SPACOMM 2013

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SPACOMM 2013 Editors

Timothy Pham, Jet Propulsion Laboratory, USA

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The Fifth International Conference on Advances in Satellite and Space Communications (SPACOMM 2013), held between April 21st-26th, 2013 in Venice, Italy, continued a series of events attempting to evaluate the state of the art in academia and industry on the satellite, radar, and antennas based communications bringing together scientists and practitioners with challenging issues, achievements, and lessons learnt.

Significant efforts have been allotted to design and deploy global navigation satellite communications systems. Satellite navigation technologies, applications, and services still experience challenges related to signal processing, security, performance, and accuracy. Theories and practices on system-in-package RF design techniques, filters, passive circuits, microwaves, frequency handling, radars, antennas, and radio communications and radio waves propagation have been implemented. Services based on their use are now available, especially those for global positioning and navigation. For example, it is critical to identify the location of targets or the direction of arrival of any signal for civilians or on-purpose applications; smarts antennas and advanced active filters are playing a crucial role. Also progress has been made for transmission strategies; multiantenna systems can be used to increase the transmission speed without need for more bandwidth or power. Special techniques and strategies have been developed and implemented in electronic warfare target location systems.

We take here the opportunity to warmly thank all the members of the SPACOMM 2013 Technical Program Committee, as well as the numerous reviewers. The creation of such a high quality conference program would not have been possible without their involvement. We also kindly thank all the authors who dedicated much of their time and efforts to contribute to SPACOMM 2013. We truly believe that, thanks to all these efforts, the final conference program consisted of top quality contributions.

Also, this event could not have been a reality without the support of many individuals, organizations, and sponsors. We are grateful to the members of the SPACOMM 2013 organizing committee for their help in handling the logistics and for their work to make this professional meeting a success.

We hope that SPACOMM 2013 was a successful international forum for the exchange of ideas and results between academia and industry and for the promotion of progress in the field of satellite and space communications.

We are convinced that the participants found the event useful and communications very open. We also hope the attendees enjoyed the charm of Venice, Italy.

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**Table of Contents**

1. Concatenated Turbo/LDPC codes for deep space communications: performance and implementation  
   Carlo Condo  
2. Capacity Saving Analysis of Lp-problem based Allocation with Packet Loss and Power Metrics in GEO-Satellite Communications Channels  
   Igor Bisio, Stefano Delucchi, Fabio Lavagetto, and Mario Marchese  
3. New Optical Communication Capabilities Using Nanosatellites  
   Francesco Sansone, Lorenzo Olivieri, and Alessandro Francesconi  
4. Toward Self-Reconfiguration of Switched Ethernet Architectures in the Next Generation of Space Launchers  
   Robert Jeremy, Georges Jean-Philippe, and Divoux Thierry  
5. Development of Wideband Dual Polarized of Microstrip Antennas for Microwave Remote Sensing  
   Akbar Eslami, Manohar Deshpande, James Lee, Henry Mishoe, Jamiru Luttamaguzi, and Ehsan Sheybani  
6. Hybrid Beamformer for TPMS Interference Suppression  
   Suk-seung Hwang, Cheol Park, and Seong-min Kim  
7. Multi-millisecond GNSS Maximum Likelihood Bit Synchronization Method  
   Gerardo Ludovico Puga, Pedro Agustin Roncagliolo, and Javier Gonzalo Garcia  
8. Turbo Decoder VLSI Architecture with Non-Recursive max* Operator for 3GPP LTE Standard  
   Ashfaq Ahmed, Maurizio Martina, and Guido Masera  
9. A Novel Range-spread Target Detection Algorithm Based on Waveform Entropy for Missile-borne Radar  
   Bo Liu, Wenge Chang, and Xiangyang Li  
10. A Brief Analysis of the Motion Compensation for FMCW SAR  
    Gaowei Jia, Wenge Chang, Xiangyang Li, and Zhiyong Zhao  
11. A Novel DDS-PLL Hybrid Structure to Generate the LFM Signal  
    Zhiyong Zhao, Xiangyang Li, Wenge Chang, and Gaowei Jia  
12. Research of Current Movements of the Earth's crust in the Territory of Almaty City  
    Zhumabek Zhantayev, Baurzhan Kurmanov, Assylkhan Bibossinov, and Kairat Sadykov  
13. Cloud top height estimation from the Meteosat water vapor imagery using computational intelligence techniques: Preliminary results  
    Stavros Kolios, Petros Karvelis, Chrysostomos Stylios, and Periklis Tagkas
The Two-Loop Tracking Strategy for Real-Time Multi-Antenna GNSS Receivers

*Pedro A. Roncagliolo, Jorge Cogo, and Javier G. García*
Concatenated Turbo/LDPC Codes for Deep Space Communications: Performance and Implementation

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Abstract—Deep space communications require error correction codes able to reach extremely low bit-error-rates, possibly with a steep waterfall region and without error floor. Several schemes have been proposed in the literature to achieve these goals. Most of them rely on the concatenation of different codes that leads to high hardware implementation complexity and poor resource sharing. This work proposes a scheme based on the concatenation of non-custom LDPC and turbo codes that achieves excellent error correction performance. Moreover, since both LDPC and turbo codes can be decoded with the BCJR algorithm, our preliminary results show that an efficient hardware architecture with high resource reuse can be designed.

Keywords—LDPC; turbo; concatenation; deep space

I. INTRODUCTION

The world of communications is characterized by a continuous strive for better performance: communication systems are usually pushed towards higher throughput, lower Bit-Error-Rate (BER) and lower power consumption with every generation. A particular application is deep space communications: due to the limited number of complete developments, their evolution in this domain is slower than in other application fields. Moreover, their requirements and constraint can differ substantially from all other communication environments. Transmission between spacecrafts and Earth are supposed to be sporadic events, but the limited amounts of available power and the long distances make failed reception, and consequent retransmission, an unacceptable event. For this reason, deep space missions do not require a high throughput, while at the same time they demand very strict BER and frame-error-rate (FER) performance. The Consultative Committee for Space Data Systems (CCSDS) suggests a set of rules (a de facto standard) for all space-related communication systems. In [1] four channel coding schemes are described, while in [2] the considered channel coding options are assembled into application-wise FEC schemes. Deep space exploration requires the use of powerful error correction codes, such as turbo codes [3], lower rate low-density-parity-check (LDPC) codes [4] and concatenated Reed-Solomon (RS) and convolutional codes. Various works have proposed deep space FEC schemes, also using codes that are different from those suggested in [1]: from custom-constructed single codes [5] to more complex concatenated schemes [6]. Concatenation between different codes has been frequently considered in order to improve performance. “Guaranteed-performance-codes” like RS or BCH codes are often used as Outer Codes (OCs) thanks to their measurable error correction capabilities, and joined to Inner Codes (ICs) such as convolutional or LDPC, used in WiMAX and DVB-S2. The same RS+convolutional FEC scheme devised in [2] allows these codes to rival with the more powerful LDPC and turbo codes. However, concatenation comes at a usually high implementation cost: decoding support for sometimes very different codes must be provided, increasing area and power consumption. Low-complexity decoders have been designed for many codes [7], [8] but steps have been recently taken towards flexibility, with area efficient multi-code decoders [9], [10].

This paper presents a deep space oriented FEC scheme by serial concatenation of LDPC and turbo codes. Section II describes how, through a particular representation of the LDPC parity check matrix, both kinds of codes can be decoded with the same algorithm, allowing very low-cost implementation of a joint decoder. The FEC scheme is described in detail in Section III. Performance of the concatenated scheme is compared with the standard requirements in Section IV, together with a set of recent works both on deep space communications and on concatenated codes. Finally, an estimation of the complexity of decoder supporting both codes is made in Section V.

II. TURBO AND LDPC DECODING ALGORITHMS

Convolutional Turbo Codes (or CTCs) are obtained as the parallel concatenation of two constituent Convolutional Code (CC) encoders. Consequently, also the decoder is made of two different parts, called Soft-In-Soft-Out (SISO) or Maximum-A-Posteriori (MAP) decoders, and connected by an interleaver II and a de-interleaver II$^\dagger$. Each MAP decoder implements the BCJR algorithm [11], which produces extrinsic metrics from a priori information. Representing the constituent CC as a trellis, let us define $k$ as a trellis step and $u$ as an uncoded symbol. Each decoder computes $\lambda_k[u] = \lambda_k^{\text{apriori}}[u] - \lambda_k^{\text{c}}[u] - \lambda_k^{\text{c}}[c^u]$ where $\lambda_k^{\text{c}}[u]$ is the a-posteriori information, $\lambda_k^{\text{apriori}}[u]$ is the a priori information and $\lambda_k[c^u]$ is the systematic component of the intrinsic information. The a-posteriori information is obtained as follows:

$$\lambda_k^{\text{apriori}}[u] = \max_{e \in \mathcal{E}(c^u) = \tilde{u}} \{b(e)\} - \max_{e \in \mathcal{E}(c^u) = \tilde{u}} \{b(e)\}$$

(1)

where $\tilde{u} \in \mathcal{U}$ is an uncoded symbol taken as a reference (usually $\tilde{u} = 0$) and $u \in \mathcal{U} \setminus \{\tilde{u}\}$ with $\mathcal{U}$ the set of uncoded
symbols; \( e \) is a trellis transition and \( u(e) \) is the corresponding uncoded symbol. Following the Max-Log-MAP approximation [12], for a small BER degradation the \( \max \{x_i\} \) function becomes \( \max \{x_i\} \). The term \( b(e) \) in (1) can consequently be defined as:

\[
b(e) = \alpha_{k-1}[s^E(e)] + \gamma_k[e] + \beta_k[s^E(e)]
\]  

(2)

\[
\alpha_k[s] = \max_{e \in s^E(e) = s} \{\alpha_{k-1}[s^S(e)] + \gamma_k[e]\}
\]  

(3)

\[
\beta_k[s] = \max_{e \in s^S(e) = s} \{\beta_{k-1}[s^E(e)] + \gamma_k[e]\}
\]  

(4)

\[
\gamma_k[e] = \lambda_k^{apr}[u(e)] + \lambda_k[e]
\]  

(5)

where \( s^S(e) \) and \( s^E(e) \) are the starting and the ending states of \( e \). \( \alpha_k[s^S(e)] \) and \( \beta_k{s^E(e)} \) are the forward and backward metrics associated to \( s^S(e) \) and \( s^E(e) \) respectively, while \( \gamma_k[e] \) is the channel intrinsic information.

LDPC codes are identified by a spare parity check matrix \( H \), of size \( M \times N \). A received codeword must satisfy all the parity checks (rows of \( H \)), i.e., \( H \cdot x^T = 0 \), where \( x \) is the codeword of length \( N \). Different decoding approaches are possible, depending on the graph representation of \( H \): the classical approach defines a bipartite graph with \( N \) Variable Nodes (VNs) and \( M \) Check Nodes (CNs), and edges between VN \( i \) and CN \( j \) if a nonzero entry is present in column \( i \) and row \( j \) of \( H \). Layered decoding [13] on the contrary, sees \( H \) as a multipartite graph composed of different layers of parity checks: this scheduling allows to exploit faster convergence thanks to multiple updates of the bit error probabilities within a single iteration.

Calling \( \lambda[c] \) the Logarithmic Likelihood Ratio (LLR) of symbol \( c \) and, for column \( k \) in \( H \), bit LLR \( \lambda_k[c] \) is initialized to the corresponding received soft value. These are the VN-to-CN messages. The following operations are executed for all parity constraints \( l \) in a given layer, and reiterated up to the desired level of reliability:

\[
\Delta \lambda_{lk}[c] = \lambda_k[c] - \Lambda^{apr}_{lk}
\]  

(6)

\[
\lambda_k[c] = \Delta \lambda_{lk}[c] + \Lambda^{po}_{lk}
\]  

(7)

where \( \Lambda^{po}_{lk} \) is the CN-to-VN message, namely the updated version of \( \Lambda^{apr}_{lk} \), that is initialized to 0, and stored for the next iteration. Several exact and approximated algorithms have been proposed to calculate \( \Lambda^{po}_{lk} \): the most common algorithm used in LDPC decoding is the Belief Propagation (BP) algorithm, of which one of the most used approximations is the min-sum and its variations [14].

It can be clearly seen that LDPC and turbo decoding processes share many characteristics. Both of them are iterative, rely on soft information, are usually implemented in their logarithmic form, while commonly being represented through special kinds of graphs. A particularly interesting exploitation of these characteristics has been proposed in [15]. Every row of \( H \) is seen as a turbo code with trellis length equal to the row weight: a direct link between turbo and LDPC codes is drawn, and turbo decoding algorithms can be applied to LDPC codes with minor adjustments. The BCJR-based LDPC decoding relies on the fact that binary LDPC codes have a 2-state trellis: state metrics can consequently be expressed as differences \( \Delta\alpha[c] \) and \( \Delta\beta[c] \), reducing the quantization noise. Defining \( \Phi(x,y) = \max(x,y) - \max(x+y,0) \) and considering the Max-Log-MAP approximation [16], the CN-to-VN message update becomes:

\[
\Lambda^{po}_{lk} = \Phi(\Delta\alpha_k[c], \Delta\beta_k[c])
\]  

(8)

\[
\Delta\alpha_k = \Phi(\Delta\alpha_k, \Delta\beta_k[c])
\]  

(9)

\[
\Delta\beta_k = \Phi(\Delta\beta_k[c], \Delta\lambda_k[c])
\]  

(10)

where \( \Delta\alpha[c] \) and \( \Delta\beta[c] \) at the edge of the trellis are initialized as the minimum value of the dynamic range.

### III. PROPOSED FEC SCHEME

In the CCSDS recommended standard [17] transmission data rates of up to 2.048 Mb/s are foreseen for the next missions, while FEC schemes must be able to guarantee bit error floors not higher than \( 10^{-9} \). The need for effective coding schemes, alongside simple decoding algorithms, makes code concatenation one of the smartest solutions. The powerful Turbo and LDPC codes have been considered for concatenation before ([18]) but the low level of details provided and the unsatisfying results leave room for further investigation.

The devised FEC scheme is shown in Fig. 1: an LDPC code is serially concatenated to a turbo code. The outer encoder encodes the input bits, and the resulting codeword is used as input for the inner encoder. Being responsible of the first, rough decoding, the IC should work well also in presence of a large number of errors. Since turbo codes have better performance than LDPC at low SNR [19], they have been chosen as IC. On the contrary, receiving the updated information from the inner decoder, the OC decoder, that is an LDPC code, can fully exploit its deep waterfall region and low error floor. An interleaver scrambles the output of the OC encoder before the second encoding, while the inverse function is inserted between the decoders. Since, depending on the rate and characteristics of the chosen codes, the block sizes may not be compatible, an optional padding block has been placed after the interleaver: zeros are added to the scrambled codeword to fit the required number of bits. This reduces the coding efficiency but allows for flexible concatenation. The padding bits are removed after the IC decoding, before the deinterleaver.

The decoding process starts in the IC decoder, which performs up to \( \text{Iter}_{in} \) iterations. After \( \text{Iter}_{in} \) the codeword is stripped of the padding bits, descrambled and passed to the OC decoder. Both turbo and LDPC code decoding algorithms involve soft information: while the IC decoder receives measures of bit error probabilities from the channel estimator, the OC decoder must rely on the metrics updated by the IC decoder. In particular, the LDPC decoder receives as intrinsic information for the initialization of \( \lambda_k[u] \) the bit-level extrinsic output of the turbo SISO \( \lambda_k[u] \). These metrics are passed through the deinterleaver along with the codeword.
The CCSDS suggests three FEC schemes for space communications in [1]: a RS-convolutional codes concatenated scheme, turbo codes and LDPC codes. Deep space communications requiring very low bit error rates address turbo codes in particular, allowing four code rates ranging from 1/6 to 1/2, and four information block lengths in the range 1784-8920. The WiMAX standard [19] relies on a wide set of Quasi-Cyclic LDPC codes (QC-LDPC), together with turbo codes of different length and rates. The two code types are mutually exclusive options in the standard: their proven effectiveness and implementation-friendly structure, however, make them ideal candidates for concatenation towards deep-space applications, regardless of the relatively low performance of WiMAX LDPC codes w.r.t. CCSDS LDPC codes. Thanks to the wide variety of available codes, it has been possible to experiment with different code combinations consisting of both WiMAX and CCSDS codes. Though CCSDS Single-Binary (SB) turbo codes employ 16 states, it is proven in Section IV that also the Duo-Binary (DB), eight-state codes used in WiMAX guarantee very good results while keeping the decoding complexity low.

IV. SIMULATIONS AND PERFORMANCE COMPARISONS

To evaluate the effectiveness of the proposed approach, simulations have been run on a proprietary tool. Its deeply customizable structure allows to select codes, channel model, SNR and result reliability level, together with decoding algorithms and related choices (number of iterations, stopping criteria). Moreover, it is possible to tweak a set of implementation-oriented characteristics, like different approximations of the chosen algorithms and number of bits assigned to the representation of the metrics.

In order to comply as much as possible with the requirements of CCSDS, the turbo codes suggested in [1] have been used as ICs in a first batch of simulations, and concatenated with WiMAX LDPC codes. The relatively high rate of the OC results in a concatenated rate that is very close to the CCSDS specifications. Block size compliance is guaranteed by using, if needed, multiple LDPC codewords as a single turbo information block, together with padding bits. The maximum allowed number of iterations has been set to 10 for both IC and OC decoder, and following most of the state of the art on deep space communications, the AWGN channel model has been chosen.

Moving towards a hardware implementation of the proposed FEC scheme, some limitations have been inserted in the simulation environment. To correctly evaluate the impact of soft information quantization on the BER and FER, the dynamic range of all metrics involved in the decoding process has been limited to 10 or 9 bits, with 3 bits of fractional part. For the same reason, both the LDPC and turbo codes have been decoded with the BCJR algorithm, thus leading to low decoding complexity.

Early experimentations have shown that the gain that can be obtained with the insertion of a bit interleaver between the two encoders is negligible w. r. t. the additional complexity, and it has not been considered in the plotted curve. This limited effect is mainly due to the sparse structure of the H matrix, that acts as an interleaver by itself [18].

Fig. 2 and 3 plot a set of meaningful BER and FER curves respectively. The “+” marker indicates the curves provided by CCSDS in [1] for SB turbo codes of rate 1/3 (continuous) and 1/4 (dashed). They are obtained with 10 decoder iterations, QPSK modulation and AWGN channel. The ×-marked curves show the performance of these codes when concatenated with a WiMAX rate 5/6 LDPC code. It can be seen that both
concatenated BER and FER follow very closely the standard’s curves: FER results are particularly encouraging, thanks to aggregated error distributions that are addressed later in this section. Moreover, plots in [1] show error floors as early as BER=$10^{-6}$ (block length 8920): the minimum BER simulated with the concatenated scheme is slightly above $10^{-10}$, with no signs of error floor.

The continuous, ■-marked curves have been obtained by substituting to the SB turbo code of CCSDS in the concatenated scheme a WiMAX DB turbo code of comparable size and same rate. The difference between the two curves is negligible, while the complexity of an 8-state turbo decoder is much lower than a 16-state one, as shown in Section V.

To evaluate the influence of IC and OC respective rates on the decoding performance, a second set of simulations has been run: the IC rate has been fixed to 1/3 and by changing the OC rate it has been possible to obtain concatenated rates equal to those of CCSDS turbo codes. In both Fig. 2 and 3, the dashed, ■-marked curve has a concatenated rate of 1/4, obtained by CTC 1/3 + LDPC 3/4. It can be noticed how the lower rate of the OC fails to deliver the same BER and FER results as CTC 1/4 + LDPC 5/6: this behavior, observed also in the CTC 1/6 + LDPC 5/6 against CTC 1/3 + LDPC 1/2 case, reveals how the IC turbo rate is more critical than for the LDPC OC.

Fig. 4 plots a set of BER curves to compare the performance of WiMAX concatenated codes against single LDPC and turbo codes. The curves showing the “+” marker refer to the concatenated FEC scheme: both use a WiMAX turbo code of rate 1/3 and 960 two-bit input symbols. The continuous line has been obtained using a rate 1/2, codeword length 1920 LDPC, while the dashed one with a rate 5/6 LDPC. The higher rate LDPC results in a less steep curve: this degradation can be addressed by rising Iter_{\text{in}}, with the two curves superimposed at Iter_{\text{in}}=20. These plots are compared to the constituent rate 1/3 CTC and rate 1/2 LDPC with different numbers of allowed iterations. Since the concatenated scheme has at its disposal up to 20 iterations (10 CTC + 10 LDPC), single-code curves are plotted with both 20 and 10 iterations maximum. The concatenated BER shows very good performance at low $E_b/N_0$, crossing the $10^{-6}$ threshold at $E_b/N_0=0.9$ dB when using the rate 5/6 LDPC OC. At higher $E_b/N_0$, its performances are even more remarkable: a total of $2 \cdot 10^8$ frames have been simulated for all the shown $E_b/N_0$ points, for a total of $1.92 \cdot 10^{11}$ information bits, and no errors were counted for $E_b/N_0$ higher than 1.0 dB when using the rate 1/2 LDPC code. The curves show no sign of error floor and constant BER decrease. At low $E_b/N_0$, the 20-iterations single CTC outperforms the concatenated schemes of up to 0.15 dB: the difference is smaller than that reported in [18] with much more favorable conditions, and the crossing point occurs at a higher BER and much lower $E_b/N_0$.

Fig. 5 shows the FER curves for the same parameters as Fig. 4: the concatenated FEC reaches very low values ($7 \cdot 10^{-8}$ with rate 5/6 LDPC code). The difference with the 20-iteration single CTC curve at low $E_b/N_0$ is substantially reduced (0.08 dB maximum) w. r. t. the BER, and the crossing point is moved at lower $E_b/N_0$. This is due to the fact that very often a failed decoding with the concatenated scheme is due to a high number of wrong bits within the same frame. Consequently, BER and FER scale differently, since errors are clustered together and affect a very low number of frames.

Table I provides a comparison of the proposed FEC scheme with similar state of the art solutions. Solution A refers to CCSDS SB turbo 1/4 + WiMAX LDPC 5/6, while solution B to CCSDS SB turbo 1/3 + WiMAX LPDC 5/6, both already shown in Fig. 2 and Fig.3. To help a fair comparison with coding schemes with different rate, the $\Delta_{\text{SHN}}$ row identifies the distance of the BER curve from the Shannon limit at BER=$10^{-6}$: in both cases the distance is less than 1.5 dB. The obtained $\Delta_{\text{SHN}}$ is similar to that of the AR4JA LDPC codes proposed by CCSDS [1], but has been obtained with a much smaller number of iterations.

In a recent work [20], a FEC scheme for 3D HDTV using an outer block turbo code (BTC) concatenated to an LDPC code is proposed. The scheme is shown to outperform the DVB-
TABLE I. PERFORMANCE COMPARISON AMONG FEC SCHEMES

<table>
<thead>
<tr>
<th>Application</th>
<th>A</th>
<th>B</th>
<th>[20]</th>
<th>[21]</th>
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<th>[6]</th>
<th>[22]</th>
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<tr>
<td>Inner Code</td>
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<tr>
<td>Rate, in n</td>
<td>1/4</td>
<td>1/3</td>
<td>2/3</td>
<td>3/5</td>
<td>1/3</td>
<td>1/3</td>
<td>1/3</td>
<td>1/3</td>
<td>1/3</td>
</tr>
<tr>
<td>Rate, out</td>
<td>5/24</td>
<td>5/18</td>
<td>467/500</td>
<td>9/10</td>
<td>467/1000</td>
<td>3/5</td>
<td>25/49</td>
<td>49/50</td>
<td>1/2</td>
</tr>
<tr>
<td>Input, in</td>
<td>3568</td>
<td>16 K</td>
<td>1000 symbols</td>
<td>2379</td>
<td>16 K</td>
<td>504</td>
<td>504</td>
<td>1792</td>
<td></td>
</tr>
<tr>
<td>Input, out</td>
<td>1920</td>
<td>15 K</td>
<td>900 symbols</td>
<td>2379</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Outer Code</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Rate, in n</td>
<td>10</td>
<td>N/A</td>
<td>N/A</td>
<td>N/A</td>
<td>N/A</td>
<td>N/A</td>
<td>N/A</td>
<td>N/A</td>
<td>N/A</td>
</tr>
<tr>
<td>Rate, out</td>
<td>50</td>
<td>20</td>
<td>15</td>
<td>N/A</td>
<td>N/A</td>
<td>N/A</td>
<td>N/A</td>
<td>N/A</td>
<td>N/A</td>
</tr>
<tr>
<td>Quantization</td>
<td>10-9 bits</td>
<td>Floating Point</td>
<td>Floating Point</td>
<td>Floating Point</td>
<td>Adaptive</td>
<td>Floating Point</td>
<td>Floating Point</td>
<td>Floating Point</td>
<td>Floating Point</td>
</tr>
<tr>
<td>Channel</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>E_b/N_0 @ BER=10^-6</td>
<td>0.43</td>
<td>0.65</td>
<td>4.4</td>
<td>1.85</td>
<td>N/A</td>
<td>N/A</td>
<td>N/A</td>
<td>2.35</td>
<td>1.75</td>
</tr>
<tr>
<td>FER @ BER=10^-6</td>
<td>8 x 10^-6</td>
<td>8 x 10^-6</td>
<td>4 x 10^-4</td>
<td>N/A</td>
<td>1.55</td>
<td>N/A</td>
<td>N/A</td>
<td>2.0 x 10^-5</td>
<td>1.55</td>
</tr>
<tr>
<td>Δ_SHN @ BER=10^-6</td>
<td>1.38</td>
<td>1.43</td>
<td>4.50</td>
<td>1.55</td>
<td>N/A</td>
<td>N/A</td>
<td>N/A</td>
<td>2.90</td>
<td>2.43</td>
</tr>
<tr>
<td>minimum BER</td>
<td>2 x 10^-9</td>
<td>10^-10</td>
<td>2 x 10^-7</td>
<td>8 x 10^-7</td>
<td>3 x 10^-10</td>
<td>N/A</td>
<td>N/A</td>
<td>1.2 x 10^-7</td>
<td>4 x 10^-7</td>
</tr>
<tr>
<td>minimum FER</td>
<td>6 x 10^-8</td>
<td>1.1 x 10^-8</td>
<td>8 x 10^-5</td>
<td>N/A</td>
<td>10^-8</td>
<td>N/A</td>
<td>N/A</td>
<td>6 x 10^-6</td>
<td>1.9</td>
</tr>
<tr>
<td>E_b/N_0 @ min</td>
<td>0.6</td>
<td>0.9</td>
<td>4.6</td>
<td>2.0</td>
<td>0.3</td>
<td>1.05</td>
<td>2.5</td>
<td>1.9</td>
<td>1.6</td>
</tr>
</tbody>
</table>

The two closely related works [18] and [23] implement CTCs as inner and LDPC codes as OCs, decoding them with Log-MAP and probability-domain BP algorithms respectively. The work in [23] presents the same system as [18], with the addition of a certain number of global decoder iterations that slightly improve the BER. The plotted curves show a degradation of the concatenated scheme (5+50 iterations) w. r. t. the single turbo code (5 iterations) of up to 0.3 dB that has not been observed in this work (Fig. 4). Rates are comparable to that of B, and they both use the AWGN channel model, while the block size of B is larger. Regardless of the much higher precision and number of iterations allowed in [18], A and B yield better BER and Δ_SHN results, with a gain ranging from 0.90 to 1.10 dB at BER=10^-6.

V. IMPLEMENTATION COMPLEXITY

The proposed FEC scheme joins the BER of turbo codes at low E_b/N_0 with that of LDPC codes at higher E_b/N_0, resulting in very steep performance curves that are well suited for deep space communications. However, concatenation is usually expensive when it comes to implementation, since decoding is required for both codes separately. Resource sharing can be attempted, but if the decoding algorithms are very different the obtained area saving are often overweighted by the additional logic required. Since LDPC and turbo codes can be decoded with the BCJR algorithm, it is possible to share a large part of the datapath. Currently, the authors...
are working on a flexible and efficient decoder meeting the requirements for concatenated LDPC and turbo decoding: the decoder must be able to switch between LDPC and turbo configurations on-the-fly, in case of IC failed decoding. Partial synthesis estimations show that around 95% of the LDPC datapath can be shared with the turbo datapath. Moreover, due to the higher number of states in the trellis of turbo codes, the LDPC decoder architecture can exploit an internal level of parallelism (up to 8× in presence of 16-state turbo codes). This aspect is particularly useful in case many LDPC codewords are used as a single information block for turbo codes. Taking as an example the multi-core LDPC-turbo decoder presented in [9], a 95% shared datapath would result in a 14% area reduction in every processing element. The limited throughput requirements set by CCSDS allow to scale down the operating frequency: in B, a throughput of 2 Mb/s would be achieved at 8 MHz, with a huge decrease in power consumption.

BP and CHASE algorithms used in [20] are very different and, to the best of our knowledge, datapath sharing has never been considered. However, the adaptive BP proposed in [25] for BTC decoding is likely to allow at least partial resource sharing. Also in case of LT and NB-LDPC codes used in [21] and [6], while the datapaths are different, both decoding processes rely on a bipartite graph: this can lead to some degree of sharing of the internal connections and memories.

In [22] LDPC decoding is performed through LLR-BP: the min-sum approximation only requires sums and minimum extractions, making its implementation very simple. In [26] resource sharing between min-sum and SISO-based turbo decoding is considered, resulting in unsatisfying improvements: the overhead would be even greater with the probability-domain BP used in [18].

VI. CONCLUSION AND FUTURE WORK

In this paper, a FEC scheme relying on serial concatenation of parallel CTCs and LDPC codes is presented. Detailed simulations and comparison with the state of the art show competitive results, outperforming or being comparable to the latest concatenation and deep space schemes. The use of the same algorithm to decode both codes leads to a small area, low power implementation that is currently under development. Early complexity estimations show that it is possible to implement the proposed approach at a very small cost w. r. t. even single code deep space communications FEC schemes.

REFERENCES


Capacity Saving Analysis of $L_p$-problem based Allocation with Packet Loss and Power Metrics in GEO-Satellite Communications Channels

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Abstract—This paper introduces a criterion, based on the $L_p$-problem, which has been employed to allocate capacity among Earth Stations. The obtained allocations permit a performance compromise between Packet Loss and Transmitted Power, which are taken into account as performance metrics. Moreover, starting from the proposed $L_p$-problem based allocation and considering specific analytical models for the Packet Loss Probability (PLP) and the Transmitted Power (TP), the paper highlights the existence of a Capacity Bound, independent of the overall capacity available $C_{TOT}$, on which the allocations converge. The main contribution of the paper concerns the performance analysis, carried out by simulation, which shows that the proposed method enables a significant savings of capacity and Transmitted Power and simultaneously, with only a limited worsening of the Packet Loss.

Keywords-Satellite Communications; Multi-Objective Programming; $L_p$-problem based Allocation; Capacity Bound; Performance Analysis;

I. INTRODUCTION

Resource allocation in modern satellite networks, such as satellite-sensor [1], LTE [2] and WiMax [3], plays a crucial role and needs to be deeply investigated. In several previous works (such as [4], [5], [6] and [7]), we consider a scenario composed of a satellite communication system of $Z$ earth stations that receive TCP traffic flows from a fixed number of sources and forward them onto a common geostationary satellite channel with an overall available capacity set equal to $C_{TOT}$ [bps]. The channel state is modeled by considering the fading effect, due to atmospheric conditions, that have a negative impact on the quality of communications. As a consequence, to compensate, we apply a Forward Error Correction (FEC) encoding system, setting the code rate as a function of the fading level $F$, expressed in [dB], undergone by each station. From the viewpoint of the higher protocol layers (i.e., above the network layer) the redundancy bits added to protect the transmitted information cause a reduction of the available capacity for transmission. Taking into account the proposed scenario, the problem considered in this paper is the well known capacity allocation problem. The goal is to share the channel capacity among the earth stations according to the policy defined in [6] whose performance is further analysed and discussed in this paper. The proposed allocation problem models each transmission entity (i.e., the Earth Stations) using some functions, which values are directly proportional to the capacity allocated, that represent some metrics that need to be optimized simultaneously (e.g., the Packet Loss Probability, shortly PLP, and the Transmitted Power, shortly TP, as done in this work). If the functions are in contrast to each other, as it happens in the case of this paper, the solution of the allocation problem must represent a compromise. For this reason we use the Multi-Objective Programming (MOP) framework to formulate the allocation problem thereby obtaining a solution also known as the Pareto Optimal Point (POP) set. To find out a single solution from this set, which is representative of the best compromise between the adopted metrics, we apply the $L_p$-problem.

An important consequence of the proposed approach is that its solution converges if $C_{TOT}$ increases (i.e., the Capacity Bound (CB) discussed in [6]). It enables a significant capacity savings with respect to the allocation of the overall resources available on the channel: $C_{TOT}$. In fact, sharing all the channel capacity enables to optimize the value of a decreasing metric, however this is not true for an increasing metric. So it is useless to allocate the whole capacity but a significant portion can be reserved to increase the number of earth stations that can transmit on the satellite channel. It is worth noticing that all the stations, that are allowed to transmit, experience the same conditions in terms of PLP and TP.

The aim of this paper is to present a performance analysis of the allocation problem proposed, by focusing on the evaluation of the benefits obtained, considering the value of the PLP and the TP, with respect to the allocation of the whole available capacity $C_{TOT}$. Another key point of this paper is the evaluation of the increment of the number of earth stations computed by using a formula of the average number of earth stations, which can transmit with fixed value of PLP and TP, is presented.

As next step of the research presented in this work, adaptive modulation and coding techniques, applied together with the resource allocation, for example based on the DVB-S2 standard as in [8] and [9], will be object of thorough study. The rest of the paper is organized as follows. The next section presents a brief survey of the state of the art of resource allocation for satellite and wireless communications systems. In Section II the model adopted for the capacity
allocation is presented while Section III describes the formulae employed for the Packet Loss Probability (PLP) and Transmitted Power (TP). Section IV presents the simulation results that highlight the advantages obtained allocating a capacity equal to the CB, obtained by using our allocation method with respect to the allocation of the overall available capacity. Finally Conclusions are drawn.

II. THE FORMULATION ALLOCATION PROBLEM

The model adopted in this paper and developed in previous works by same authors ( [5] and [6]), is based on three different components. The physical entities represent the earth stations, which are composed by different queues called virtual entities. Each queue is characterized by several objective function that model some performance metrics such as the Packet Loss Probability and the Transmitted Power, as done in this work.

A. The MOP based capacity allocation problem

To formalize the MOP based capacity allocation we define two vectors: the first one, reported in (1) is the control vector and contains all the capacities allocated to each virtual entity for each physical entity considered. Formally we consider that each physical entity is identified by \( z \in [1, Z] \). \( Y_z \) is the number of virtual entities of the \( z \)-th physical entity. Each virtual entity is identified by \( y_z \in [1, Y_z] \). \( M_{yz} \) is the number of objective functions for each virtual entity \( y_z \). Each objective function, of a given \( y_z - th \) virtual entity, is identified by the index \( m \in [1, M_{yz}] \). \( C_{yz} \) is the capacity allocated to the virtual entity \( y \) of the physical entity \( z \)

\[
C = (C_{11}, C_{21}, C_{31}, ..., C_{1Y_1}, ..., C_{1Z}, C_{2Z}, C_{3Z}, ..., C_{Y_Z})
\]

(1)

In this work the capacity allocation is supposed to be carried out in a centralized fashion: the control vector is the output of a centralized decision maker and represents the decision taken by the allocation algorithm, that is the capacity assigned to each queue in each earth station. It is worth noticing that the overall capacity allocated to the \( z - th \) physical entity, \( C_z \) is the sum of the capacities allocated to all its virtual entities that are

\[
C_z = \sum_{y=1}^{Y_z} C_{yz}
\]

The second vector defined is the so called objective function vector and contains the value of each objective function of each virtual entity in each physical entity. All the components of this vector are the metrics, the PLP and the TP analytically defined in Section III that need to be optimized simultaneously.

\[
F(C) = (F_{C_{11}}(C), ..., F_{M_{11}}(C), ..., F_{C_{1Y_1}}(C), ..., F_{M_{1Y_1}}(C), \ldots, F_{C_{1Z}}(C), ..., F_{M_{YZ}}(C), Y_Z(C))
\]

(2)

In (2), the generic term \( F_{m,y_z}(C) \) is the \( m - th \) objective function of the \( y - th \) virtual entity of the \( z - th \) physical entity. Considering the Multi Objective Programming theory formalized in [5] with the control vector defined in (1) and the objective function vector (2) is possible to formalize the MOP based capacity allocation problem as follows:

\[
\begin{align*}
\begin{cases}
C_{opt} = (C_{11,opt}, C_{21,opt}, ..., C_{Y_1,opt}, ..., C_{1Z,opt}, C_{2Z,opt}, ..., C_{Y_Z,opt}) = \arg \min_{C} F(C) ; \\
C_{yz} \geq 0, \forall y_z \in [1, Y_z], \forall z \in [1, Z] \\
\sum_{z=1}^{Z} \sum_{y=1}^{Y_z} C_{yz} \leq C_{TOT}
\end{cases}
\end{align*}
\]

(3)

An important boundary is imposed in (3): the sum of the capacity allocated to each virtual entity must be lower or equal to the overall capacity available in the channel \( C_{TOT} \). As a consequence the MOP capacity allocation problem determines a feasibility region that contains all the control vector that may be admissible solutions. The goal of this allocation policy is to determine a control vector \( C_{opt} \) that simultaneously minimizes all the components of the objective function vector.

The solution of the defined problem is not single but it is a set, called Pareto Optimal Points (POP) set. The position of this set depends on the characteristics of the objective functions considered: if all objective functions are strongly decreasing [10], then a solution \( C_{opt} \) is a POP if and only if the solution is on the constraint boundary

\[
\sum_{z=1}^{Z} \sum_{y=1}^{Y_z} C_{yz} = C_{TOT}.
\]

If this condition is not true the POP set may also stay also inside the feasibility region which means that the Pareto Optimal may also be a point for which

\[
\sum_{z=1}^{Z} \sum_{y=1}^{Y_z} C_{yz} < C_{TOT}.
\]

The strongly decreasing assumption concerning the objective-function vector is quite typical because common performance functions applied in telecommunication networks such as Packet Loss Probability, Packet Delay and Packet Jitter are quantities that decrease their values when the allocated capacity value increases. This is not true if other important metrics are also used: power, but also processing and computation effort. In those cases, as done in this paper considering the PLP as one of the adopted metrics, the allocation of the overall available capacity \( C_{TOT} \) may not be the optimal choice. In fact as previously said the optimal solution may also be also inside the feasibility region and not on the constraint boundary. The aim of this paper is to identify a solution through the MOP approach, as well as to compare the performance by way of the results obtained sharing the overall capacity available among the considered earth stations.

B. The \( L_p \)-problem based capacity allocation

As previously said, the MOP based capacity allocation problem determines as a solution a set of points. In this
paper we apply the $L_p$-problem to obtain a single control vector that represents a compromise between the adopted metrics. The idea is to allocate capacity so that the value of each objective function is as close as possible to its ideal value. The set of ideal capacities (i.e. the ideal vector (4)) is composed of the ideal decision variable vector elements $C^{F_k,y_z}_{y_z,id}$ for which $F_{k,y_z}$ attains the optimum value

$$
C^{F_k,y_z}_{id} = (C^{F_k,y_z}_{1,1,id}, C^{F_k,y_z}_{2,1,id}, ..., C^{F_k,y_z}_{1',id}, ..., C^{F_k,y_z}_{1,2,id}, C^{F_k,y_z}_{2,2,id}, ..., C^{F_k,y_z}_{Y_z,1',id}, ..., C^{F_k,y_z}_{Y_z,2,id}) = \arg\min_{C} F_{k,y_z}(C);
$$  

(4)

$\forall k \in [1, M_{y_z}], \forall y_z \in [1, Y_z], \forall z \in [1, Z]$  

Each element $C^{F_k,y_z}_{y_z,id}$ can assume a value between 0 and $C_{TOT}$, independently of any physical constraint and of the values of the other components of vector (4). It is called ideal (utopian) for this reason. For example, if a generic objective function is decreasing versus the allocated capacity, it is obvious that it is ideal to use all the possible capacity $C_{TOT}$, while if it is increasing versus capacity, it is ideal allocating no capacity at all. The values of vector (4) are considered known in the remainder of the paper while the vector in (5) contains each objective function attaining its ideal value.

$$
F_{id} = (F_{1,1,1,id}(C^{F_{1,1,1}}_{id}), ..., F_{k,y_z,1,id}(C^{F_{k,y_z,1}}_{id}) ..., F_{M_{Y_z},Y_z,1}(C^{F_{M_{Y_z},Y_z,1}}_{id}))
$$

(5)

To compute the distance to the ideal vector we apply the generic norm $p$ and in (6) and in (7) is reported the formulation of the $L_p$-problem based capacity allocation ($L_p$CA)

$$
C_{all} = (C_{1,all}, C_{2,all}, ..., C_{Y_z,all}, ..., C_{1,all}, C_{2,all}, ..., C_{Y_z,all}) = \arg\min_{C \subseteq C_{opt}} J_p(C)
$$

(6)

where $J_p(C)$ is a function representing the generic norm, usually indicated with the symbol $L_p$ [10]:

$$
J_p(C) = \left( \sum_{z=1}^{Z} \sum_{y=1}^{Y_z} \sum_{k=1}^{M_{y_z}} \left| w_{k,y_z} \right| F_{k,y_z}(C^{F_{k,y_z}}_{id}) \right)^{1/p} + \left| F_{k,y_z,1,id}(C^{F_{k,y_z,1}}_{id}) \right|^{1/p}
$$

(7)

and $\sum_{k=1}^{M_{y_z}} w_{k,y_z} = 1, w_{k,y_z} > 0, \forall k \in [1, M_{y_z}], \forall y_z \in [1, Y_z], \forall z \in [1, Z]$ so to assure the Pareto optimality of the solution. Modifying the values of the weights it is possible to differentiate the importance of the considered objective functions. As a consequence, the proposed capacity allocation is elastic and can be applied to different scenarios with heterogeneous traffic flow with different QoS requirements (e.g., telephony, video-conferencing, audio/video streaming, web transactions).

The most interesting contribution of this work, that is a consequence of the allocation method proposed, concerns a comprehensive analysis of the performance assured by the existence of a Capacity Bound, that has been demonstrated in [6]. This point represents a Pareto Optimal Point (POP) on which the $L_p$CA problem converges. It implies that, from the practical viewpoint, a Satellite Service Provider may provide capacity allocations to all the entities involved in the allocation process without employing the overall available capacity. It allows dedicating the rest of the capacity to further possible entities. As shown in Section IV, moreover, it can be done without penalizing the overall performance and without capacity wasting.

III. THE OBJECTIVE FUNCTIONS

In this paper each physical entity represents an Earth Station that transmits through a satellite channel and is composed by a single queue (as a consequence, physical and virtual entities are not differentiated). Each considered entity is represented by two objective functions that are the Packet Loss Probability, shortly PLP, due to congestion ($F_{1,1,z} = P_{loss,z}(C_z)$) and the Transmitted Power, shortly TP, ($F_{2,1,z} = W_{tx,z}(C_z)$) and the constrain is defined by the amount of available capacity ($\sum_{z=1}^{Z} \sum_{y=1}^{Y_z} C_{y_z} \leq C_{TOT}$).

A. Packet Loss Probability Function

The PLP model used in this paper is referred to the loss due to congestion for a Transmission Control Protocol (TCP) based traffic flow and is analytically reported in (8):

$$
P_{loss,z}(C_z) = \frac{k_z \cdot N_z^2}{(R_z \cdot C_z \cdot rtt_z + Q_z)^2}
$$

(8)

It is worth noticing that the PLP model adopted in this work is a decreasing function with respect to the allocated capacity $C_z$. Moreover in this paper, the values of the parameters reported in (8), applied in the performance analysis section, and the related meanings are: $k_z$=128/81 is a constant depending on TCP protocol, $N_z$=10 is the number of active TCP connection for the $z$-th station, $Q_z$ is the buffer size, equal to 10 packets, for the $z$-th station, $rtt_z$ is the round trip time, equal to 512 [ms], $l$ =1500 [byte] is the TCP packet size and $R_z$ and $C_z$ are the code rate and the capacity allocated to the $z$-th station, respectively. Channel conditions vary over the time and, in this paper, the undergone fading level $F_z$ for each station represents the satellite channel status. Each Earth Station is supposed to apply different code rates in dependence on the channel status modeled considering the fading level. Code rates are assigned as in Table I. This hypothesis allows considering packet losses due to congestion because channel errors are made negligible by
applying encoding. On the other hand, as previously said, from the application layer viewpoint, represents a reduction in the available capacity for the considered station. So the detriment of the channel conditions due to the fading effect is modeled by a reduction of the capacity available to a station with respect to the portion allocated as reported in equation (8) where the applied code rate is the capacity reduction factor.

### B. Transmitted Power Function

The TP of the \( z \)-th station is reported in (9):

\[
TP_z(h_z, C_z) = (\frac{G_t}{h_z} - 1) \cdot \frac{1}{h_z} \tag{9}
\]

Considering the formulation of the TP is possible to view that it is an increasing function with respect to the allocated capacity \( C_z \). \( h_z \), called link constant in this paper, takes into account the parameters related to the satellite link. In more detail, it contains the transmission antenna gain \( G_T \), of the \( z \)-th station, the receiver antenna gain, on the satellite \( G_R \) (common for each station both equal to \( 10^4 \)), the Boltzmann constant \( k \) equal to \( 1.38 \cdot 10^{-23} \text{J} \text{K}^{-1} \), the noise temperature \( T \) set to 290 [K] (considering additive white Gaussian noise), the bandwidth of the satellite channel \( W=1 \) [MHz] and the Free Space Loss (FSL) set equal to \( 10^{19} \) as defined in [11]. In practice, the coefficient \( \frac{1}{h_z} \) is:

\[
\frac{1}{h_z} = \frac{k \cdot T \cdot W \cdot FSL}{G_T \cdot G_R} \tag{10}
\]

The Transmitted Power function is obtained by combining two equations: \( C_z = B \cdot \log_2 \left( 1 + \left( \frac{G_t}{T} \right) \right) \) the Hartley-Shannon law, and \( \left( \frac{G}{T} \right) = \frac{G_t \cdot G_R \cdot TP}{h_T \cdot B \cdot FSL} \) that represents the carrier to noise ratio [11].

It is worth noticing that the employment of equation (9) as transmitted power function implies the hypothesis that the transmission bit rate of the \( z \)-th station is equal to the allocated capacity \( C_z \). Both the objective functions considered in this paper are continuous and differentiable in \( \mathbb{R} \) so assuring the existence of a solution of the \( L_p \)-problem applied.

### C. The Capacity Bound

In this paper we consider \( Z \) physical entities, a single virtual entity for each physical entity \( (Y_z = 1 \) \((z \in [1, Z])\) and two objective functions for each virtual entity \( (k = 2 \forall z \in [1, Y_z], \forall z \in [1, Z]) \). Considering the two objective functions previously introduced, the vector \( F(C) \), defined in (2) it is possible to define a capacity bound where our allocation converges. In practice, given fixed channel conditions, if the overall capacity available for the entire communications system significantly grows, the POP solution provided by solving (3), considering \( J_p(C) \) as defined in (7), will not significantly change tending, in the sense of a horizontal asymptote, to a quantity called Capacity Bound \( C_{\text{bound}} \). From a formal viewpoint,

\[
C_{\text{bound}} = \sum_{z=1}^{Z} C_{\text{z bound}}, C_{\text{bound}} < C_{\text{TOT}} \tag{11}
\]

where \( C_{\text{z bound}} \) is the portion of capacity allocated to the \( z \)-th Earth Station when the overall allocation converges on the defined bound. The mentioned \( C_{\text{bound}} \) exists and is finished if \( C_{\text{z bound}} \), \( \forall z \in [1, Z] \) is a quantity independent of \( C_{\text{TOT}} \) when \( C_{\text{TOT}} \) tends to infinity. [6] defines three conditions, regarding the objective functions used, that assure the existence of the capacity bound. Moreover it demonstrates also that the proposed \( L_p \)-CA, with the considered metrics, satisfies these conditions.

### IV. Performance Analysis

The scenario considered in this performance evaluation has been implemented through the \( \text{ns-2} \) simulator. It is composed by \( Z = 2 \) Earth Stations that transmit TCP traffic over a common geostationary satellite channel. The overall duration of the simulation is 300 [s]. The allocation is done each 5 [s] (i.e., allocation period) and the Fading Level suffered by each \( z \)-th station, expressed by \( F_z \), assumes values, kept constant in each allocation period, that belong to Table I. It is worth noticing that \( F_z \) is not included in (10) because the fading is supposed to be compensated applying the FEC encoding. As a consequence the terms \( F_z \) does not modify the value of TP, which depends only on the FSL.

The main contribution of this section is the investigation of the advantages of the proposed \( L_p \)-CA approach in terms of saved capacity, which are strictly related to the existence of the Capacity Bound. In more detail, the performed tests consider the two earth stations with a random fading level considered uniformly distributed among all possible levels and the code rate is consequently chosen according to the aforementioned table. Obviously, the fading level value is considered known when the allocation algorithm acts. Each value of the performance metrics reported in the figures below represents the average of the values obtained by a number of simulation runs aimed at guaranteeing a confidence interval of the 95%.

The Transmitted Power [W] (TP) values, reported in the following figures, are computed through the objective function (9) itself on the basis of the allocated capacity \( C_z \); the packet loss has been computed, for each allocation period, as the number of packets lost for congestion over the number of sent packets, measured through
the available ns-2 simulation counters. For this reason, in the following, the results are presented in terms of Packet Loss Rate (PLR). Five weighs configurations are applied to give more, or less, importance to an objective function with respect to the other for both the stations \( (z = 1 \text{ and } z = 2):\) 
\[ w_{1,1} = 0.1 \text{ and } w_{2,1} = 0.9; \quad w_{1,1} = 0.25 \text{ and } w_{2,1} = 0.75; \quad w_{1,1} = 0.5 \text{ and } w_{2,1} = 0.5; \quad w_{1,1} = 0.75 \text{ and } w_{2,1} = 0.25; \quad w_{1,1} = 0.9 \text{ and } w_{2,1} = 0.1. \]

The practical purpose of this section is to explicitly evaluate the ratio between the obtained capacity allocations and the related performance with \( L_pCA, \) which operates at the \( C^{\text{bound}} \) value, and the reference case in which the overall capacity \( C_{TOT} \) is employed.

To reach the aim, we define the quantity, \( C^\text{ref}_z, \) for the \( z - \text{th} \) station, which represents the capacity allocated to the \( z - \text{th} \) station according to its channel condition, applying the \( L_pCA \) and imposing that the overall capacity is used, in practice \( \sum_{z=1}^{Z} C^\text{ref}_z = C_{TOT}. \) Similarly, the quantity \( PLR_z(C^\text{ref}_z) \) and \( TP_z(C^\text{ref}_z) \) are, respectively, the Packet Loss Rate and the Transmitted Power of the \( z - \text{th} \) station obtained allocating to it \( C^\text{ref}_z. \)

In this work we define the Capacity Saving Ratio \( (C_{SR}) \) as the quantity reported in (12), which is the percentage of the capacity saved using \( C^{\text{bound}} \) (i.e., to employ \( C^{\text{bound}} \) for the \( z - \text{th} \) station), with respect to the overall capacity available, \( C_{TOT} \) (i.e., to employ \( C^\text{ref}_z \) for the \( z - \text{th} \) station):

\[
C_{SR} = \frac{C^{\text{bound}} - C_{TOT}}{C_{TOT}} \cdot 100 \tag{12}
\]

In the same way, we define the Transmitted Power Saving Ratio \( (TP_{SR}) \) as the percentage of the average TP saved with respect to the TP obtained sharing the overall available capacity: (13).

\[
TP_{SR} = \left| \frac{\sum_{z=1}^{Z} (TP_z(C^{\text{bound}}) - TP_z(C^\text{ref}_z))}{\sum_{z=1}^{Z} TP_z(C^\text{ref}_z)} \right| \cdot 100 \tag{13}
\]

Similarly, the Packet Loss Rate Increasing Ratio \( (PLR_{IR}) \) is defined as the percentage of PLR increase obtained allocating \( C^{\text{bound}} \) to the \( z - \text{th} \) Station with respect to the PLR obtained allocating the overall capacity among the station:

\[
PLR_{IR} = \left| \frac{\sum_{z=1}^{Z} (PLR_z(C^{\text{bound}}) - PLR_z(C^\text{ref}_z))}{\sum_{z=1}^{Z} PLR_z(C^\text{ref}_z)} \right| \cdot 100 \tag{14}
\]

The defined metrics are reported in figures 1, 2 and 3. In all cases, the advantage of the \( L_pCA \) employment is clear and very satisfactory.

A further consequence of the capacity saving obtained with the \( L_pCA \) is the possibility to serve other stations with the same channel, without the necessity of increase the \( C_{TOT} \) and, simultaneously, without a degradation of the performance in terms of PLR and TP. Obviously, the number of stations depends on their channel condition but it is possible to compute the average number as we define in the following:

\[
N(C_{TOT}, C^{\text{bound}}, Z) = \left| \frac{C_{TOT} - C^{\text{bound}}}{C^{\text{bound}}} \right| \frac{Z}{\sum_{z=1}^{Z} PLR_z(C^\text{ref}_z)} \tag{15}
\]

where \( Z \) are the Earth Stations that transmit traffic over the satellite link, and \( N \) is the number of Earth Stations that can be served with the same channel. In equation (15) the average increase of Earth Station number \( N \) is calculated as the inferior integer part of the product between the capacity saving obtained applying the \( L_pCA \) given by \( (C_{TOT} - C^{\text{bound}}) \) and the inverse of the average capacity allocated to a single station (i.e., \( 1/Z \)). Obviously, this quantity increases if the overall capacity \( (C_{TOT}) \) increases or if the allocated capacity \( (C^{\text{bound}}) \) decreases.

In Figure 4 is reported the average increase of the number of Earth Stations \( N(C_{TOT}, C^{\text{bound}}, Z) \) that can transmit on
the satellite channel. In this case \(Z\) is set equal to 2, \(C_{TOT}\) varies in the range \([1 − 10]\)Mbps.

V. Conclusions

The work extends the capacity allocation method proposed in [5] modeled as the MOP problem. In particular, a criterion based on the \(L_p\)-problem is proposed to find out a capacity allocation, among Earth Stations, representative of a compromise if Packet Loss Probability (PLP) and Transmitted Power (TP) are taken into account as performance metrics. Moreover, applying the proposed \(L_p\)-problem based allocation and considering PLP and TP analytically modeled as in Section III, the obtained allocations converge on a Capacity Bound. This bound is independent of the overall capacity available \(C_{TOT}\). The performance analysis shows that the proposed method enables a significant capacity and TP saving and, simultaneously, a limited worsening of the packet losses. From the practical viewpoint, the paper contribution implies that a Service Provider may provide capacity allocations to \(Z\) Earth Stations without employing its overall available capacity and may dedicate the rest of it to other possible entities without penalizing the overall performance and avoiding capacity wasting.

![Figure 3: Transmitted Power Saving Ratio using \(L_p\)CA.](image)

![Figure 4: Average increasing of the number of Allowed Earth Station using \(L_p\)CA.](image)
New Optical Communication Capabilities Using Nanosatellites

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Abstract—In this paper, the benefits derived from the exploitation of nanosatellites for complex telecommunication missions are presented, along with their limitations, and laser technology is identified as a viable solution to dramatically increase the downlink and crosslink capabilities of such miniature platforms. Then, the critical technological issues concerning the integration of an optical telecom terminal on a nanosatellite platform are discussed, considering the perspective advancements in areas such as power generation, propulsion and attitude control. The incompatibility of laser terminals pointing requirements with the current and perspective nanosatellite attitude stability has been identified as the most critical issue to be addressed in order to implement optical links on nanosatellites. A possible solution is presented at the end of the discussion.

Keywords-nanosatellite; optical link

I. INTRODUCTION

The term nanosatellite refers to the class of satellites weighing between 1 and 10 kg, which were initially conceived to offer students and young engineers valuable educational experiences with real flight programs; their exploitation was boosted by the introduction of the CubeSat standard in 2003. The interest showed by space agencies and academic institutions in such miniature platforms is constantly increasing because of their inherent advantages. The reduced mass, prototyping and production time of nanosatellites, as well as low launch cost as piggyback payloads allow universities, corporations and small companies to access to space easily. As of December 2012, 149 missions based on nanosatellites were flown and many others are on schedule [1], [2].

The inherent low cost of a single nanosatellite can be further reduced when mass production of standard units is applied, making these miniature platforms particularly appealing for the realization of satellite constellations suitable for earth observation, mapping or telecommunication applications. In particular, LEO telecom constellations based on nanosatellites could be attractive since they can provide global continuous coverage, overcoming the visibility limitations of large GEO platforms, and can be cost-effective even with respect to the existing LEO systems. The integration of high data rate, compact telecommunication terminals on nanosatellite platforms could pave the way to mission scenarios unexpected to this day.

In this work, three illustrative perspective scenarios have been identified and are briefly presented below.

1) Extension of GEO satellites coverage to polar areas and urban canyons. A LEO constellation of nanosatellites providing continuous coverage over polar regions can be used as a bridge to extend the coverage of GEO telecom satellites (see Fig. 1a), which is usually limited to the ±71° latitudinal band when a minimum elevation angle of 10° is considered. Also, the system can be used to better cover densely urbanized areas. This way, the capabilities of GEO satellites can be increased at a very low relative cost.

2) Data relay constellation for scientific satellites in LEO. Scientific missions in LEO produce a large amount of data, while few, short windows are available for downlink to ground stations. A constellation of miniature platforms capable of high data rate down/crosslinks providing continuous global coverage will make it possible to download data to ground stations not visible to the scientific satellites, almost in real time, as shown in Fig. 1b.

3) Low cost constellation for situational awareness. In case of unavailability of ground infrastructures due to natural or man-made disasters, situational awareness can be maintained exploiting a LEO nanosatellite constellation, as depicted in Fig. 1c. Crosslink capabilities could dramatically reduce the access time to any point of the Earth surface.

However, the advantages given by the exploitation of nanosatellites are counterbalanced by severe technical limitations, which make them unattractive for complex or high performance missions. Such limitations are represented by the short life (few years) and the limited on-board resources in terms of power and volume available to the payload, attitude control and telecommunications. In particular, downlink and crosslink capabilities are limited by both the on-board power and the poor attitude stabilization typical of nanosatellites, which often rely only on permanent magnet bars and hysteresis rods, preventing the exploitation of narrow-beam antennas for high data rate links. Typically, nanosatellite communication systems use the VHF or UHF frequency band, and are based on monopole or microstrip patch antennas, with wide beamwidth and data rate often bounded to a few tens of kbit/s. Sometimes the S band is
exploited for higher data rate. This is the case of the Generic Nanosatellite Bus [3], whose dual communication system uses a UHF monopole antenna receiver system for uplink at 4 kbps and an S-band transmitter patch antenna set for downlink, capable of data rates between 32 kbps to 256 kbps.

In this paper, the advantages of exploiting optical links on nanosat platforms are discussed in section II. The state of the art of the main technologies required for the realization of the aforementioned mission scenarios is summarized in section III, along with the perspective advancements in those fields. Finally, section IV presents the system that is developed by the authors to cope with the current technological limitations which these days prevent the integration of laser communication terminals onboard nanosatellites.

II. CREATING NEW COMMUNICATION CAPABILITIES USING NANOSATELLITES

In this section, the benefits derived by the exploitation of optical link devices onboard nanosatellites are discussed.

Telecom systems based on laser technology are characterized by very narrow beamwidth and high frequency carriers (hundreds of THz), which make it possible to obtain very high data rate transmissions with more compact and less power consuming devices, with respect to RF systems. In addition, optical links can provide secure channels, as they are immune to jamming and cannot be intercepted. Moreover, there are no licence regulations on the use of optical frequency bands and bandwidths, so great commercial opportunities are possible. Although it cannot be considered a standard technology yet, several optical link devices were developed and tested in the last years [4]-[6]; among them, examples of compact devices are represented by STRV-2 [7] and SOTA [8].

As an example, two RF and one laser-based communication systems for high data rate links are compared in order to determine which technical solution is more suitable for integration on nanosatellites class platforms. The carriers of the two RF systems are set to 2.5 and 20 GHz respectively, the former representing the state of the art of RF communication systems for miniature satellites and the latter representing perspective advancements in this field [9]. The selected laser wavelength is 1.55 μm. A LEO – LEO crosslink scenario is considered with an intersatellite distance of 5000 km and the link requirements are expressed in terms of desired data rate and BER, which are set to 500 Mbit/s and 1E-6 respectively. The former is set roughly 3 orders of magnitude greater than typical maximum data rates for nanosatellites. Then, in order to compare systems with equivalent performance, the RF and optical terminals design assumptions, such as antenna diameter and bandwidth, are set in order to provide the same link margin. For simplicity, the emitter and receiver terminals are supposed to be identical. After that, the link budget is assessed, leading to the estimation of the required power and system mass for all the considered solutions. As regards the optical system, the total mass and power consumption are estimated using empirical correlations between these parameters and the telescope aperture, which were derived from a literature survey on existing optical link devices. The estimates account also for the mechanical active pointing system. On the other hand, the antenna and transmitter mass, as well as the transmitter input power for the RF systems have been estimated considering the relations provided by [10]. An additional mass of at least 1 kg is added to account for the antenna steering mechanism, derived from a literature survey of analogous systems [11]-[12]. Table 1 summarizes the terminals design assumptions and presents the results given by the link budget calculation and the total mass and power consumption estimation; the total mass of the RF systems is given by the sum of the antenna, transmitter and steering mechanism masses. The pointing accuracy required by each system was also calculated according to [13] and is reported in the last line of Table 1.

The results point out that the laser-based system requires a much lower power input with respect to the RF systems and its mass is considerably lower. In fact, the total estimated masses of the RF terminals exceed the nanosat standard limits. However, the pointing accuracy needed to achieve the communication requirements is much more
TABLE I. RESULTS FROM THE LINK BUDGET FOR A RF SYSTEM AND A LASER-BASED TERMINAL PROVIDING A LINK WITH DATA RATE OF 500 MBIT/S, BER OF 10⁻⁴ AND SAME LINK MARGIN

<table>
<thead>
<tr>
<th>DESIGN ASSUMPTIONS</th>
<th>Unit</th>
<th>RF 1</th>
<th>RF 2</th>
<th>Optical</th>
</tr>
</thead>
<tbody>
<tr>
<td>Antenna diameter</td>
<td>[m]</td>
<td>0.3</td>
<td>0.3</td>
<td>0.02</td>
</tr>
<tr>
<td>Carrier frequency</td>
<td>[GHz]</td>
<td>2.5</td>
<td>20</td>
<td>193.5E3</td>
</tr>
<tr>
<td>Wavelength</td>
<td>[m]</td>
<td>0.120</td>
<td>0.015</td>
<td>1.55E-6</td>
</tr>
<tr>
<td>Bandwidth</td>
<td>[MHz]</td>
<td>500</td>
<td>500</td>
<td>500</td>
</tr>
<tr>
<td>Modulation</td>
<td>[ ]</td>
<td>BPSK</td>
<td>BPSK</td>
<td>BPSK</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>RESULTS</th>
<th>Unit</th>
<th>RF 1</th>
<th>RF 2</th>
<th>Optical</th>
</tr>
</thead>
<tbody>
<tr>
<td>Trx gain</td>
<td>[dB]</td>
<td>15.31</td>
<td>33.37</td>
<td>91.34</td>
</tr>
<tr>
<td>EIRP</td>
<td>[dB]</td>
<td>54.97</td>
<td>54.97</td>
<td>76.71</td>
</tr>
<tr>
<td>Trx power</td>
<td>[W]</td>
<td>&gt;1000</td>
<td>144.62</td>
<td>0.035</td>
</tr>
<tr>
<td>E/N₀</td>
<td>[dB]</td>
<td>12.00</td>
<td>12.00</td>
<td>12.00</td>
</tr>
<tr>
<td>Link margin</td>
<td>[rad]</td>
<td>1.30</td>
<td>1.30</td>
<td>1.30</td>
</tr>
<tr>
<td>Beamwidth</td>
<td>[rad]</td>
<td>0.49</td>
<td>0.06</td>
<td>96E-6</td>
</tr>
<tr>
<td>Total mass</td>
<td>[kg]</td>
<td>&gt;100</td>
<td>15</td>
<td>3.5</td>
</tr>
<tr>
<td>Total power</td>
<td>[W]</td>
<td>&gt;1000</td>
<td>395</td>
<td>20</td>
</tr>
<tr>
<td>Pointing accuracy</td>
<td>[rad]</td>
<td>0.020</td>
<td>0.003</td>
<td>4E-6</td>
</tr>
</tbody>
</table>

stringent for the optical system, as a consequence of the very narrow beamwidth.

III. TECHNOLOGICAL GAPS

In this section the main technological issues that must be addressed to enable the exploitation of nanosatellite platforms for optical communication are briefly discussed.

A. Miniaturized Optical Link Terminal

The mass of existing laser-based communication terminals for space applications ranges between 5 and 160 kg, while their power consumption is within 28 and 150 W. The main drivers for sizing an optical terminal are the telescope aperture and the laser output power, which together contribute to the transmission gain. With the recent availability of compact, reliable and more powerful laser diodes the telescope dimensions can be reduced while maintaining a high transmission gain, thus enabling the realization of low-mass terminals. An optical communication system for microsatellite platforms, with preliminary mass, power consumption and data rate assessment of 5 kg, 45 W and 2.5 Gbit/s respectively, is currently under development [14]; therefore, it is expected that in the near future terminals suitable for integration on nanosatellite platforms can be realized with current technology.

B. Power Generation

An estimated power consumption of 20 W for the miniaturized laser terminal results from the analysis performed in section II. According to [15], so far the average available power on-board nanosatellite platforms does not exceed 10 W, even when GaAs solar cells are used. The limiting factor is essentially the small area available to solar cells, even when deployable solar arrays are exploited. A promising solution to this issue is represented by rollable solar arrays based on thin polymide foils, which are characterized by a very high specific power generation (>200 W/kg) [16], [17]. A total power consumption of 60 W can be delivered by a deployable, thin film solar array of 0.4 m², for a total weight of only a few hundred grams. Thus, it seems that power available on nanosatellite platforms will increase dramatically in the next years.

C. Propulsion

Considering a mission scenario in which a nanosatellite constellation is exploited for telecommunication, propulsion capabilities are essential for orbit maintenance and correction. In high LEO (1000 km or more), the most significant long-term semi-major axis perturbation is due to solar radiation pressure, as the Earth atmosphere density is almost negligible; for a 10 kg platform, a total ΔV of 50 m/s is estimated for orbit corrections throughout a 5 years mission. It seems that such performance can be achieved using currently available technology, as promising results were obtained in the last years exploiting pulsed plasma thrusters and field emission electric propulsion systems [18], [19].

D. Pointing Accuracy and Stability

As pointed out in section II, lasercom devices require very stringent pointing accuracy and stability as a consequence of their narrow beamwidth. The needed pointing accuracy is at least in the order of a ten microradians. To this date, this requirement is far beyond the capabilities of any kind of attitude control system considering both nanosatellites and larger platforms. In addition, the bus attitude stability is critical to establish an optical connection between two spacecraft. As a reference point, the ADCS of the Generic Nanosatellite Bus for the BRITE mission [20], which is scheduled to launch in 2013, will be capable of unprecedented performance for nanosatellites, with attitude accuracy and stability of <1° and 60'' respectively, exploiting a miniature starracker for fine attitude determination and a set of nano reaction wheels. Such performance do not meet the aforementioned requirements, so it is clear that they can be achieved only with a dedicated pointing system. However, moving gimbals could generate additional disturbances the bus ADCS must deal with, which sums to the vibrations generated by large deployable solar arrays based on thin polymide films. At best, considering a nanosatellite platform provided with 3-axis attitude control and damping system, the residual vibrational environment of the bus in LEO is characterized by a few milliradians amplitude oscillations at low frequency (<5 Hz).

In summary, in the near future the capabilities of nanosatellites are expected to increase significantly as power generation and propulsion are concerned; in addition, recent advancements in laser technology will enable the realization of miniaturized optical terminals for
telecommunications. However, both the state of the art of ADCS and the expected advancements in this area seem insufficient to meet the attitude stability requirements needed for a steerable laser communication terminal.

IV. CONCLUSIONS AND FUTURE WORK

The exploitation of LEO constellations based on nanosatellites for telecommunication scenarios is attractive for the benefits derived from the low mass, production time and overall cost of such miniature platforms. However, the exploitation of nanosatellites for complex missions is still prevented by their current severe resources limitations. In this paper, optical communication has been identified as a viable solution to increase significantly the communication data rate available to payloads onboard nanosatellites. The technological aspects which play a key role in the realization of new mission scenarios have been discussed considering the current state of the art of nanosatellite technologies and their perspective advancements in the near future. Considering the actual limitations of ADCS of nanosatellites, the laser terminal pointing accuracy and stability have been identified as the most critical issues which must be addressed to enable the exploitation of optical telecom terminals onboard nanosatellite platforms.

In this framework, the exploitation of a miniaturized, actively stabilized platform as an interface between the optical terminal and the host spacecraft is the chosen solution to cope with the current nanosatellite ACDS limitations. The active rejection of both low and high frequency vibrations from the bus to the laser terminal and vice versa not only will enable the exploitation of optical links on miniature satellites, but it will also allow to relax the requirements on both the ADCS of the host spacecraft and the pointing mechanism of the optical terminal. The authors are currently developing an integrated system composed by a miniaturized laser communication terminal and an actively controlled ultra-stable parallel platform with 3 rotational dof (shown in Fig. 2). This system is conceived to be mounted on top of a nanosatellite bus with 3-axis attitude control based on standard technologies (3 reaction wheels and 3 torquerods), but provided with sufficient power generation capabilities to sustain the optical communication terminal. The system performance will be evaluated though laboratory test using an attitude simulator testbed.

Figure 2. Preliminary configuration of the actively-stabilized platform.

REFERENCES

Abstract—Nowadays, many embedded systems use specific data buses (like CAN in vehicles or MIL-STD-1553B in launchers) to ensure the exchange of data. To reduce the cost, the mass and to increase performance, a solution is to rely on a components off-the-shelf (COTS) technology. As switched Ethernet is a well-known solution and widely implemented, this technology is studied for the next generation of space launchers. In this paper, we focus on the reliability issue. In case of network components failure and without reconfiguration mechanism, the network will not carry on control data anymore. Performance of Ethernet reconfiguration solutions may be not enough efficient regarding real-time constraints. This paper aims hence at coping with real-time guarantees of self reconfiguration for Ethernet networks, especially in the space launchers. An experimental evaluation has been leaded on a real case study during a Centre National d’Etudes Spatiales (CNES) French Research & Technology (R&T) activity.

Keywords—fault-tolerant network; industrial Ethernet; real-time.

I. INTRODUCTION

Nowadays, many embedded systems (aircrafts, satellites, launchers) use several specific data buses which ensure the exchange of either the control data or the service data (measurement or supervision). Particularly, the Ariane 5 launchers use a MIL-STD-1553B data bus [1] to transport the control data while the telemetry data flow is managed by a dedicated (and proprietary) bus. The MIL-STD-1553B bus relies on a centralized communication scheme inherent to its command/response mechanism. As illustrated in Fig. 1, the duplex architecture was seen as a way to ensure reliability (this choice was not derived inherently from system requirements). In this architecture, the bus and the end-nodes are therefore redundant. When a failure occurs on the bus (or a bus segment), the control of the launcher is based only on the redundant sensors data (not concerned by the failure). As a consequence, the communication system is tolerant to a single failure.

Currently, there are many new objectives for the next generation of space launchers [2]:

- reduction of the total cost of the launch system,
- avionics mass reduction and
- avionics performance improvement (keeping at least the same reliability)

To meet these objectives, there is an attempt to rely on an unique components off-the-shelf (COTS) technology. In this context, the different networks could be unified in a single communication system which will support more and more traffics or even large data (such as images, videos or extended telemetry data). As a consequence, the bus MIL-STD-1553B become unsuitable (due to its low bandwidth).

Among several COTS networks, switched Ethernet is the most cost and mature effective solution [3]. Indeed, this technology is a well-known standard and widely implemented in several domains [4] such as manufacturing systems, automotive [5], transport, aviation [6], motion control [7]. As in all industrial area, it reclams real-time and reliability requirements, such that the network must fulfill with Quality of Service (QoS) criteria like end-to-end delays, jitter, losses and availability. Several works have dealt with the temporal requirements of switched Ethernet architectures [8], [9], [10]. From this temporal aspect, the feasibility of switched Ethernet architecture (depicted in Fig. 2) in the next generation of space launchers has been highlighted in [11]. The purpose of this paper is to increase the robustness of switched Ethernet architectures up to (at least) the same reliability level than with MIL-STD-1553B.

As in all systems, the reliability in a switched Ethernet architecture can be increased with redundancy. This one can be classified into two categories (see IEC SC65C WG15 and [12]): “redundancy in the network” or “redundancy in the nodes”. The first one suggests to extend the network topology with redundant links (and even switches): end stations are attached to a single switch only, and hence interconnected to a single network. In the second one, the nodes have on the contrary (at least) two ethernet cards connected to two (parallel) redundant Ethernet networks.

Figure 1. Overview of current MIL-STD-1553B architecture
In reality, MIL-STD-1553B solution is based on a duplex architecture. However, it could lead to cost increase and all messages sent by all end nodes on the bus concerned by failure are still lost (the application was based only on the redundant sensors data). As introduced in [12], we suggest to refer to a “redundancy in the network” solution in which the network can reconfigure itself to ensure the continuity of service. It may be noticed that Automatic Repeat reQuest (ARQ) protocols can improve the reliability, but regarding the real-time requirement, none are considered here. To face with a failure of the local interconnection point of a device, redundancy of the end nodes will be used. Our objectives are therefore to have a reconfiguration mechanism transparent for the application point of view, and to minimize the time of the reconfiguration action (and so the losses). It corresponds to a real-time requirement for the reconfiguration strategy [13], and will be qualified in the following as:

**THE RECOVERY TIME:** the time necessary to switch to a backup solution and ensure the continuity of service. An application can tolerate a limited time of interruption of service, called grace time [14].

To minimize the recovery time, the literature reports many solutions [15] which rely on redundancy management. Redundancy is here supported by a physical meshed architecture. Relevant protocols in this category, Media Redundancy Protocol (MRP) [14] and Rapid Spanning-Tree Protocol (RSTP) [16] are based on two main steps: fault detection and topology reconfiguration. Fault detection is typically achieved by the asynchronous periodic transmission of particular frames between switches. Recovery time issues for real-time systems are coming from these exchanges and from the asynchronism due to the distribution of the algorithm. In order to reduce the recovery time, MRP protocol considers a centralized approach, but only for a ring topology. For all these reasons, this paper will suggest a novel centralized approach for the physical topology supervision and the logical topology reconfiguration. Another feature is that it can be applied in any physical meshed topology.

The remainder of the paper continues as follows: the section II presents an overview of our proposition and describes formally the proposal. Section III illustrates the proposed method by its application to a real case study obtained in the framework of a “CNES french R&T activity” and highlights the performance of our solution. Finally, in section IV some conclusions and future work are given.

**II. RESEARCH PROPOSAL: AUTONOMIC SWITCHED ETHERNET RECONFIGURATION**

The objective of the paper is to suggest a fast recovery solution enabling to ensure fast recovery time in harsh environment. In the framework of CNES activities, this paper deals primarily with space applications in which typical expected QoS criteria are: 1 ms end-to-end delays, any loss of frames, no current requirements in terms of bandwidth (note that in the future, it could be the case, e.g., for the telemetry traffic). This solution is particularly proposed for any meshed switched topologies.

**A. Proposition overview**

The core idea of our proposition is to reject the logical trees to control all paths in a deterministic way. It is suggested to apply a layer 2 routing to control the “route” taken by each flow. This consists of filling in the forwarding table a priori. Indeed, this table acts as a filter which enables the switch to match an incoming frame to an appropriate MAC-destination address entry and forward the frame to the next hop (with a specific output port) towards the final destination as specified in this table.

Let us remember that the objective of our proposition is that the network could reconfigure itself when a failure occurs. As a consequence, the network should be able to detect a failure and modify (or more exactly reconfigure) the forwarding table to restore an active (and known) path. To achieve this, we suggest to introduce a central agent capable of detecting any failure and reconfiguring an active logical topology. This agent has a global view of the network (and so of the network state). This view can be obtained by requesting all switches periodically (via Simple Network Management Protocol - SNMP). Indeed, any switches store the states of its ports in a specific database called Management Information Base (MIB). This information is updated thanks to a physical link integrity function (defined in the standard 802.3) on any switches.

Therefore, from the global view of the network, the central agent can detect a link or a switch failure. In case of link failure, the port state (concerned by the failure) is modified (from “up” to “down” state). When a switch failure occurs, the switch is not reachable anymore (no response to the periodic request).
Once a failure detected, the central agent sends a reconfiguration message in order to modify a (or many) output port(s) in the forwarding table. Although exchanges induced by this reconfiguration mechanism will lead to an over-consumption of the network capacity, it will remain highly limited compared to a capacity of 100 Mb/s and to the current amount of traffic (around 500 K/s) for Ariane 5 launcher. It leads to a set of new paths between two nodes. As in RSTP, paths will be computed along a spanning-tree. In this proposition, paths with unsuitable end-to-end delays will be rejected [11]. In fact, paths will be computed by a routing algorithm such as Dijkstra’s algorithm or Bellman-Ford algorithm. To maintain a deterministic control, they will be computed off-line before the configuration of the network, i.e. in the design phase. The number of paths will be related to the number of flows and to the reliability (e.g. for a single recovery, at least two times the number of flows, but in reality less).

The next subsection presents in a formal way, the reconfiguration algorithm which will be implemented in the central agent. It is planned that hardware implementation issues will be dealt with as well as the distribution of this functionality for facing with this central point of failure.

B. Formal considerations

In a formal way, we consider a network graph \( G = (X, A) \) where \( X \) is the matrix containing the state of all switch ports and \( A \) the matrix representing the physical connectivity between the switches. Let \( n \) the number of switches in the architecture and \( m \) the number of ports that a switch has (in this paper, it is assumed that all switches have the same number of ports). The element of \( X \) is written as \( X_{i,j} \in \{0,1\} \) which represents the state of switch \( j \in \{1, n\} \) port \( i \) with \( i \in \{1, m\} \). \( X \) is defined such as \( X = [X_1 \ldots X_n \ldots X_m] \) where \( X_j \) represents the port states of switch \( j \). The element of \( A \) is noted as \( A_{i,j}(i') \in \{0,1\} \) represents the physical capacity to communicate between the port \( i \) of the switch \( j \) and the port \( i' \) of switch \( j' \). This network graph will have to be updated periodically as written in line 3 of Algorithm 1. Indeed, in monitoring the states of ports (the matrix \( X \)), the network graph is accordingly modified. It is important to note that the performance of our solution will depend on the value of this monitoring period. This criterion will be evaluated in the next section.

Let us remember that our goal is to fill in or update the forwarding table. For that, we have to look for the output ports associated to each destination on all switches. A list \( D \) of all stations (represented by their MAC addresses (\( @MAC \))) is defined. It is necessary to know on which port of which switch are located the end-stations. A matrix \( D' \) is thereby defined to contain the network attachment of the different stations such as \( D'_{i,j} \in D \) represents the connection of end stations (with MAC address \( @MAC \)) on the port \( i \) of switch \( j \). As a matter of course, for each switch and for each MAC address in the list \( D \), we can identify on which switch \( j' \) is located an end-station (see lines 4-6 in Algorithm 1).

This information and the updated network graph are the inputs of the well-known Bellman-Ford routing algorithm (see line 7 in Algorithm 1). The result of this step is the next hop to reach the destination. Knowing the switch to cross and the physical connectivity (the matrix \( A \)), we can deduce the output port to fill in the forwarding table. Of course, if it is concerned by the failure, it is modified (see lines 8-9 in Algorithm 1) in the forwarding table.

III. EXPERIMENTAL WORK

Our research laboratory collaborates closely with the CNES to lead R&T activities. In this framework, the objective is to reach a sufficient level maturity of switched Ethernet technology for the future Ariane launchers. This level can be assessed by the Technology Readiness Level (TRL) [17]. In this work, the TRL 3 is expected and consists to study separately each elements (and problem area) of the technology. It should constitute a “proof-of-concept” without being in an operational environment. As a consequence, a computer network is used to show the feasibility of our algorithm, the reachable interruption time of service and the impact of our algorithm on the number of lost frames. However, a real applicative traffic is considered to perform the laboratory tests and to measure parameters of interest.

The tested architecture is depicted on the Fig. 3. It corresponds to the upper stages (Cryogenic Upper Stage - CASE - and Cryogenic Propulsion stage - EPC - as shown in Fig. 2). In this experiments, the switches are Cisco 2950 series (IOS version: 12.1(22)EA1). In a formal way, this example can be initially described by the matrix \( X \), \( A \), \( D' \) and the vector \( D \) as can be seen in (1) and (2).

\[
X = \begin{bmatrix} 1 & 1 & 1 & 1 \\ 0 & 0 & 0 & 1 \\ 1 & 1 & 1 & 1 \\ 1 & 1 & 1 & 1 \end{bmatrix}, \quad D = \begin{bmatrix} @MAC1 \\ @MAC2 \\ @MAC3 \\ @MAC4 \end{bmatrix}, \quad D' = \begin{bmatrix} @MAC1 \\ @MAC2 \\ @MAC3 \\ @MAC4 \end{bmatrix}, \quad \dim(X) = 24 \times 4, \quad \dim(D) = 13 \times 1 \quad \text{and} \quad \dim(D') = 24 \times 4
\]
Among the tested monitoring period, the reconfiguration time and the update time in the MIB.

The variation of the interruption time can be explained by the detection time of “link loss” which is between 50 ms and 150 ms (defined in the standard 802.3 [18]), plus the reconfiguration time and the update time in the MIB. Among the tested monitoring period, the 100 ms period is the closest to the detection time. That’s why, it can be considered a period equals to 100 ms. Moreover, when the frame with the forwarding update is sent, we record this date. We can therefore deduce a (bounded) approximation of the reconfiguration time which is the difference between the end of interruption of service and this date. The reconfiguration time is around 20 ms.

A second test has been leaded on three of the flows which are the flow 1 on the device 13, the flow 2 on the device 9 and the flow 3 on the device 8. These flows are interesting because of their different periods (288 ms, 72 ms and 18 ms). The objective is to highlight the impact of our algorithm on the number of lost frames. The same failure as previously (on the link Sw1-Sw3) is considered. The forwarding table of the switch Sw3 is initially configured as shown in Table II.

Consider, for instance, the flows of the device 13 (to device 1) and the link Sw1-Sw3 failure. As the graph is updated periodically (see lines 2-3 of Algorithm 1), the failure will be observed in the matrix X (once the failure is electrically detected via the physical link integrity test in the switch and reported in his MIB) as can be seen in 3.

\[
X = \begin{pmatrix}
1 & 1 & 1 & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & 0 & 0 & 0 \\
1 & 1 & 1 & 0 & 0 & 0 & 0 \\
1 & 1 & 1 & 0 & 0 & 0 & 0
\end{pmatrix}
\]  

The algorithm calculates (or more precisely it has been offline precalculated as mentioned in II.B) the next hop (the new output port to reach the destination, i.e. device 1 in this example). If we apply the algorithm on this example, for the third switch and the device 1 (which is the destination of the flows of the device 13) with MAC address @MAC1 (lines 4-5 in Algorithm 1), we obtain:

line 6: \((i,j') = (1,1)\)

line 7: the Bellman-Ford algorithm calculate the next hop to reach the switch \(j' = 1\) from the switch \(j = 3\) according to the current graph \(G\). The Bellman-Ford will have the choice between the ports 23, 24 as next hop because the matrix

![Figure 3. Architecture of case study](image1.png)

![Figure 4. Service interruption time](image2.png)
A gives the possible output ports according to the physical connectivity (i.e. ports \{22,23,24\}) and the matrix \(X\) gives the output ports states (i.e. port 21 = 0 so “up”, port 23 = 1 so “up” and port 24 = 0 so “down”). Consider the following Bellman-Ford result: next hop = 23. As a result, the vector \(R_i\) equals to \(<@MAC1, 24>\) (before the failure) is updated to \(<@MAC1, 23>\).

Consequently, the forwarding table of the switch \(Sw3\) will be therefore modified as shown in Table III (the parameter \(dot1dStaticAllowedToGoTo\) corresponding to the output port becomes 0x00 00 02 (i.e. the port 23). Notice that the forwarding table of the switch \(Sw1\) will be also modified (by the algorithm) such a way that the messages of the devices 8, 9, \ldots, 13 are sent on the output port 22. Moreover, in this example, any modifications will be done on the switch \(Sw2\) because its forwarding table is initially configured to send the messages from the devices 8, 9, \ldots, 13 to the devices 1, 2, \ldots, 7 via the output port 22, and the messages from the devices 1, 2, \ldots, 7 to the devices 8, 9, \ldots, 13 via the output port 23 (which is the initial Bellman-Ford result).

Table IV shows the number of lost frames in the case of Sw1-Sw3 failure (the cable Sw1-Sw3 is unplugged) by using our solution (measures) or RSTP (the interruption time for RSTP has been considered as being of 1 s [15] for computing the losses). Any flows sent by the devices 2, 3, \ldots, 8 are lost. Indeed, the failure does not occur on their paths. For the other flows sent by the devices 9, 10, \ldots, 13, the number of lost frames depends on the period of the flow and the interruption time. It is therefore possible to see that there is no loss for all flows with a period of 288 ms (vs. 4 by using RSTP). Only two frames have been lost for the flows with a period of 72 ms (vs. 14 by using RSTP).

In the case of flows with a period of 18 ms, 14 frames are lost (vs. 56 by using RSTP). However, in the worst case, it could be observed 1 (respectively 3 and 15) loss(es) for the flows with a period of 288 ms (respectively 72 ms and 18 ms).

Fig. 5 and 6 show the evolution of the end to end delays. When a failure occurs, the reconfiguration involves a longest path (because the shortest path relatively to the number of hops had been set first). Consequently, the delay increases when the reconfiguration is effective as shown in the two graphs. In this example, the delay growth is equivalent to the switch latency (around only one transmission time because of switching mode: store & forward). However, this growth is proportionally limited (relatively to the service interruption time). Let us note that the zoom on each graph enables to estimate (and retrieve as presented in Table IV) the losses in regard to the interruption time due to the failure.

### IV. Conclusion and Future Work

This paper suggests a novel centralized approach for the physical (meshed) topology and the logical topology reconfiguration. Experimental measurements have shown that the recovery time has been decreased and the number of losses is lower compared to RSTP (although it is still non-null especially for the flow with a small period).

Finally, the future work will integrate this solution on

#### TABLE I. EXPERIMENTAL TRAFFICS (BYTES, MS)

<table>
<thead>
<tr>
<th>Device</th>
<th>Flow 1</th>
<th>Flow 2</th>
<th>Flow 3</th>
<th>Flow 4</th>
<th>Flow 5</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>length</td>
<td>period</td>
<td>length</td>
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<td>length</td>
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<td>72</td>
<td>12996</td>
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<td>72</td>
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</table>

#### TABLE II. INITIAL FORWARDING TABLE OF THE SWITCH SW3

<table>
<thead>
<tr>
<th>dot1dStaticAddress</th>
<th>dot1dStaticReceivePort</th>
<th>dot1dStaticAllowedToGoTo</th>
<th>dot1dStaticStatus</th>
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</thead>
<tbody>
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<td>0x00 00 00 01</td>
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<tr>
<td>@MAC7</td>
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<td></td>
</tr>
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<tr>
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<td>0x04</td>
<td>3</td>
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</tbody>
</table>

#### TABLE III. FORWARDING TABLE OF THE SWITCH SW3 AFTER RECONFIGURATION

<table>
<thead>
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<th>dot1dStaticAddress</th>
<th>dot1dStaticReceivePort</th>
<th>dot1dStaticAllowedToGoTo</th>
<th>dot1dStaticStatus</th>
</tr>
</thead>
<tbody>
<tr>
<td>@MAC1</td>
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<td>0x00 00 00 02</td>
<td>3</td>
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<tr>
<td>@MAC7</td>
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<tr>
<td>@MAC8</td>
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<td>0x80</td>
<td>3</td>
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<tr>
<td>@MAC9</td>
<td>0</td>
<td>0x40</td>
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<tr>
<td>@MAC13</td>
<td>0</td>
<td>0x04</td>
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</table>
a laboratory demonstrator (target TRL4) in order to study the impact of this active solution on others QoS parameters (e.g. the consumed bandwidth and its potential impact on the applicative end-to-end delays) and its robustness in embedded environment. Industrial switches will hence be used, which may very well decrease the detection time (and so the interruption time) if the “down state” is instantaneously notified (contrary to the values of “link loss” in the link integrity test defined in the standard 802.3).

REFERENCES


| Table IV. LOST FRAMES PER FLOW BY USING OUR SOLUTION (A) OR RSTP (B) IN THE CASE OF THE SW1-SW3 FAILURE |

<table>
<thead>
<tr>
<th>Device</th>
<th>Flow 1</th>
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<th>Flow 3</th>
<th>Flow 4</th>
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Figure 5. Delay of the flow 1 on device 13 (period: 288 ms)

Figure 6. Delay of the flow 3 on device 8 (period: 18 ms)

Development of Wideband Dual Polarized of Microstrip Antennas for Microwave Remote Sensing

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Abstract—A novel wideband dual polarized L-band antenna for NASA’s microwave remote sensing applications has been designed, fabricated, and validated. The novel antenna structure consists of a rectangular microstrip antenna excited by capacitive coupled four probes. One of the linear polarizations is obtained by feeding two diagonally opposite probes with equal but out of phase signals. Other orthogonal polarization is obtained by feeding other remaining probes with equal but out of phase signals. The novel structure provides the band width in excess of 500 MHz (with the center frequency of 1.3 GHz) and the cross polarization below -40 dB. For validation purpose, two prototype antennas were fabricated and tested. The test results collected show relatively good agreement with simulation results. For achieving wider bandwidth, it is required to use, in our design, the probes with diameter larger than conventional coaxial connector. A novel feed structure using larger size cylindrical via is designed. The new feed structure design is fabricated and validated before its integration with the antenna.

Keywords—microstrip; L-band antenna; microwave; NASA.

I. INTRODUCTION

Microwave radars/radiometers are often used by NASA and other agencies for remote sensing of the Earth science parameters. The spatial and temporal resolutions obtainable with the microwave radars/radiometer observation depend upon their frequency bandwidth. It is always desirable for high spatial resolution to have radar operating at much higher bandwidth. The current antenna technology offers antenna with 20-50 MHz bandwidth. We proposed to develop a wideband (40%, 500MHz), dual polarized L-band array antenna for the next generation of microwave remote sensing radars and radiometers. The novel wideband antenna configuration will enable the scientists to observe the Earth science parameters with less than meter spatial resolution. In addition, the dual polarization capability with very low cross-polarization coupling (< -50 dB) will allow to measure the Stokes parameters with enhanced accuracy/sensitivity (-50 dB). Accurate measurement of the Stokes parameters will yield better estimates of permafrost and biomass which are critical parameters for development of Earth’s carbon cycle model. It has been reported that dual polarization capability will provide biomass estimation with RMSE error of ~ 120 Mg/ha. 500 MHz bandwidth will allow achieving resolution in the range of 50 x 50 cm.

Researchers use different techniques and technology to design and fabricate an antenna. Microstrip has been one of the most popular microwave transmission lines in antenna. However, coplanar waveguide transmission lines have also been used in microwave transmission line as well. They are often selected over high frequency transmission line option such as stripline. John Coonrod has reported differences between microstrip and coplanar waveguide transmission lines to find the best performance applications [1]. Also, antenna size is always something that must be related to the wavelength and thus the frequency. Higher frequencies can utilize smaller antennas. However, the size of the antenna depends to the radiation resistance, the antenna feeding impedance, and radiation pattern [2]. Neil Chamberlain [3] has reported a method to fabricate a single-layer antenna that has application in military and commercial environments. He used a large-diameter center post as its supporting structure that allows for fabrication of a sufficiently rigid antenna element that can survive launch loads.

In this work, we present a procedure to design wideband dual polarized L-band antenna element and its microstrip feed line structure. We also describe a procedure used to fabricate and test these microwave components using the laboratory facility available at Microwave Instrument Technology Branch of NASA Goddard Space Flight Center (NASA-GSFC) in Maryland.

Figure 1. The computer model of L-band antenna
II. DESIGN METHODOLOGY

Two tasks need to be fulfilled to design and build microstrip antenna. First task starts with design of microstrip feed line using large size coaxial line. Fig. 1 shows the wideband dual polarized L-band antenna structure to be modeled. Our electromagnetic simulation of this antenna showed that for wideband operation, the four probes need to be of diameter much larger than the diameters of central conductor of a standard coaxial connector. Hence, a perfectly matched transition (as shown in Fig. 2) is required to connect conventional small size coaxial connector to a large size coaxial line. The desired transition has to transfer 100% of power between the standard coaxial inputs to the larger size coaxial output. However, to test such a transition we would require unconventional large size coaxial connectors. To overcome this difficulty, we used back-to-back transitions (shown in Fig. 2), which results in a structure shown in Fig. 3, and which only require standard coaxial connectors to test the power transfer through a large size coaxial line section. The transition shown in Fig. 3 consists of metal plate of size 43.18 x 17.78 x 1.27 cm (17 x 7 x 0.5 inches). On each side of the metal plate a dielectric substrate with dielectric constant of 2.94 was glued to the metal plate. On the top and bottom surface of these composite layers, a microstrip line having characteristic impedance of 50 Ohms was etched out to form 50 Ohms microstrip lines.

The width of the microstrip lines were calculated using [4]

\[ Z_0 = \left( \frac{87}{\sqrt{\varepsilon_r + 1.41}} \right) \ln \left( \frac{5.98 h}{0.8 w + t} \right) \text{ ohms} \]  \hspace{1cm} (1)

In (1), \( Z_0 = 50\Omega \) is the characteristic impedance, \( \varepsilon_r = 2.94 \) is dielectric constant of the insulator, \( h = 0.8 \text{ mm} \) is thickness of the dielectric insulator, and \( t = 0.08 \text{ mm} \) is thickness of the copper strip. The width, \( w \), of the microstrip using (1) is found to be 1.7 mm which was rounded to 2.00 mm. The coaxial line section used in the proposed transition has a cross section as shown in Fig 4.

In order to design the larger size coaxial line with characteristic impedance of 50 Ohms, (2) for a coaxial line was used to estimate inner and outer diameter of coaxial line section [5].

\[ Z_0 = \left( \frac{138}{\sqrt{\varepsilon_r}} \right) \log \left( \frac{D}{d} \right) \text{ ohms} \]  \hspace{1cm} (2)

In (2), the D represents the diameter of the hole in
the aluminum plate, and the \( d \) represents the diameter of the center conductor pin located at the center of the hole. The ratio of \( D/d \) can be calculated with assuming the value of \( Z_0 \) is 50 ohms and \( \varepsilon_r = 1 \) for value of air dielectric.

Using the dimensions obtained by these two equations, the transition was fabricated using fabrication facilities available at Microwave Instrument Technology Branch of NASA-GSFC. A photo of fabricated transition is shown in Fig. 5.

The second task starts with design of wideband dual polarized L-band antenna. The simulation model of the proposed L-band antenna is shown in Fig. 1. The critical parameters that control the antenna performance are (1) diameter of probe, (2) rectangular patch size, (3) height of patch from the ground, (4) separation of probes from center. Using the CST Microwave Studio electromagnetic simulation model, the antenna performance was optimized with respect to above parameters [6].

Commercial SolidWorks Computer Aided Design (CAD) software is utilized to draw and design antenna components for fabrication purpose. The geometry of the desired model is given in Fig. 6.

The two different types of L-band antennas were fabricated and tested. The first antenna had a thick substrate, while second antenna had a thin substrate. The L-band antenna consisted of four layers including an aluminum plate at the base, substrate with copper strip in the middle, aluminum ground plate attached to top of the middle substrate, and substrate with copper radiation patch at top. Four radiation poles are connected through ground plate and substrate to four copper transmitting strip lines. The commercial CST Microwave Studio software was used to obtain the desired specifications of the dimensions of the L-band antenna. The antenna was fabricated to the specified dimensions and it is shown in Fig. 7.
The second antenna has same specifications except using different substrate to compare the results.

III. RESULTS

The microstrip and two L-band antennas were tested and reported as following. The fabricated microstrip model shown in Fig. 5 was tested with the PNA analyzer to validate the design specification. The graphs in the results section indicate that the microstrip met the desired design. The results shown in Fig. 8 confirm that there was very low return loss (low S11 and S22) at both ports. The top two smith charts within the graph both represent return loss in ports one and two. The S11 and S22 curves are very minimal and tight to the center as expected. The bottom charts both represent transfer of power through the design structure from one port to the other. The results show about -0.13db of loss of power between the ports one and two. Note that these transitions were fabricated and tested for frequency range of 0.3-0.6 GHz as shown in Fig. 8. An identical approach was used to design a transition for the L-band.

Two samples of antenna shown in Fig. 7 were fabricated and measured. It was found that both samples gave identical performance within the experimental tolerances.

First the L-band dual polarized antenna with thick substrate was tested for its input Voltage Standing Wave Ratio (VSWR) and its impedance matching performance outside the anechoic chamber. It was also tested for its radiation performance inside the electromagnetic anechoic chamber room. Anechoic means no echoes which imply the room is free of reflections which gives accurate measurements. The antenna has four ports and to perform measurements two ports were terminated into matched load and other two were excited through a 3dB hybrid coupler. Fig. 9 shows input VSWR when two of four ports were excited through a 3dB hybrid coupler and other two ports were terminated into matched loads. Fig. 9 shows both measured and simulated results obtained using CST Microwave Studio software. A good agreement between calculated and measured VSWR is observed. Desired results are lower than -10 db’s for entire desired band showing a good impedance match for the antenna.
The next step was to test L-band dual polarized antenna with thin substrate. To configure the four port antenna into a dual polarized two port antenna, pair of two diagonally opposite probes is fed through hybrid couplers. This feeding arrangement converts the four port antenna into two port antenna. S-parameters (S11, S21, S12, and S22) of the dual polarized antenna were measured and shown in Fig. 10. The curves for S11 and S22 show that the antenna is impedance matched over the frequency band covering 1.0 – 1.7 GHz band. The curves for S12 and S21 show that the cross coupling between the two linear polarization is below -30 dB.

IV. CONCLUSION AND FUTURE WORK

In this project, the performance of fabricated microstrip antenna has been evaluated. Both measured and calculated results are presented and found to be in a good agreement. The SolidWorks CAD commercial software is utilized to visualize components for fabrication process. The model has been fabricated in a traditional machine shop and tested several times to collect data from the PNA Analyzer. Despite working with traditional method for fabricating microstrip and antenna, good agreement was achieved between the simulated and measured results. This project was involved extensive research, design, fabrication, constant testing, and troubleshooting. The wideband L-band designed and tested in this work is planned to be used in a large L-band array antenna for NASA’s remote sensing.

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REFERENCES

Hybrid Beamformer for TPMS Interference Suppression

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Abstract—The Tire Pressure Monitoring System (TPMS) is a type of the wireless communication device for vehicles, defined as a safety assist system for enabling drivers to efficiently check and manage tires. In the TPMS communications, external electric and electronic devices may interfere with the exact data transmission. In this paper, we propose a hybrid TPMS beamformer based on a switching beamformer and minimum-variance distortionless–response (MVDR) beamformer for selecting an efficient algorithm depending on circumstances in order to eliminate the interference and to transfer the accurate data to drivers. We also suggest an effective signal decision algorithm based on the signal-to-interference and noise ratio (SINR) in order to ensure the transferred data reliability. The performance of the suggested hybrid TPMS beamformer and the effective signal decision algorithm is identified through a computer simulation example.

Keywords-Tire Pressure Monitoring System (TPMS); Hybrid Beamformer; Switching Beamformer; Minimum-Variance Distortionless-Response (MVDR); Interference Suppression.

I. INTRODUCTION

The Tire Pressure Monitoring System (TPMS) is a safety assist device for measuring tire temperature and pressure by means of a sensor in each tire of a vehicle, transmitting the measured data to the signal processing unit of the vehicle, and displaying them in the display unit to inform the driver about the tire state information in real time [1]. The TPMS employs various carrier frequencies such as 433.92 MHz in U.S.A and Europe, and 433.92 MHz and 447 MHz in Korea.

Most of TPMSs currently employ the one-way wireless communication technique with low data transmission efficiency and decision reliability. In order to overcome this problem, we suggest a method for deciding the TPMS data reliability of high performance based on the bidirectional communications. An advantage of the bidirectional communications is that it may save the power of the battery in the sensor unit installed in each tire, because it transmits the data when only it is required. Also, since it contains Ack/Nack data which indicates the efficiency of the transmitted data and E/N data which indicates the condition of a tire, it may efficiently improve the data reliability. Decision of data efficiency by the suggested algorithm is based on the measured signal-to-interference and noise ratio (SINR).

The TPMS employing various frequencies across the globe may experience the serious interference by high-power signals from external electric or electronic interference sources. At the sample index $k$, the received signal is given by

$$ z(k) = \sum_{i=1}^{4} a_i g_i(k) h_i(k) + A j(k) + n(k) \quad (1) $$

where $a_i$ is an array response vector (size $M \times 1$) for the $i$th tire based on the uniform linear array (ULA), $g_i(k)$ is a cyclostationary Gold code (size $N$) for the $i$th tire, and $h_i(k)$ is a measured data bit for the $i$th tire, which remains

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The upper limit of the summation in (1) is four, because we assume that the number of tires is four. \( A \) is a \( M \times P \) array response matrix, \( P \) is the number of interference signals, and \( j(k) \) is an interference signal vector (size \( P \)). Also, \( n(k) \) is an AWGN vector composed of independent and identically distributed components, each with zero mean and variance \( \sigma^2 \). The angle-of-arrival (AOA) array response vector uses the equation described in [4,5]. Although we generally assume that the interelement spacing is a half of the wavelength, we are currently investigating the effect for reducing interelement spacing to a quarter or one-eighth of the wavelength. In order to save the length of entire space for the antenna elements, we are studying to employ the rectangular or the circular antenna array structure in the center of a vehicle.

An equation of the received signal SINR [6] for deciding effective signals is given by

\[
SINR = 10 \log_{10} \left( \frac{S.\text{Power}}{N.\text{Power} + I.\text{Power}} \right)
\]

(2)

where \( S.\text{Power} \) (Signal Power) is the signal power transmitted from each tire, \( N.\text{Power} \) is the noise power, and \( I.\text{Power} \) is the interference power. The received signal SINR is measured using preamble data, which a receiver knows beforehand.

III. EFFECTIVE SIGNAL DECISION ALGORITHM

In this section, we propose a TPMS effective signal decision algorithm based on the duplex TPMS data architecture [2].

A. Effective Signal Decision Algorithm in Signal Processing Unit

A flow chart for the effective signal decision in the signal processing unit installed in a vehicle is shown in Fig. 2. The signal processing unit measures SINR of the received signal, compares the measured SINR with a threshold. If the measured SINR is greater than the critical value, it sends the Ack signal, which means that the transmitted signal is effective, to the sensor unit to request transmitting the main data. On the contrary, if the measured SINR is smaller than the critical value, it does not send any signal and the sensor unit recognizes Nack which means that the transmitted signal is not effective. In this case, the sensor unit requests retransmitting the reference data. After receiving the main data, the signal processing unit decides the normal or abnormal state of a tire using the main data. If the tire is decided normal, the signal processing unit operates in the normal mode which indicates the normal tire state (for example, receiving data once per minute). On the contrary, if the tire is decided abnormal, the signal processing unit transmits E signal (for example, bit -1) which indicates an emergency mode, and operates in the emergency mode which indicates a dangerous tire state (for example, receiving data once per second). The data processed in the signal processing unit is sent to the display unit to enable a driver to check the tire condition.
**B. Effective Signal Decision Algorithm in Sensor Unit**

Fig. 3 shows a flow chart for the suggested effective signal decision method based on a duplex TPMS communication system in the sensor unit. The sensor unit receives Ack/Nack and E/N bits from the signal processing unit. If the sensor receives the Ack bit (for example, bit 1), it transmits main data to the signal processing unit. On the contrary, if the sensor unit receives Nack bit (for example, no bit), it retransmits reference data (preamble data) to the signal processing unit. If the signal transmitted from the main data is decided N bit (for example, bit 1), the sensor unit operates in the normal mode (for example, transmitting data once per minute). If it is decided E bit (for example, bit -1), the sensor unit operates in the emergency mode (for example, transmits data once per second).

**IV. HYBRID BEAMFORMER**

In order to efficiently suppress interference signals from external devices and receive the accurate data from the sensor unit, we propose a hybrid beamformer, which selectively uses the switching beamformer and MVDR beamformer after comparing the measured SINR and the given threshold. The switching beamformer has obviously low computational complexity comparing with the MVDR beamformer, because it does not require an auto-correlation matrix but the MVDR beamfore requires that. Hybrid beamformer forms beam in a desired direction to receive TPMS data. The received signal includes TPMS signals of each tire, interference from external devices, and noise. The hybrid beamformer basically uses the switching beamformer to weaken interference signals. If the output SINR of the switching beamformer is greater than the threshold, the output signal of the switching beamformer is analyzed to display the data in the display unit and thus to inform a driver about the tire condition. However, if the measured SINR is smaller than the threshold, the hybrid beamformer changes the mode from the switching beamformer to the MVDR beamformer as shown in Fig. 4.

Since the switching beamformer has very low computational complexity and has the relatively good performance of interference suppression, it is used as a basic beamformer of the hybrid beamformer. That is, if the measured SINR is greater than a specific threshold, the hybrid beamformer operates in the switching beamformer mode. However, if the SINR of the switching beamformer output is smaller than the threshold, it results the low ratio of the accurate data reception due to the interference signals. In this case, we employ the MVDR beamformer, which forms beam in the desired direction and nulls in the directions of the interference signals to minimize them, at the same time [7]. The MVDR beamformer has excellent performance of interference elimination, but it needs calculation of an auto-correlation matrix resulted in very high computational complexity. The proposed hybrid beamformer optimizes interference elimination performance and computational complexity.
complexity efficiency with two beamformers, which have obvious advantages and disadvantages.

A. Switching beamformer

The switching beamformer is defined as to alternately use more than one weight vectors based on the AOA vector to form beams to directions of interest. The switching beamformer weight vector for the $i^{th}$ tire signal with a beamformer architecture shown in Fig. 5 is given by

$$w_{s} = \frac{a_{i}}{\sqrt{a_{i}^{H}a_{i}}}$$

where $^{H}$ denotes complex conjugate transpose. The aforementioned weight vector generates a beam factor of size one for the $i^{th}$ tire to receive TPMS signal.

B. MVDR Beamformer

The MVDR beamformer calculates a weight vector for minimizing the power of the beamformer output while maintaining the power of the desired signal. The MVDR weight vector is computed from

$$\min w^{H}Rw \text{ subject to } a^{H}w = 1$$

where $R = E[zz^{H}(k)]$ is an auto-correlation matrix of the received signal [8]. The MVDR weight vector [9,10] for the $i^{th}$ tire is given by

$$w_{MVDR} = \left[a_{i}^{H}R^{-1}a_{i}\right]^{-1}R^{-1}a_{i}$$

It forms a beam with the factor of size ‘1’ to the direction of the $i^{th}$ tire signals and nulls interference signals to minimize them, at the same time. The hybrid beamformer output for the $i^{th}$ tire is given by

$$y_{i}(k) = w_{i}^{H}z(k).$$

This output contains the TPMS signal of the $i^{th}$ tire, noise, and residual interference signal. $w_{i}$ selectively uses the weight vector of the switching beamformer or MVDR depending on circumstances. The output of the hybrid beamformer is despreaded using the unique Gold code for the $i^{th}$ tire.

V. COMPUTER SIMULATION

Computer simulation result is described below to identify the performance of SINR threshold decision for an effective data decision method and the hybrid beamforming technology. We use a million received signal data for determining the SINR threshold for the suggested effective data decision method and set the number of antennas as one. Fig. 6 shows data reliability from –10 dB to 10 dB with respect to SINR. From the figure, we observe that increasing SINR results in rising data reliability. This result is utilized to determine the threshold for the proposed hybrid beamformer.

Six receiving antennas are used to identify the performance of the suggested hybrid TPMS interference suppression. It is based on the assumption that the received signal included three interference signals and noise, and a Gold code of $N=15$ is given to the TPMS signal of each tire. The incident angles of transmitted signals for each tire are assumed 60°, 120°, 240°, and 300°, respectively. The incident angles for 3 interference signals are randomly assumed 89°, 175°, and 340°, respectively. It is identified that the output SINR threshold value, which is a reference of data reliability 99% is about 4.4 dB from Fig. 6. This value is assumed as a threshold in the simulation for evaluating hybrid beamformer performance.

Fig. 7(a) and 7(b) show beampatterns for the switching beamformer which forms beams to direction of incidents angles 60° and 300° with respect to the right-front and right–rear tire TPMS signals and which forms beams to direction of incident angles 120° and 240° with respect to the left-front and left–rear tire TPMS signals, respectively. Fig. 8(a) and 8(b)
show beampatterns for the MVDR beamformer which forms beams for right and left tire TPMS signals, respectively. Note that the MVDR beamformer forms nulls to incident angles of interference signals, unlike the switching beamformer. Since the array response vector includes the cosine function, the proposed hybrid beamformer alternately uses two weight vectors (not four weight vectors) for the front tire in the right side and the front tire in the left side to receive the data from all four tires [2].

Fig. 9 shows output SINR curve per signal-to-noise ratio (SNR) for interference-to-signal ratio (ISR) = 3 dB, for 99% data reliability (4.4dB SINR). From the figure, we observe that the switching beamformer is used when the SNR is greater than -4 dB because the output SINR is greater than the threshold. In
the section SNR = -4 dB or lower, the SINR is smaller than the threshold and the switching beamformer mode is converted to the MVDR beamformer mode. Fig. 10 shows the output SINR curve per ISR for SNR = 0 dB, for 99% data reliability (4.4 dB SINR). It has a section where SINR decreases and then increases at ISR = 4 dB, which shows the result that the hybrid beamformer operates in the switching beamformer mode and then is converted to the MVDR beamformer mode where SINR is below the threshold.

VI. CONCLUSION AND FUTURE WORK

Enactment for compulsory use of TPMS to inform a driver about the tire condition in real time is globally underway in order to prevent serious traffic accidents due to abnormal tires. Since the TPMS employs wireless communication technique, it is essential to ensure the data reliability. For the high TPMS data reliability, we considered the duplex wireless communication technique, which is more developed than the one-way wireless communication method used in the conventional TPMS, and suggested a data reliability decision method according to the threshold based on SINR of received signals. We also proposed a hybrid TPMS beamformer based on the switching beamformer and MVDR beamformer in order to effectively eliminate interference caused from external devices. An optimum algorithm between two beamformers is selected based on the measured SINR according to circumstances for exact data transmission. The unique Gold code is employed to each tire in order to eliminate interference signals from other tires and to reduce the power consumption of a battery in the sensor unit of TPMS. The performance of the proposed hybrid TPMS beamformer was illustrated via the computer simulation example. In the future work, we will consider the effects of the car body and the ground, and the coupling between antennas for TPMS.

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Multi-millisecond GNSS Maximum Likelihood Bit Synchronization Method

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Abstract—This paper introduces a new maximum likelihood bit synchronization algorithm that can be tailored to be used with different types of GNSS signals (GPS, GLONASS). A preliminary version of this work was presented in a previous paper by the same authors. In that paper, it was shown that this method allows to determine the positions of the bit edges in the signal using multiple millisecond correlation values. Using longer correlation times for bit synchronization decreases the peak processor load on embedded GNSS receivers that are based on carrier and code tracking loops for signal tracking, since for this kind of processing, longer correlation times allow for lower tracking loop update rates. In that previous paper it was also shown (through the use of simulations) that this new method presented lower error rates compared to the histogram method for fixed length observation intervals. The present paper extends these results in two ways. First, it generalizes the proposed method to N-millisecond integrations, providing a framework for automatic generation of the Viterbi state machine that drives the synchronization. Second, this paper provides a simple analytical expression that can be used to estimate the synchronization error rate of the algorithm, and identifies its most important design parameters.

Keywords—GPS; GLONASS; receiver; data modulation

I. INTRODUCTION

Global Navigation Satellite Systems (GNSS) enable any vehicle equipped with a receiver to determine its position and velocity based on the signals transmitted by a constellation of satellites placed in carefully controlled orbits. For each signal present in the antenna the receiver determines a pseudorange measurement, which is a quantity related to the true geometrical range (distance) between the receiver and the transmitting satellite. Using four or more of these measurements, the receiver can determine its spatial coordinates with an error in the order of a few meters. User velocity determination is performed in a similar fashion, using quantities related to the Doppler deviation of the signal carrier frequency.

There are currently two GNSS systems in full operational status: GPS, the GNSS System maintained by the United States Department of Defense (DoD), and its Russian counterpart GLONASS. There are other GNSS systems being developed by different national entities, but as of 2012 these are still on the planning stage or have been only partially deployed (COMPASS, Galileo). A detailed description of the signal of GPS and GLONASS can be found in the literature [1][2][3][4].

Both GPS and GLONASS satellites transmit data at 50 bps. GPS encodes the bits using bipolar NRZ pulses, while GLONASS uses Manchester coding. Before being able to decode the data frames transmitted by a satellite, the receiver needs to perform a synchronization process during which the position of the data bit edges is determined.

Because of the structure of the signal, there is only partial uncertainty in the knowledge of the position of the bit edges. GPS and GLONASS signals are spread using a DS-SS (Direct Sequence Spread Spectrum) periodic spreading sequence. This spreading sequence is 1023 chips long for GPS, and 511 chips long for GLONASS. In both cases the spreading sequence repeats itself after 1 ms. Data bit sign transitions are aligned with the start of the new period of the spreading sequence. One data bit pulse extends exactly 20 spreading sequence periods. Since the start of the each period of the spreading sequence is known to the receiver, then there are only 20 different possibilities for the data bit sequence alignment. During bit synchronization, the receiver correlates the signal with the periodic spreading sequence using integration intervals that extend over an integer number of spreading sequence periods, and processes the results in order to detect the position of the data bit edges.

The classic algorithm for GPS data bit synchronization is the histogram method [2][8]. This method searches for sign changes in consecutive 1 ms correlation results. The performance is adequate for carrier-to-noise-density \( C/N_0 \) (which is a quantity related to the signal-to-noise ratio) ratios above 30 dB, which is the normal operating condition for most outdoor GNSS receivers. Because of its simplicity, this method has been extensively used for general purpose embedded receivers.

There are other more complex methods that present higher sensitivity, allowing receivers to perform data bit synchronization under extremely low \( C/N_0 \) conditions such as those endured by GPS receivers for indoor applications and street level car receivers. These algorithms work by finding the bit edge candidate position that maximizes the recovered average bit energy [6][7][10], which is equivalent to choosing the maximum likelihood candidate [6]. These methods can work with \( C/N_0 \) levels down to 12 dB [7].

Slightly modified versions of these algorithms can be used for GLONASS. The modifications are necessary because
GLONASS uses Manchester coding for its data bits instead of bipolar NRZ.

Both the histogram and the energy maximization algorithms need to work on correlation samples obtained using 1 ms integration times in order to determine the bit alignment with millisecond level resolution. This means that the phase and code tracking loops in the receiver must be updated at a rate of 1 kHz in order to determine the correlation parameters (carrier initial phase and frequency, spreading sequence initial phase and chip rate) that need to be configured at the start of each correlation. This update rate is higher than the update rate typically used during later processing stages (100 Hz), but the amount of processing per update is roughly the same in both cases. Thus, the synchronization stage of GNSS receivers that use these algorithms represents an important fraction of the worst-case processor load of the design. This worst-case processor load is an important metric for real-time systems since it determines whether the system will be able to fulfill all of its processing deadlines in due time [5]. Also, since the synchronization stage represents a very small portion of the total processing time of a given GNSS signal, using the histogram, the bit energy maximization, or any other method that works on 1 ms correlation samples can lead to receiver designs with high peak-to-mean processor load ratios.

The worst-case processor load for a given design can be reduced using data bit synchronization algorithms that can achieve millisecond resolution using multi-millisecond correlation samples that require lower tracking loop update rates. This improvement in the worst-case processor load should not come at the expense of a decrease in the reliability of the synchronization algorithm.

In a previous work [11], a 3 ms maximum likelihood bit synchronization algorithm was proposed. The algorithm was simulated and compared against two other methods (histogram and a maximum bit energy algorithm similar to the one shown in [6]) on the grounds of error rates of each method for fixed size observation intervals.

The present paper extends these results in two ways. First, it generalizes the proposed method to N-millisecond correlation samples, providing a set of tools to help the design of the Trellis diagram that guides the synchronization state machine transitions. Second, this paper provides a simple analytical expression that can be used to estimate the synchronization error rate of the algorithm, and identifies its most important design parameters.

It is important to mention that while our previous work was aimed at achieving bit synchronization of 50 bps data signals of GLONASS and GPS, the present paper focuses on bit synchronization of a generic 100 bps bipolar NRZ. This is because both GPS and GLONASS signals can be thought as 100 bps bipolar NRZ signals whose data bits are encoded with a repetition code, and thus this 100 bps synchronization can be used as the first stage of a full 50 bps GPS/GLONASS bit synchronization algorithm. This is the approach used in a dual system GPS/GLONASS receiver that is being developed by the GNSS group in the Facultad de Ingeniería of the Universidad Nacional de La Plata.

The rest of this paper is organized as follows. Section II presents a basic baseband signal model that can be used to work with N-millisecond correlation samples and enunciates a few base assumptions. Later, the fundamental ideas of the maximum likelihood synchronization method are presented. Section III presents several simple rules to automatically build the synchronization Viterbi state machine for any given value of N. These rules were developed in order to code the simulations that were in turn used to validate the analytical expression of the probability of synchronization error that is introduced in Section IV. Simulation results are compared against theoretically calculated values of the probability of synchronization error in Section V. Finally, the conclusion is in Section VI.

II. ALGORITHM

A. N-millisecond Correlation Samples Model

The following sequence $I_1[i]$ can be used as the model of 1 ms correlation values produced by a GNSS receiver when processing a signal with data encoded as a sequence of 100 bps bipolar NRZ symbols.

$$I_1[i] = B[(i + m)/10] + n_1[i],$$ (1)

where $B[n]$ is a random process that models the random bit sequence ($B[n] = \pm 1$), $m$ is the unknown initial bit alignment (in milliseconds), $n_1[i]$ is random Gaussian random variable, independent for each $i$, and such that $E\{n_1[i]\} = 0$, $Var\{n_1[i]\} = \sigma_1^2$.

This model assumes that the beginning of the 1 ms correlation intervals are aligned so that they differ from each and every data bit edge by an integer number of milliseconds. This is a safe assumption since typically GNSS receivers are free to arrange the correlation intervals in the most convenient way, and the 1 ms correlation interval duration fits an integer number of times within the bit duration.

The previous model also requires some sort of normalization of the amplitude of the samples so that $E\{|B[n]|\} = 1$.

Samples generated when using N-millisecond correlation times can be modeled as

$$I_N[j] = \sum_{a=0}^{N-1} I_1[Nj + a].$$ (2)

Based on the definition of $I_1[i]$,

$$I_N[j] = S[j] + n_N[j],$$ (3)

where $n_N[j] = n_1[Nj] + ... + n_1[Nj + N - 1]$ is also a Gaussian random variable such that $E\{n_N[i]\} = 0$, and $Var\{n_N[i]\} = N\sigma_1^2$. The term $S[j]$ is the integrated value of the data signal during the N-milliseconds interval. It can
be seen that depending on the presence and position of a data bit sign change during the correlation interval, the value of this term will be $S[j] = \pm (N - 2k)$, where $k \in [0, N]$ is the position of the data bit sign change in milliseconds, relative to the start of the integration time.

B. Synchronization

In this section, we will skim over the main ideas of the synchronization algorithm since they were already discussed in [11].

Given a data bit sequence $B[n]$ and an alignment $m$, the sequence of N-millisecond correlation values $S[j]$ that will be observed is completely determined. This is also true the other way around: given a sequence of $L$ observed correlation samples $I_N[j]$, the data bit sequence $B[n]$ during the observation interval and the bit alignment $\hat{m}$ can be estimated if we find the sequence of expected correlation values $S[j]$ that best matches $I_N[j]$.

Using the Maximum Likelihood criterion it can be shown that the highest probability candidate $\hat{S}[i]$ is the one that minimizes the log-likelihood index $J$

$$\hat{S}[i] = \min_o J(I_N, S^{(o)}),$$

which is defined as

$$J(I_N, S^{(o)}) = \sum_{a=0}^{L-1} (I_N[a] - S^{(o)}[a])^2.$$  

In order to find the maximum likelihood candidate a Viterbi algorithm is used. One state is defined for each possible alignment of the beginning of the next N-millisecond correlation relative to the start of the current bit interval and for each sign. Since there are 10 possible different alignments and two bit values ($\pm 1$), 20 states need to be defined. Notice that the size of the state machine is independent of the length of the correlation interval $N$.

The state numbering scheme that will be used throughout this paper is shown in Figure 1. While any kind of state naming scheme is possible, this scheme is particularly useful because it allow us to enunciate a few very simple rules to automate the creation of the state machine transitions for any correlation length $N$.

There are $20 + 2N$ transitions in this state machine. States that are closer to the start of the bit than a correlation interval length have two possible arrival paths from previous states; the rest is only linked to a single previous state. Figure 2 shows the Trellis diagram for the case when $N = 3$. In Section III, the rules used to automate the creation of this Trellis will be given.

Initially each state has no accumulated quadratic error. For each observed integration value $I_N[j]$, the quadratic errors of the observed values against the expected values along each possible transition within the Trellis are calculated and added to the accumulated quadratic error of the state at the origin of this transition. If there is a single input path to the destination state of the transition, this updated accumulated quadratic error becomes the accumulated quadratic error of the destination state. If instead there are two possible arrival paths to the destination state, then the path with the highest updated accumulated quadratic error can be safely discarded since it can be anticipated that no matter what the future sequence of observed values is, in the end this candidate sequence will have a total accumulated error higher than at least one other candidate. The surviving path at the merge point then determines the quadratic error of the destination state. This way, at each step $2N$ candidates are eliminated at the merge points in the Trellis, and $2N$ new candidates are created at the fork points. After the last input sample $I_N[j]$ has been processed, the maximum likelihood observed sequence $\hat{S}[j]$ can be retrieved from the transition history of the candidate with the least accumulated error among the 20 survivors. If only the maximum likelihood alignment estimation is required, all the information needed is in the index number of the state with the least error: if $s$ is the state number with the least error, then at the start of the next correlation interval $\text{mod}(s, 10)$ milliseconds will have
passed after the last bit edge position.

III. TRELlIS CONSTRUCTION RULES

Because of the regularity of the state transitions, the Trellis for any value $N \in [2, 10]$ can be generated using a few very simple rules.

Let $st$ be the state number, using the same numbering scheme shown in Figure 1. Each state has at least one input transition that comes from the $st_{ss}$-th state

$$st_{ss} = \text{mod}(10 + st - N, 10) + 10 \lfloor st/10 \rfloor,$$

where mod is the remainder of the integer division. The expected observed value along this transition is

$$E_{ss} = (-1)^{\lfloor st/10 \rfloor} N.$$

If $\text{mod}(st, 10) < N$ there is a second input transition, from the $st_{ds}$-th state

$$st_{ds} = \text{mod}(st_{ss} + 10, 20).$$

This transition models the observed correlation value when there is a data bit sign change from the previous bit to the current one. Because of the sign change, the integrated areas at each side of the bit edge partially cancel each other, and thus the expected observed value depends on the alignment the state relative to the bit start

$$E_{ds} = (2 \text{mod}(st, 10) - N) (-1)^{\lfloor st/10 \rfloor}.$$

These rules were used to generate the trellis in Figure 2.

IV. ERROR EXPRESSION

Figure 3 shows the values of the final accumulated quadratic errors of the 20 states after having processed a signal with carrier-to-noise-density ($C/N_0$) ratio of 35 dB using 5 ms integrations ($N = 5$). Each state has an associated candidate correlation value sequence that can be recovered from the history of transitions within the Trellis. Pairs of states $s$ and $s + 10$ are both associated to candidates with the same bit alignment. These pairs of states also have similar accumulated quadratic errors because of the way the Viterbi algorithm forks and merges candidates as it moves forward.

The state with the least accumulated quadratic error determines the maximum likelihood bit alignment. Accumulated quadratic errors grow higher the farther away we move from the state with the maximum likelihood solution. The reason for this is that the surviving candidates associated to state numbers close to the one with the least error are not random but are in fact very similar to the maximum likelihood candidate. The farther away from the maximum likelihood candidate that we move, the smaller the amount of likeness, and thus the higher the accumulated error.

This observation about the likeness of the surviving candidates is not only qualitative. For high enough signal-to-noise ratios, it can be safely assumed that the second and third most likely alignments are associated to candidate correlation sequences that followed the same bit sequence than the maximum likelihood correlation sequence, but whose bit edges are displaced $\pm 1$ ms from the true value.

This, in turn, can be used to justify the following statement: the probability of incorrectly identifying the bit edge position using this algorithm is the probability of choosing instead a candidate sequence whose alignment differs by $\pm 1$ ms from the correct value.

Let $S^{\alpha}$ be the correct candidate for the received data bit sequence. Let $S^{\beta}$ be the candidate that would have been correct if the data bit sequence was the same, but the bit edges were delayed by 1 ms. The probability of mistakenly choosing $S^{\beta}$ over $S^{\alpha}$ as the maximum likelihood candidate is the probability of the event

$$J(I_N, S^{\alpha}) > J(I_N, S^{\beta})$$

$$\sum_{a=0}^{L-1} (I_N[a] - S^{\alpha}[a])^2 > \sum_{a=0}^{L-1} (I_N[a] - S^{\beta}[a])^2. \quad (10)$$

Since the bit sequences that generated $S^{\beta}$ and $S^{\alpha}$ only differ by 1 ms, the sequences $S^{\beta}$ and $S^{\alpha}$ differ only at $F$ sequence values, one for each data bit sign change that occurred (since $N \leq 10$, there can be at most a single bit sign transition during each correlation interval). Many terms then cancel out, leaving

$$\sum_{a=0}^{F} (I_N[a] - S^{\alpha}[a])^2 > \sum_{a=0}^{F} (I_N[a] - S^{\beta}[a])^2. \quad (11)$$

It is easy to see that since the original data bit sequences only differ by a single millisecond, the following equality holds for the remaining terms:

$$S^{\alpha}[a] - S^{\beta}[a] = \pm 2, \quad (12)$$
and thus
\[ \sum_{F} (I_N[a] - S^\alpha[a])^2 > \sum_{F} (I_N[a] - S^\alpha[a] + 2)^2. \]
Replacing \( I_N[a] = S^\alpha[a] + n_N[a] \):
\[ \sum_{F} (n_N[a])^2 > \sum_{F} (n_N[a] + 2)^2 \]
\[ 0 > \sum_{F} (\mp 4n_N[a] + 4) \]
\[ \sum_{F} \pm n_N[a] > F. \] (13)

\( n_N[a] \) are independent gaussian random variables, thus:
\[ P \left\{ \sum_{F} \pm n_N[a] > F \right\} = Q \left( \frac{\sqrt{F}}{\sigma_N} \right), \] (14)
where \( Q(x) \) is the probability that a Gaussian random variable will obtain a value larger than \( x \) standard deviations above the mean.

Finally, since there are two candidates that differ by exactly \( 1 \text{ ms} \) among the surviving candidates of the Viterbi processing, the final synchronization error expression is:
\[ P_e = 2Q \left( \frac{\sqrt{F}}{\sqrt{1000N}} \right), \] (15)

where in the last expression \( \sigma_N^2 = N\sigma^2 \) was replaced by its expression as a function of \( C/N_0 \).

It can be seen that the most important parameters in order to determine the synchronization error rate are the number of data bit sign transitions observed \( F \), the length of the correlation interval \( N \) and the carrier-to-noise-density \( C/N_0 \) ratio of the signal. While the value \( N \) is usually imposed by the constraints in other parts of the receiver (e.g. the update rates of the tracking loops), and the \( C/N_0 \) ratio is an external constraint imposed by the system and the environment, the designer can choose to increase or decrease the probability of error by choosing the minimum number of bit sign changes that need to be observed during synchronization before deciding the most probable bit alignment of the signal.

Using \( F \) as a design variable imposes a compromise between the error rate and the time it takes to perform synchronization. GLONASS signals use Manchester coding, which inserts a forced sign transition during the bit time. Thus it is safe to expect between 50 and 100 bit sign changes each second. GPS, on the other hand, encodes the bits using NRZ; because of this the expected data bit sign changes will be in the range between 0 and 50 each second.

V. Simulations

Figure 4 shows the probability of synchronization error as a function of the correlation integration length \( N \) for \( N = 2, 3, 5, 10 \) and the signal carrier-to-noise-density ratio, \( C/N_0 \). The values of probability of error calculated using (15) are also shown in the same figure for comparison. For these simulations, random 100 bps BPSK bit sequences were simulated, each one with \( F \) bit sign transitions and random initial edge alignments. The value of \( F \) was chosen to be 25, which is often taken as the expected number of bit sign transitions during each second for GPS signals (the expected number for GLONASS is 75).

It can be seen that the theoretical values predicted by (15) match the probability of error obtained during simulations for values of \( C/N_0 \) higher than 27 dB. This range includes the operating range for most outdoor GNSS receivers. Theoretical and simulated curves start to differ for values below that because, as the signal-to-noise ratio drops, the base hypothesis about the probability of incorrectly identifying the bit edge position being equal to the probability of mistakenly choosing a candidate whose alignment differs by \( \pm 1 \text{ ms} \) from the correct value stops being reliable.

VI. Conclusion

This paper introduced a new maximum likelihood bit synchronization algorithm that can be tailored to be used with different types of GNSS signals (GPS, GLONASS). A preliminary version of this work was presented in a previous paper by the same authors. In that paper, it was shown that this method allows to perform bit synchronization using 3 ms correlation interval samples, while at the same time it improves the probability of error compared to the histogram method.

The present paper expands our previous work by exploring the use of different correlation lengths \( N \). As a test case, we work on the bit synchronization of a 100 bps NRZ signal, since that is the kind of synchronization that is used during the first stage of bit edge disambiguation phase of a dual-system GPS/GLONASS receiver being developed in the Facultad de Ingeniería of the Universidad Nacional de La Plata.

An analytical expression to calculate the probability of synchronization errors is also provided in this paper. This expression relates the probability of error to the different parameters of the algorithm, providing the designer with valuable information during the development and testing stage of a receiver. The expression was verified using simulations, and proved to be accurate in typical use scenarios.

It is important to state that the algorithm itself and the results herein presented are in no way limited to the kind of codification and data rate presented in this paper. If a designer found more convenient to use specialized algorithms for each type of signal (50 bps NRZ GPS and 50 bps Manchester coded GLONASS signal) on a dual-system receiver, or in the design of an only GPS or only GLONASS receiver, the results provided in this paper could be reused with only slight modifications, e.g., using larger
state machines (40 states for 50 bps signals) and specialized state machine transitions for each GPS and GLONASS. Such was the approach in our previous work [11].

REFERENCES


Turbo Decoder VLSI Architecture with Non-Recursive max\(^*\) Operator for 3GPP LTE Standard

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Abstract—This paper presents a highly parallel turbo decoder architecture for 3GPP LTE standard with a new non-recursive max\(^*\) operator. High parallelism is introduced at several levels to achieve high throughput, to meet LTE requirements. The decoder supports all codes specified by LTE and features low complexity, obtained by using the new non-recursive max\(^*\) operator. The decoder achieves a maximum throughput of 376.4 Mbps at 250 MHz, occupying an area of 1.62 mm\(^2\) on 90-nm Standard Cell ASIC technology. The decoder shows better decoding efficiency (Bits/Cycle/Iterations) and throughput to area ratio (Throughput/mm\(^2\)) than many of the previously implemented decoders.

Keywords—3GPP LTE; iterative decoding; parallel turbo decoder; VLSI architecture

I. INTRODUCTION

Turbo codes [1] have gained huge attention in the last twenty years. Indeed Turbo codes achieve Near Shannon limit capacity, excellent error correction performance, intrinsic parallelism and high coding gain which made them an eligible candidate for a large number of wireless communication standards e.g., WiMax [2], 3GPP LTE [3], UMTS-LTE [4], CCSDS [5]. LTE (Long Term Evolution) is the next step forward in cellular 3G and 4G services. It is designed to meet carrier needs for high speed data and media transport as well as high capacity voice support. The standard specifies data rates up to 100 Mbps in the down-link (two MIMO channels), 50 Mbps in the up-link (single channel) and 20 MHz bandwidth channel.

VLSI implementation of turbo decoders is a challenging task because of high throughput constraints of the standards. A lot of research has been carried out to implement efficient decoders satisfying area, speed and power constraints [6][7][8]. Typically two SISO (Soft Input Soft Output) processors are concatenated in parallel in order to achieve high throughput. Parallelism can be introduced in the decoding process at several levels to improve decoding speed [9] e.g., using multiple processors, exploiting the trellis representation parallelism to decode a set of bits, internal parallelization by using multiple windows in each processing element [10], which eventually reduces the latency. However, highly parallel architectures introduce high complexity in terms of hardware and logic. In the proposed decoder architecture, several of previously mentioned methods have been adopted to improve throughput, along with a novel implementation of non-recursive max\(^*\) operator [11]. In the non-recursive max\(^*\) implementation the complexity and overhead of the operator is reduced by computing the n-input max\(^*\) as a whole instead of recursively applying the 2-input max\(^*\) operation. In the low complexity max\(^*\) operator, the two maximums are calculated and using the two outputs, the final maximum is found.

The remainder of the paper is organized as follows. Section II describes the code adopted in LTE. In this section the explanation of BCJR algorithm, that is the core of turbo decoding, is given. Section III details the proposed architecture including the overview of the main modules, focusing on the novelty in the architecture i.e., the non-recursive max\(^*\). The implementation results along with the comparisons are shown in Section IV. Finally, the whole architecture is concluded in Section V.

II. LTE TURBO CODES AND DECODING ALGORITHM

The 3GPP LTE turbo codes are based on the parallel concatenation of two 8-state Convolutional Codes (CCs) and one Quadratic Polynomial Permutation (QPP) interleaver [12]. The constituent code used in 3GPP LTE is a single binary systematic CC. The generated input frame is fed into the convolutional encoder of rate 1/3. An information sequence \(u\) is encoded by a convolutional encoder, and an interleaved version of \(u\) is encoded by a second convolutional encoder. The initial values of the shift registers of the 8-state constituent encoders are all zeros. After encoding of the current frame, termination is performed: during the last three cycles of the frame, values are input into the encoder to force it to the zero state. The total number of output bits generated by the encoder is \(L = 3 \cdot K + 12\), where \(40 \leq K \leq 644\) is the frame length. The last 12 bits are the trellis termination bits.

A. Decoder

The main processing elements inside each decoder are the Soft-In–Soft-Out (SISO) modules [13] which executes the Bahl–Cocke–Jelinek–Raviv (BCJR) algorithm [14], usually in its logarithmic form [15]. The SISO module processes the intrinsic log-likelihood ratios (LLRs), received from the channel, of a coded symbol \(c\) and calculates the extrinsic information for the information bits \(u\). The LLRs are exchanged by the SISO modules through the interleaving/deinterleaving permutation laws \(\Pi /\Pi^{-1}\). The output extrinsic LLR of symbol \(u\) at the \(k\)th step is computed as
\[
\lambda_{ext}^k = \delta \left( \max_{R_1} \{b\} - \max_{R_0} \{b\} - \lambda_{apr}^k \right) 
\]

where \( \delta = 0.75 \) is a scaling factor [16]. The trellis transition for input equal to ‘0’ is represented by \( R_0 \), whereas the trellis transition due to input equal to ‘1’ is represented by \( R_1 \). The \( \max \) finds transitions with maximum probability w.r.t. input equal to \( x \), where \( x \in \{1, 0\} \). \( \lambda_{apr}^k \) is the a-priori information received from the other SISO module, whereas \( \lambda_{ext}^k \) is the extrinsic information generated by the SISO at trellis step \( k \) and it is transferred to the other SISO module by means of the interleaver memory. \( b \) is defined as,

\[
b(k) = \alpha_{k-1}(s) + \gamma_k(s, s') + \beta_{k-1}(s') 
\]

(2)

\[
\alpha_k(s') = \max_s \left( \alpha_{k-1}(s) + \gamma_k(s, s') \right) 
\]

(3)

\[
\beta_{k-1}(s) = \max_s \left( \beta_k(s) + \gamma_k(s, s') \right) 
\]

(4)

\[
\gamma_k(s, s') = y_k + \lambda_{apr}^k + y_p^k 
\]

(5)

where \( \alpha \) and \( \beta \) are known as forward and backward state metrics and are calculated during forward and backward recursions, while \( \gamma \) is the branch metric and is calculated for each transition. SISO takes the following steps to calculate extrinsics and the output bits:

1) It calculates \( \gamma \) for the trellis section \( k \) using the systematic and parity channel LLRs and the a-priori information coming from the other SISO, as in (5).

2) From the calculated \( \gamma \), it computes \( \alpha \) by adding the previous \( \alpha \) to the corresponding \( \gamma \). So (3) can be expanded as,

\[
\alpha_k = \max_s \left( \alpha_{k-1}^1 + \gamma_k^1, \alpha_{k-1}^2 + \gamma_k^2 \right) 
\]

(6)

3) \( \beta \) is calculated in the same way like \( \alpha \), but in this case the recursion is in reverse direction (see (4)).

4) After calculating \( \alpha \), \( \beta \) and \( \gamma \), the final step is to calculate the extrinsic information and the decoded bit value, as in (1) and (2). The calculated extrinsics are passed to the other SISO by means of interleaver memory, while the decoded bit is stored in the decoded bits memory after a certain number of iterations.

III. PROPOSED ARCHITECTURE

The current 3GPP LTE standard features native code rate of 1/3, while higher code rates can also be achieved through external rate matching. All the block sizes are even numbers and are divisible by both 4 and 8, besides all the block sizes greater than 512, 1024 and 2048 are also divisible by 16, 32 and 64 respectively [17]. The complete architecture is shown in Fig. 1.

The major modules of the decoder are: (i) the 16 SISO processors, which execute the BCJR algorithm, (ii) the 16 memories storing the extrinsic values, (iii) two 16x16 switching
networks for data transfer from processors to memories and from memories to processors, (iv) the address generator which generates addresses using (7) and (v) the control unit. 3GPP LTE standard uses QPP (Quadratic Permutation Polynomial) addresses and the minimum of these 16 address is calculated through MIN-16 block, which is the effective address for each extrinsic memory. During the forward recursion, all generated addresses are also stored in the AddrBuffer, which is a LIFO of W height, where W is the maximum window length. After forward recursion through one complete window, the backward recursion is started and also on the same time the SISO produces the extrinsics, which are passed to the memories through the switching matrix. The address for the write operation (in the extrinsic memories) is always taken from the AddrBuffer.

α-init and β-init modules calculate and hold the initial values of α and β. The encoder always starts from all-zero state, which implies that α-init is a pre-computed value, while for β, three tail bits are used with a backward recursion to calculate the value of β at N – 1 state.

The batcher sorters [19] are used to manage the switching data from SISO to memory and from memory to SISO [20]. The switching matrix takes the data, coming from a memory bank, and the address, generated by the corresponding Addr generator, and it sorts the data w.r.t. the addresses. Two switching matrix are used in order to have parallel read and write operations.

There are 16 SISO processors, where each processor passes the α and β to the neighboring SISO, as suggested in [6]. The β-in for the first, second, fourth and eights is multiplexed with either the β-out of the neighboring SISO or from the β-init module. The SISO processors will be discussed in detail in the next section.

The control unit enables the decoder to work for different code lengths. Each code length is associated to a parallelism factor and f1 and f2 are given in the standard and their value depends on the block size N [17]. The decoder is designed to support the maximum code length i.e., N=6144. The maximum depth of the extrinsic memories is N/P, where N=6144 and P=16. On each clock cycle the address generators output 16 addresses and the minimum of these 16 address is calculated through MIN-16 block, which is the effective address for each extrinsic memory. During the forward recursion, all generated addresses are also stored in the AddrBuffer, which is a LIFO of W height, where W is the maximum window length. After forward recursion through one complete window, the backward recursion is started and also on the same time the SISO produces the extrinsics, which are passed to the memories through the switching matrix. The address for the write operation (in the extrinsic memories) is always taken from the AddrBuffer.

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The control unit enables the decoder to work for different code lengths. Each code length is associated to a parallelism degree that can be 1,2,4,8 and 16. For example, it enables only 1 SISO processor and only one address generator when the smallest block size is selected. Similarly, the select signals for
the multiplexers at the inputs of \( \alpha \) and \( \beta \) processors are sent by the control signal, in order to pass the correct \( \alpha' \)'s and \( \beta' \)'s to the processors at the end of one \( N/P \) bits and \( N/(P/NW) \) bits respectively. Similarly, the control unit is adaptive enough to generate control signal for the data path in order to have correct operation scheduling.

A. SISO processor

The SISO, shown in Fig. 2, is the main processing unit in turbo decoding. In this work, 16 parallel SISO processors are employed, each of which implements (1) to (5) to calculate the extrinsic values and the output bits. The complete architecture of the SISO unit is shown in Fig. 2. Each SISO gets channel LLRs i.e., systematic and parity, the apriori information from memories and the starting state metrics for \( \alpha \) and \( \beta \) recursions from other SISOs. The incoming frame is divided in \( P \) input buffers, where \( P \) is the number of SISOs. Each SISO decodes the corresponding input bits. So each SISO works on \( \frac{N}{P} \) bits, where \( N \) is the block size. According to Fig. 2, the SISO performs the following decoding operations:

1) The \( \alpha \)-BMU (Branch Metric Unit) calculates the \( \gamma' \) (branch metrics) combining the LLRs and the apriori information using (5). At the same time, the BMU-MEM stores all the incoming LLRs and the intrinsic information. BMU-MEM is a LIFO memory. The architecture of the BMU is shown in Fig. 2b.

2) After each calculation of \( \alpha \)-BMU, the \( \alpha \)-processor calculates the \( 8 \) new \( \alpha' \)s using previous \( \alpha \) and the \( \gamma \) coming from BMU. Eight processing elements (PE-0 ... PE-7) execute in parallel following (3). The new \( \alpha' \)s are normalized subtracting the maximum from each of them. Finally the calculated \( \alpha' \)s are stored in the \( \alpha \)-MEM, which is a LIFO memory as well. The architecture of \( \alpha \)-processor is shown in Fig. 2c.

3) When the \( \alpha' \)s for a complete window are calculated, the \( \beta \)-BMU and the \( \beta \)-processor start calculating the the \( \gamma' \) and \( \beta' \)s, respectively, in the backward direction using (5) and (4).

4) The \( \lambda \)-processor executes in parallel with the \( \beta \)-processor. So after each calculation of a \( \beta \), the \( \lambda \)-processor calculates the extrinsic, using (1), and also estimates the output bit. The architecture is shown in Fig. 2d.

Each SISO uses three different memories i.e., \( \alpha \)-MEM, BMU-MEM and the \( \beta \)-MEM. The \( \alpha \)-MEM stores the \( \alpha' \)s while traversing the trellis in the forward direction and the BMU-MEM stores the incoming intrinsic and the a-priori LLRs. In Table I, the number of bits required to represent each data managed by the decoder is shown. Table II shows the total memory utilization in the architecture.

<table>
<thead>
<tr>
<th>Channel LLRs ( \lambda' ,\lambda'' )</th>
<th>6</th>
</tr>
</thead>
<tbody>
<tr>
<td>Ext./Intrinsics ( \lambda )</td>
<td>8</td>
</tr>
<tr>
<td>Branch metrics ( \gamma )</td>
<td>12</td>
</tr>
<tr>
<td>State Metrics ( \alpha \beta )</td>
<td>12</td>
</tr>
</tbody>
</table>

Table II. MEMORIES USED IN THE ARCHITECTURE

<table>
<thead>
<tr>
<th>Unit</th>
<th>Size (kbit)</th>
</tr>
</thead>
<tbody>
<tr>
<td>( \alpha )-MEM</td>
<td>36.864</td>
</tr>
<tr>
<td>BMU-MEM</td>
<td>9.216</td>
</tr>
<tr>
<td>( \beta )-local-MEM</td>
<td>24.576</td>
</tr>
<tr>
<td>Addr. LIFO</td>
<td>0.216</td>
</tr>
<tr>
<td>Addr. init</td>
<td>114.56</td>
</tr>
<tr>
<td>Input Buff.</td>
<td>110.592</td>
</tr>
<tr>
<td>Output Buff.</td>
<td>6.144</td>
</tr>
<tr>
<td>Extrinsic Mem.</td>
<td>49.153</td>
</tr>
<tr>
<td>Total</td>
<td>351.32</td>
</tr>
</tbody>
</table>

B. Non-recursive max* operator

The \( \lambda \)-processor executes (1) and (2). In (1), the non-recursive max* operator [11] is used to find the maximum from all the inputs. In general the max* operator is recursive in nature, where on each recursion it finds the maximum from 2 inputs and then the result is compared with the third one and so on. In recursive max* for \( n \) input values the max* operator is applied recursively \( n-1 \) times. From (8) it is evident that the max* operator having \( n \) input values can be computed non-recursively as it requires only knowledge of the maximum among \( n \) values and an additive correction term depending on the second maximum value among \( n \) values. The non-recursive max* is implemented to find the two maximum from the 8 inputs using a simple maximum finding tree and a small look-up table (LUT), as shown in Fig. 3. Fig. 4 shows the max* implementation with 2 inputs, which is used in calculation of (3) and (4). The max* operator can be given as

\[
\max^* (x_i) \approx \max (x_i) + \log \left[ 1 + \exp (-\delta) \right] \tag{8}
\]

The log \( \left[ 1 + \exp (-\delta) \right] \) in (8) is the correction term, which is pre-computed and is stored in a small LUT. The difference between the first maximum and the second maximum, namely \( \delta \), becomes the address for the LUT. The data from the LUT is then added to the first maximum to find the max* output.

IV. RESULTS

To decrease the amount of memory and to increase the throughput, large number of windows are used for large code lengths. Each SISO processor works with up to 16 windows. In this case, each SISO is decoding 384 bits. The window is 24 bit long. So at each clock cycle, the address generators produce 16 addresses. The data from the memory and the corresponding addresses from the address generators are fed into the permutation network and finally based on the sorting of addresses, the data reaches its destinations [18].

The throughput of the decoder is calculated as in [23]:

\[
\text{Throughput} = \frac{N \times F}{2.1 \left( \frac{f}{P} + W \right)} \tag{9}
\]

where \( N \) is the number of decoded bits, \( F \) is the clock frequency at which the decoder is synthesized, \( I \) is the number of iterations, \( P \) is the number of SISO processors, and \( W \) is the window length. The BER and FER results, shown in Fig. 5 and Fig. 6 respectively, are obtained with 5 and 9 iterations for code length 6144. Stopping the decoding after 5 iterations

---

**Table I. QUANTIZATION**

<table>
<thead>
<tr>
<th>No. of bits</th>
<th>Channel LLRs ( \lambda', \lambda'' )</th>
<th>6</th>
</tr>
</thead>
<tbody>
<tr>
<td>Ext./Intrinsics ( \lambda )</td>
<td>8</td>
<td></td>
</tr>
<tr>
<td>Branch metrics ( \gamma )</td>
<td>12</td>
<td></td>
</tr>
<tr>
<td>State Metrics ( \alpha \beta )</td>
<td>12</td>
<td></td>
</tr>
</tbody>
</table>
### Table III. RESULTS COMPARISONS: Radix/Processor (Rdx/Proc.), CMOS Technology Process (Tech.), Iterations (Iter.), Clock Frequency (Freq.), Memory (Mem), Active Area (Area), Normalized Area (N.A.), Throughput (TP), Throughput to Area Ratio (TAR = Mbps/mm), Decoding Efficiency (DE = Bits/Cycle/Iterations)

<table>
<thead>
<tr>
<th>Design</th>
<th>Proposed</th>
<th>[6]</th>
<th>[7]</th>
<th>[8]</th>
<th>[17]</th>
<th>[20]</th>
<th>[18]</th>
<th>[21]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Standard</td>
<td>LTE</td>
<td>LTE, WiMAX</td>
<td>LTE</td>
<td>LTE</td>
<td>LTE</td>
<td>LTE</td>
<td>LTE</td>
<td>LTE</td>
</tr>
<tr>
<td>Rdx./Proc.</td>
<td>2/16</td>
<td>4/8</td>
<td>2/8</td>
<td>2/8</td>
<td>2/64</td>
<td>4/8</td>
<td>2/16</td>
<td>2/32</td>
</tr>
<tr>
<td>Tech (nm)</td>
<td>90</td>
<td>130</td>
<td>90</td>
<td>90</td>
<td>65</td>
<td>90</td>
<td>90</td>
<td>65</td>
</tr>
<tr>
<td>Iter.</td>
<td>5</td>
<td>8</td>
<td>8</td>
<td>8</td>
<td>6</td>
<td>5.5</td>
<td>5</td>
<td>5.5</td>
</tr>
<tr>
<td>Freq.</td>
<td>250</td>
<td>250</td>
<td>275</td>
<td>275</td>
<td>400</td>
<td>400</td>
<td>400</td>
<td>400</td>
</tr>
<tr>
<td>Mem (Kb)</td>
<td>351.32</td>
<td>N/A</td>
<td>N/A</td>
<td>N/A</td>
<td>N/A</td>
<td>129</td>
<td>413.35</td>
<td>N/A</td>
</tr>
<tr>
<td>Area (mm²)</td>
<td>1.62</td>
<td>10.7</td>
<td>2.10</td>
<td>2.10</td>
<td>8.3</td>
<td>3.57</td>
<td>2.10</td>
<td>13.82</td>
</tr>
<tr>
<td>N.A. (mm²)</td>
<td>1.62</td>
<td>5.12</td>
<td>2.10</td>
<td>2.10</td>
<td>15.91</td>
<td>1.71</td>
<td>2.10</td>
<td>13.82</td>
</tr>
<tr>
<td>TP (Mbps)</td>
<td>376.47</td>
<td>187.53</td>
<td>130</td>
<td>123</td>
<td>3140.72</td>
<td>298.6</td>
<td>254.4</td>
<td>13.82</td>
</tr>
<tr>
<td>TAR</td>
<td>232.39</td>
<td>36.62</td>
<td>61.90</td>
<td>61.42</td>
<td>82.83</td>
<td>228.42</td>
<td>135.43</td>
<td>82.34</td>
</tr>
<tr>
<td>DE</td>
<td>7.53</td>
<td>6</td>
<td>3.78</td>
<td>3.75</td>
<td>19.66</td>
<td>7.16</td>
<td>7.11</td>
<td>12.88</td>
</tr>
</tbody>
</table>

Instead of 9 introduces a penalty of about 0.5 dB at BER level $10^{-4}$, but offers increased throughput and reduced energy dissipation. It is also shown in Fig. 5 and 6 that the non-recursive max* achieves slight gain over the recursive one.

The synthesis is carried out on Synopsys Design Vision with 90 nm standard cell technology at a clock frequency of 250 MHz. The proposed architecture achieves a maximum throughput of 376.4 Mbps, occupying an area of 1.62 mm².

Table III shows the comparison of the proposed architecture with some already published decoders. The comparison is carried out in terms of implementation technology, supported codes and decoded modes, maximum achieved throughput, internal parallelism and occupied area. The area of each implementation is scaled up to 90 nm process technology for fair comparison. The scaling factors of $(90/70)^2$ and $(90/65)^2$ are used for 130 nm and 65 nm technologies, respectively. A parameter called Throughput to Area Ratio (TAR) [24] is used to evaluate the area efficiency of the designed decoder. Another metric used for comparison purposes is Decoding Efficiency (DE), defined as the number of decoded bits per clock cycle per iteration. DE evaluates the degree of parallelism actually offered by the decoder architecture. As shown in Table III, the proposed architecture achieves better TAR than the other implementations. The DE is better than [6], [7], [8], [20], [18] and [22], whereas [17] and [21] achieve better DE, but they are very expensive architectures, with huge area occupation. So the proposed architecture achieves more than sufficient throughput for 3GPP LTE standard at the cost of lower occupied area than required for most of the alternate solutions.

**V. Conclusion**

In this paper, an already developed parallel turbo decoder for 3GPP LTE standard is modified with a new architecture of non-recursive max* operator. The analysis shows that with the new max* operator, significant area can be saved and throughput enough to meet the standard requirement can be achieved. The new max* operator is implemented with a very simple logic i.e., by finding just the two maximums and then adding them with a correction term, taken from a LUT.

**REFERENCES**


A Novel Range-spread Target Detection Algorithm Based on Waveform Entropy for Missile-borne Radar

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Abstract—A novel range-spread target detection algorithm in white Gaussian clutter for missile-borne radars is presented. For missile-borne radars, range migration of the target echoes during a coherent processing interval (CPI) is serious, which disperses the echo energy of the target and increases the difficulty of target detection. In this paper, range alignment for target echoes is accomplished in the frequency domain during the process of digital pulse compression. Then a range-spread detection algorithm based on waveform entropy (WE) of the average combination of high range resolution profiles (HRRP) is addressed, which has the property of constant false-alarm rate (CFAR). Finally, the detection performance is assessed by Monte-Carlo simulation, and the results indicate that the detection performance of the proposed detector is superior to the traditional energy integrator detector and is robust for different HRRPs of the target.

Keywords—range resolution; range-spread; waveform entropy; target detection; CFAR.

I. INTRODUCTION

High range resolution (HRR) radars are widely used in precision guidance in recent years [1], for the echoes of HRR radars involve abundant target information and can be used for target recognition and accurate tracking [2]. For HRR radars, a relatively large target can be assumed to be the composition of multiple physical scatterers which distribute in different range cells in radar echo, so is called a range-spread target [3]. Consequently, traditional point-like target detection schemes for low range resolution (LRR) radars may fail for HRR radars [4]. Many achievements have been made in range-spread target detection during the past decades [5-9] (and references therein). Precisely, range-spread targets detection in white Gaussian noise of a known spectral density level is addressed in [5-6]. In [7], constant false-alarm rate (CFAR) detectors based on a generalized likelihood ratio test (GLRT) for range-spread targets are derived. Adaptive detection of distributed targets has been made in range-spread target detection in [8], with reference to Gaussian disturbance clutter. In [9], CFAR detection of distributed targets in non-Gaussian disturbance modelled as a compound-Gaussian process is studied. Nevertheless, the detection algorithms above do not consider the relative motion between the radar and the target, so they are not applicable in the scenario of moving range-spread target detection. In addition, some of them consume so much computation that they are difficult to implement in engineering.

In this paper, we propose a novel range-spread target detection algorithm for anti-ship terminal guidance HRR radars. Firstly, the range-spread target echo model of a missile-borne HRR chirp radar is established and the range alignment of HRRP is accomplished in the frequency domain. Secondly, the range-spread target detection method based on the waveform (WE) of the radar echo after coherent integration is addressed. Finally, the simulation indicates that the detection algorithm is superior to the traditional energy integrator detector and is robust for different HRRPs of the target. It should be pointed out that the proposed detector based on waveform entropy needs not to estimate the parameters of the clutter, which is a must for many traditional detection algorithms. Moreover, the computational complexity of the detector makes it suitable for missile-borne radar signal processing.

The paper is organized as follows. In Section II, we build and analyze the echo model of a range-spread target for missile-borne HRR radar. In Section III, range alignment of the target echoes is accomplished in the frequency domain. We address the range-spread target detection algorithm in Section IV. And the performance of the proposed method is assessed in Section V. At last, in Section VI, some conclusions are given.

II. RADAR ECHO MODEL

The transmitted signal of a chirp radar is expressed as:

\[
s(i,t) = \operatorname{rect}(\frac{i}{T_p}) \cdot \exp(j\mu t^2) \cdot \exp(j2\pi f_c t).
\]

(1)

where \( i \) represents fast time; \( t_i = mT_s \) is slow time, \( m=0,1,\ldots,M-1 \), \( M \) is the number of pulses for coherent integration; \( T_s \) is the pulse repetition interval; \( t = \hat{t} + t_i \) is the absolute time; \( \mu = B/T_p \) is the frequency slope of chirp pulse, \( B \) is the signal’s bandwidth, \( T_p \) is the pulse width; \( f_c \) is the carrier frequency.

According to the radar theory, the echo of a target is the convolution of the transmitted signal with the target range-scattering function. The range-scattering function of a static point-like target at the range of \( R_0 \) can be written as [3]:

\[
CF_0(t) = a_0 \cdot e^{j2\phi_0} \cdot \delta(t - 2R_0/c).
\]

(2)

where \( a_0 \) is the amplitude, \( \phi_0 \) is the initial phase, and \( c \) is the velocity of light. Thus the range-scattering function of a static range-spread target can be expressed as:

\[
CF(t) = \sum_{k=1}^{K} a_k \cdot e^{j2\phi_k} \cdot \delta(t - \tau_k).
\]

(3)

where \( K \) is the number of scattering centers of the range-spread target, \( a_k \), \( \phi_k \) and \( \tau_k \) are the amplitude, initial phase and delay of the \( k \)-th physical scatterer of the range-spread target in range cell, respectively.
Assuming that the radar is working on tracking condition and moving towards the target at sea, and the radial velocity between the radar and the target remains constant during a coherent processing interval (CPI). Accordingly, the instantaneous range between the radar and the target is:

$$R_i(t_c) = R_{0i} - vt_i, \quad k = 1, 2, ..., K.$$  \hspace{1cm} (4)

where $R_i(\cdot)$ represents the instantaneous range of the $k$-th scatterer of the range-spread target, $R_{0i}$ is the initial range of the $k$-th scatterer and $v$ is the radial velocity between the radar and the target. Substituting (4) into (3), the instantaneous range-scattering function of a range-spread target is obtained:

$$CF(t, t_c) = \sum_{k=1}^{K} a_k \cdot e^{i2\pi f_{k0}} \cdot \delta \left( t - \frac{2R_i(t_c)}{c} \right)$$

$$= \sum_{k=1}^{K} a_k \cdot e^{i2\pi f_{k0}} \cdot \delta \left( t - \frac{2(R_{0i} - vt_i)}{c} \right).$$  \hspace{1cm} (5)

Therefore, the radar echo of the range-spread target is expressed as:

$$r(t, t_c) = \sum_{k=1}^{K} a_k \cdot e^{i2\pi f_{k0}} \cdot \text{rect} \left( \frac{t - 2(R_{0i} - vt_i)}{T_p} \right)$$

$$\cdot \exp \left\{ j\pi f_{k0} \left[ t - 2(R_{0i} - vt_i)/c \right] \right\}$$

$$\cdot \exp \left\{ j2\pi f_k \left[ t - 2(R_{0i} - vt_i)/c \right]/c \right\}.$$  \hspace{1cm} (6)

where $f_{k0} = 2\nu/c$ is the Doppler frequency. After mixing and low-pass filtering, the baseband target echo is written as:

$$r_{\text{baseband}}(\hat{t}, t_c) = \sum_{k=1}^{K} a_k \cdot e^{i2\pi f_{k0}} \cdot \text{rect} \left( \frac{\hat{t} - 2(R_{0i} - vt_i)/c}{T_p} \right)$$

$$\cdot \exp \left\{ j\pi f_{k0} \left[ \hat{t} - 2(R_{0i} - vt_i)/c \right]^2 + j2\pi f_k \hat{t}/c \right\}$$

$$\cdot \exp \left\{ -j4\pi f_k (R_{0i} - vt_i)/c \right\}.$$  \hspace{1cm} (7)

From (7), it can be seen that the relative radial velocity between the radar and the target produces time-frequency coupling to the chirp signal [10], which results in mismatching between the target echo and the matched filter. After matched filtering, HRRPs of the range-spread target can be written as:

$$y(\hat{t}, t_c) = \sum_{k=1}^{K} a_k \sqrt{\mu T_p} \cdot \text{rect} \left( \frac{\hat{t} - 2(R_{0i} - vt_i)/c}{T_p} \right)$$

$$\cdot \sin \left[ \pi B \hat{t} \frac{2(R_{0i} - vt_i)/c}{T_p} \frac{f_k}{\mu} \right]$$

$$\cdot \exp \left\{ j2\pi (\phi_0 + \frac{1}{8}) \right\}$$

$$\cdot \exp \left\{ j2\pi f_k (R_{0i} - vt_i)/c \right\}.$$  \hspace{1cm} (8)

Due to mismatching, envelopes of the pulse compression results in (8) are not sinc functions any more. According to [10], the amplitude loss in (8) is negligible, while the time shift resulting from the time-frequency coupling of chirp signal is significant and will affect range measurement accuracy. From (8), the time shift of the target resulting from the time-frequency coupling of the chirp signal is:

$$\Delta t = f_k/T_p.$$  \hspace{1cm} (9)

Additionally, the radial velocity $v$ between the radar and the target produces range migration during a CPI. From (8), the migration time between the adjacent pulses is expressed as:

$$t_m = 2vT_p/c.$$  \hspace{1cm} (10)

The range migration factor $P$ is defined as:

$$P = B(M - 1)\nu = 2Bv(M - 1)T_p/c.$$  \hspace{1cm} (11)

The range migration factor $P$ represents the number of the range resolution cells that the target echo spreads during the CPI.

Coherent integration is widely adopted to enhance the signal-to-noise ratio (SNR) of radar echo [11], which improves the target detection performance of the radar system. But when there is high speed relative motion between the radar and the target, the range migration will affect the result of coherent integration. Fig. 1 shows the range migration effects on coherent integration results. Simulation parameters: $B = 80MHz$, $M = 32$, $T_p = 20\mu s$, $T_c = 250\mu s$, sampling rate $f_s = 120MHz$.

Coherent integration is widely adopted to enhance the signal-to-noise ratio (SNR) of radar echo [11], which improves the target detection performance of the radar system. But when there is high speed relative motion between the radar and the target, the range migration will affect the result of coherent integration. Fig. 1 shows the range migration effects on coherent integration results of a point-like target with range migration factor $P = 0, 2, 4$ and $8$, respectively. It can be seen that the range migration disperses the echo energy of the target in coherent integration and so makes it more difficult to target detection.

Assuming it is required that the range migration during the CPI should be no more than half of the range resolution cell [12], according to (11), the radial velocity between the radar and the target should satisfy the following equation:

$$v < \frac{c}{4B(M - 1)T_p}.$$  \hspace{1cm} (12)

III. RANGE ALIGNMENT

For missile-borne radar, equation (12) usually can not be satisfied. Therefore it is necessary to correct the range migration between the HRRPs before coherent integration. Supposing the velocity of the missile $\nu_m$ can be obtained by the missile-borne inertial navigation system (INS) [13], the velocity measurement error is $\Delta \nu_m$. If we adopt $\nu_m$ as an estimation of the radial velocity between the radar and the target, the estimation error is:

$$\varepsilon = \Delta \nu_m + \nu_r.$$  \hspace{1cm} (13)
where \( v_r \) is the radial velocity of the target ship. For the time being, the radial velocity estimation error \( \varepsilon \) generally satisfies (12) and so \( v_r \) can be employed to correct the range migration between the HRRPs during a CPI.

Supposing the velocity of the missile \( v_m \) provided by the INS is 2000 m/s, the real radial velocity between the missile and the target \( v = 20 \) m/s, the error of the coarse estimation of the radial velocity \( \varepsilon = 20 \) m/s. The echoes of a range-spread target with four scattering centers are shown in Fig. 3, the horizontal is the range axis, and the ordinate is the Doppler frequency axis, where Fig. 3 (a) shows the echoes without range alignment, Fig. 3 (b) shows the echoes after range alignment. Fig. 4 shows the coherent integration results of the target echoes before and after range alignment. It can be seen, compared with the coherent integration result without range alignment, the echo energy of the target has been effectively gathered after range alignment.

IV. TARGET DETECTION

After Range alignment and coherent integration, the echo energy of the range-spread target is accumulated effectively. Denoting the data after coherent integration by matrix \( Z \{ M \times N \} \), where \( N \) is the maximum of the range cells number that the desired target spreads. The detection problem can be formulated in terms of the following binary hypotheses test:

\[
\begin{align*}
H_0 : z_m = w_m, & \quad m = 0, 1, \ldots, M - 1, \\
H_1 : z_m = x_m + w_m, & \quad m = 0, 1, \ldots, M - 1.
\end{align*}
\]

where \( w_m \) are the received signal vector, sea clutter vector, and HRRP of the desired target, respectively. All of them are row vectors with length \( N \), and \( w_m \) is assumed to be zero-mean complex white Gaussian noise.

In practical applications, to restrain the side lobe of the Doppler [14], a weighting function is generally adopted during coherent integration, which expands the Doppler of the target echo. Moreover, the target ship may corner sometimes, which expands the Doppler as well. Nevertheless, the maximum number of Doppler cells that the target echo spreads in Doppler axis usually can be known in advance. Setting a sliding-window of width \( M \) on the Doppler axis, where \( M' \) is the max number of the Doppler cell that the target echo may take up, \( M' \leq M \), and \( r_d \) is the Doppler resolution cell of the radar system. The echo energy within the window is integrated, whichever makes the integrated value of the largest window is chosen as the target detection window, denoted by \( W \). The detection problem can be re-described as the following binary hypothesis test:

\[
\begin{align*}
H_0 : z_m(n) = w_m(n), & \quad m = m_0, m_0 + 1, \ldots, m_0 + M_0 - 1, n = 0, 1, \ldots, N - 1, \\
H_1 : z_m(n) = x_m(n) + w_m(n), & \quad m = m_0, m_0 + 1, \ldots, m_0 + M_0 - 1, n = 0, 1, \ldots, N - 1.
\end{align*}
\]

where \( m_0 \) is the starting Doppler cell number of the target in \( W \). Envelops of the \( M_0 \) HRRPs are high correlated, which can be utilised to construct the detector. An average combination of the radar echo within the detection window \( W \) is defined by:

\[
A(n) = \frac{1}{M_0} \sum_{m=0}^{m_0+M_0-1} z_m^2(n), n = 0, 1, \ldots, N - 1.
\]

In fact, the average combination defined in the above equation can be seen as incoherent integration of the \( M_0 \) echoes. Because of the independence of the clutter for
The waveform entropy has the following properties:

(WE) \[15\]. Entropy is a measure of the uncertainty of random variables \([16]\), in order to adopt the concept of entropy, setting:

\[
p(n) = |A(n)| / ||A||, \quad n = 0, 1, ..., N - 1.
\]

Therefore, from (23), the hypothesis is also independent of 2

\[
\text{H}_0 \implies \text{H}_1.
\]

(16), the detection statistic is

\[
R = WE[A(n)] = \frac{1}{M_0} \sum_{n=0}^{M_0-1} z_m^2(n) \quad H_0 \implies T.
\]

where \(T\) is the detection threshold, \(n = 0, 1, ..., N - 1\).

Under the \(H_0\) hypothesis in (16), the detection statistic is written as:

\[
R = WE \left[ \frac{1}{M_0} \sum_{n=0}^{M_0-1} w_m^2(n) \right], \quad n = 0, 1, ..., N - 1.
\]

Setting:

\[
q(n) = \frac{m_0-M_0}{m_0-N} \sum_{n=0}^{M_0-1} w_m^2(n) - \frac{n}{N} \sum_{n=0}^{N-1} w_m^2(n), \quad n = 0, 1, ..., N - 1.
\]

Then, the detection statistic under the \(H_0\) hypothesis is expressed as:

\[
R = -\sum_{n=0}^{N-1} q(n) \cdot \log_2 q(n).
\]

Here, we assume the variance of \(w_m(n)\) is \(\sigma^2\). Then,

\[
\frac{1}{\sigma^2} w_m^2(n) \sim \chi^2(l)\]

where \(\chi^2(l)\) denotes a chi-square distribution with freedom \(l\).

Similarly,

\[
\frac{1}{\sigma^2} \sum_{n=0}^{M_0-1} w_m^2(n) \sim \chi^2(2M_0),
\]

and,

\[
\frac{1}{\sigma^2} \sum_{n=0}^{N-1} \sum_{m=0}^{M_0-1} w_m^2(n) - \chi^2(2M_0N).
\]

From (22), (25) and (26), we can obtain:

\[
N \cdot q(n) \sim F(2M_0, 2M_0N), \quad n = 0, 1, ..., N - 1.
\]

So the probability distribution function (PDF) of \(q(n)\) is independent of the clutter variance \(\sigma^2\). Therefore, from (23), we can find that the detection statistic \(R\) under the \(H_0\) hypothesis is also independent of \(\sigma^2\). Thus, the false alarm probability is independent of the external clutter environment, which implies that the proposed detector in (20) is a CFAR detector.

The Neyman-Pearson criterion is employed to make the judgement. Although the closed-form expression for the false-alarm probability \(p_{fa}\) is difficult to derive, the threshold for a special false-alarm probability can be obtained by the widely used Monte-Carlo method. While the detection statistic is smaller than the threshold, \(H_0\) hypothesis is selected.

V. PERFORMANCE ASSESSMENT

In this section, we assess the performance of the waveform entropy detector given by (20) resorting to Monte-Carlo simulations. The simulation parameters of the radar are listed in table 1, which are assumed according to the principle in [1].

\[
\text{Copyright (c) IARIA, 2013. ISBN: 978-1-61208-264-6}
\]
Fig. 6 shows the HRRPs of two range-spread targets simulated by the computer, where HRRPs of target 1 and target 2 are shown in Fig. 6(a) and Fig. 6(b), respectively. Each of the two targets has four scattering centers, while the distributions of the scattering centers are different.

<table>
<thead>
<tr>
<th>B</th>
<th>T_p</th>
<th>T_s</th>
<th>M</th>
<th>f_s</th>
</tr>
</thead>
<tbody>
<tr>
<td>80 MHz</td>
<td>250 μs</td>
<td>20 μs</td>
<td>32</td>
<td>120 MHz</td>
</tr>
</tbody>
</table>

Table I. Parameters of a HRR Chirp Radar

In the following we compare the performance of the proposed detector based on waveform entropy with the integrator detector [5]. Gaussian clutter is generated by computer, while the variance is adjusted to SCR. Taking the computational complexity into account, we assumed \( p_{fa} = 0.0001 \) and 1000 independent trials are carried out at each SCR.

Fig. 7 shows the detection performance of the proposed detector based on waveform entropy and the integrator detector. Simulations for target 1 and target 2 are shown in Fig. 7(a) and Fig. 7(b), respectively. It can be seen that the detection performance of the proposed detector is superior to the integrator detector.

The detection performance of the proposed detector for target 1 and target 2 with different Doppler distributions (\( M_d \)) is shown in Fig. 8. From Fig. 8, we can see that the detection performance of the proposed detection method is robust to the HRRP of the desired target, and yet is connected with the Doppler distribution of the target. In this simulation, the detection performance of the proposed detector decreases 1 dB while \( M_d \) doubles. In spite of this, the detection performance of the proposed detector is superior to the integrator detector even when \( M_d = M \).

VI. CONCLUSION AND FUTURE WORK

HRR radars are widely used in precision guidance in recent years, for the echoes of HRR radars involve abundant target information and can be used for target recognition and accurate tracking. This paper presents a novel range-spread target detection algorithm for missile-borne HRR radars. Firstly, the range-spread target echo model of missile-borne HRR chirp radar is established and the range alignment of HRRP is accomplished in the frequency domain. Then a CFAR range-spread detection algorithm based on the WE of the average combination of the HRRPs is addressed. The simulation indicates that the detection algorithm is superior to the traditional energy integrator and is robust for different
HRRPs of the target. The proposed detector based on WE needs not to estimate the parameters of the clutter, and the computational complexity of the detector makes it suitable for missile-borne radar signal processing. Future work will focus on the practical application of the proposed algorithm, and the target detection performance will be further tested by using the measured data.

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A Brief Analysis of the Motion Compensation for FMCW SAR

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Abstract—The combination of frequency modulated continuous wave (FMCW) and SAR (synthetic aperture radar), is one of the most important and dynamic research around the world. For FMCW SAR, traditional approximation of stop-go is invalid, hence the traditional motion compensation method is no longer suitable for FMCW SAR. The signal model of FMCW SAR in the presence of motion error is derived in the paper, as well as the difference introduced by inter-pulse motion errors. Then a new motion compensation method is proposed. Simulated and experimental FMCW SAR data is generated and the well-focused image shows the proposed method is available.

Keywords—Frequency Module Continuous Wave; Synthetic Aperture Radar; Motion Compensation; Line of Sight; intra-pulse; inter-pulse.

I. INTRODUCTION

Being a kind of advanced imaging sensor, SAR has been assembled with a variety of platforms, like airplane, airship, helicopter, missile, UAV (unmanned airplane vehicle), even the satellites. Most of these airborne platforms would be disturbed by the air current and are going to move on a non-ideal trajectory, accordingly, declines the quality of SAR imagery, even leads to a failure of imaging.

Airplane is a popular platform for SAR system, in the background of the glorious boom of UAVs, the combination of SAR and UAV, leads to a significant development for both the SAR sensor and UAVs. For one thing, it expands the application of SAR, for another, UAVs could have a stronger power of observation. However, the limited load of UAV could not satisfy the big, heavy and high-power pulsed SARs. Fortunately, the technology of FMCW could make the SAR smaller, low power and low-weight. It is the reason why FMCW SAR has been one of the most important and dynamic research in the field of radar around the world [1-5].

In FMCW SAR Systems, due to the fact that the transmitted wave is continuous and the length is in milliseconds, traditional approximation of stop-go in SAR theory is no longer valid. The motion during inter-pulse could not be ignored and the original method of motion compensation is not available either. But the error motion could still be separated into two terms: range error along the LOS and velocity error along track. Resampling the raw data along the track or transmitting the pseudo-continuous wave whose duty cycle is approximately to 1 then adjusting the PRF online are alternative for the elimination of velocity error. Thus the processing of eliminating velocity error could be seen as the preparation for imaging. However, it becomes complicated when dealing with the range error. A new method which is available for compensating the range error is proposed in this paper. In addition, a laboratorial FMCW SAR system is built-up, based on which the presented approach could be validated.

In Section II, the SAR system geometry and signal in presence of trajectory deviation is addressed. In Section III, difference phenomenon introduced by inter-pulse motion error is presented while the related approach is presented too. The proposed method is verified with the simulated and experimental data in Section IV. Finally, the conclusion is pointed out in Section V.

II. SIGNAL MODEL OF FMCW SAR IN NON-IDEAL CASE

Let us refer to Fig.1, in which the SAR system geometry in the presence of the 3-D motion error is presented. The real trajectory is shown as the dotted line, while the straight line along X denotes the ideal trajectory. O means the origin of the coordinates and β is the off-nadir angle. \( P(X(t), Y(t), Z(t)) \) are the azimuth, range and height position of the antenna phase center (APC), while \( \Delta X(t), \Delta Y(t), \Delta Z(t) \) are the instantaneous 3-D error. \( R(t, R_0) \) denotes the instantaneous range target-to-antenna distance and \( t \) is the time variable within time axis. It could be expressed as \( t = \hat{t} + t_s \), in which \( \hat{t} \) is the fast time and \( t_s \) is the slow time.

Assuming the static target T in the scene, whose position is \( \{X_r, Y_r, Z_r\} \), then the instantaneous range between T and APC is

\[ R(t, R_0) = \sqrt{(X_r + \Delta X(t))^2 + (Y_r + \Delta Y(t))^2 + (Z_r + \Delta Z(t))^2} \]

Figure 1. FMCW SAR system geometry in presence of trajectory deviation
\[
R(t, R_g) = \sqrt{(X(t) - X_s)^2 + (Y(t) - Y_s)^2 + (Z(t) - Z_s)^2}
\]  \(1\)

Correspondingly, \(R(t, R_g)\) would be expressed in ideal case as

\[
R(t, R_g) = \sqrt{(Vt - X_s)^2 + Y_t^2 + Z_t^2} = \sqrt{(Vt - X_s)^2 + R_g^2}
\]  \(2\)

where \(V\) means the ideal velocity, and \(R_g = \sqrt{Y_t^2 + Z_t^2}\) denotes the distance of closest approach. The expression shown in (2) reveals that the varied points at the same range have the same closest distance and they will have a similar signal form. It is known as invariance of azimuth and made the range cell migration compensation more convenient and quick. Unfortunately, the motion error destroys the invariance and convenience.

\[
A \sim |\sin(\beta t + \Delta Z(t) \cos \beta)| \ll |(X(t) - X_s)| + R_g \quad (1)
\]

(1) could be expressed after the Taylor expansion as

\[
R(t, R_g) = \sqrt{(X(t) - X_s)^2 + R_g^2 - 2\Delta Y(t)R_g \sin \beta - 2\Delta Z(t)R_g \cos \beta + \Delta^2 \theta(t) + \Delta^2 \beta(t)}
\]

\[
= \sqrt{(X(t) - X_s)^2 + R_g^2 - 2\Delta Y(t)\sin \beta - 2\Delta Z(t)R_g \cos \beta}
\]

\[
= \sqrt{(X(t) - X_s)^2 + R_g^2 - 2\Delta Y(t)(\sin \beta + \Delta Z(t) \cos \beta)}
\]

\[
= \sqrt{(X(t) - X_s)^2 + R_g^2 - \Delta v(t, R_g) \cos \theta}
\]  \(3\)

where

\[
\cos \theta = \frac{R_g}{\sqrt{(Vt - X_s)^2 + R_g^2}} \approx \frac{R_g}{\sqrt{(Vt - X_s)^2 + R_g^2}}
\]

\(\theta\) is the instantaneous squint angle between APC and scatter \(T\). The first approximation in (3) is derived from the situation that \(\max{\{\Delta Y(t), \Delta Z(t)\}} < R_g\) and the second approximation benefit from the ignorance of the second and senior order term, most importantly, these approximations are available generally.

In (3) it is denoted that the 3-D motion error has been separated into cross-track and along-track motion error. (3) Could be expressed as follows in the case that ignoring the inter-pulse motion, exactly for the pulsed SAR,

\[
R(t, R_g) = \sqrt{(X(t_n) - X_s)^2 + R_g^2 - \Delta v(t_n, R_g) \cdot \cos \theta}
\]  \(4\)

in which \(t_n\) denotes the slow time, being a sample of the whole time \(t\). \(X(t_n)\) is the azimuth location of platform for pulsed SAR. \(\Delta v(t_n, R_g) = \Delta Y(t_n) \sin \beta + \Delta Z(t_n) \cos \beta\). Intuitively, (4) could be obtained by exchanging \(t\) with \(t_n\) and the process is shown in Fig. 2.

The sampled line is suitable for pulsed SAR but unavailable for FMCW SAR due to the fact that the inter-pulse motion error could not be ignored. For FMCW SAR, assuming the period of the transmitted LFM is \(T\), which is on the order of \(10^{-3}\) s. It is reasonable to consider the velocity contains same variation during \(T\), shown as \(V(t_a + \hat{t}) = V(t_a)\). \(V(t_a + \hat{t}) = V(t_a)\) \(\hat{t} \in [0, T]\), in which \(V\) means the velocity of platform along LOS. Consequently, (3) is separated as

\[
R(t, R_g) = \sqrt{(X(t) - X_s)^2 + R_g^2 - \Delta v(t, R_g) \cos \theta}
\]

\[
\approx \sqrt{(X(t) - X_s)^2 + R_g^2 - \Delta v(t_a, R_g) \cos \theta}
\]

\[
\approx \sqrt{(X(t) - X_s)^2 + R_g^2 - \Delta v(t_a, R_g) - \xi_a \cdot \hat{t}}
\]  \(5\)

where \(\xi_a = V(t_a) \sin \beta + V(t_a) \cos \beta\), the first approximation is available in the case of \(\Delta Y(t) \approx \Delta Y(t_a) + V(t_a) \Delta t\), \(\Delta Z(t) \approx \Delta Z(t_a) + V(t_a) \Delta t\) while the second one is \(\cos \theta \approx 1\), especially satisfied for high-band SAR.

The raw echo of FMCW SAR in the ideal case is \([2-3]\)

\[
s(R, t) = w[R - \frac{2\pi(t)}{\lambda}]v[R, t] \exp(-j \frac{4\pi}{\lambda} R(t, R_g))
\]

\[
\text{exp}(-j \frac{2\pi}{\lambda} R(t, R_g)) \exp(j \frac{4\pi}{\lambda} R(t, R_g))
\]

\[
\exp(-j \xi_a \hat{t})
\]  \(6\)

(6) is approximated usually as

\[
s(R, t) = \{w[R - \frac{2\pi(t)}{\lambda}]v[R, t]\} \exp(-j \frac{4\pi}{\lambda} R(t, R_g)) \exp(-j \frac{4\pi}{\lambda} R(t, R_g))
\]

\[
\exp(-j \xi_a \hat{t})
\]  \(7\)

where \(w[R - \frac{2\pi(t)}{\lambda}]v[R, t]\) are the rectangle envelopes of echo in range and azimuth dimension respectively (ignoring the inference of antenna’s weight). Substituting (5) into (6), the signal model of FMCW SAR which is in the presence of motion errors is derived. Since the range error has a negligible effects on the envelop, same to the inter-pulse motion error on RVP term, the finally echo is
\[ s(t,\tau_c;\varepsilon) = w(t) \frac{2R(t)}{c} v(\tau_c) \exp\left(-\frac{4\pi}{\lambda} R(t, R_0)\right) \exp\left(-\frac{4\pi}{\lambda} \tau_c \varepsilon \cdot i\right) \]

\[ \exp\left(\frac{4\pi k R^2(t, R_0)}{c^2}\right) \exp\left(\frac{8\pi k R^2(t, R_0)}{c^2}\right) \exp\left(\frac{16\pi k R^2(t, R_0)}{c^2}\right) \]

*(where)*

\[ R(t, R_0) = \sqrt{(X(t) - X_0)^2 + R_0^2} \]

\[ R(t, R_0) = \sqrt{(X(t) - X_0)^2 + R_0^2} \]

After a simplification and approximation, (9) changes to

\[ s(t,\tau_c) = [w(t) \frac{2R(t)}{c} v(\tau_c) \exp\left(-\frac{4\pi}{\lambda} R(t, R_0)\right) \exp\left(-\frac{4\pi k R^2(t, R_0)}{c^2}\right) \exp\left(-\frac{16\pi k R^2(t, R_0)}{c^2}\right)] \]

\[ \exp\left(\frac{4\pi k R^2(t, R_0)}{c^2}\right) \exp\left(\frac{8\pi k R^2(t, R_0)}{c^2}\right) \exp\left(\frac{16\pi k R^2(t, R_0)}{c^2}\right) \]

The convolution term outside the bracket represents the RVP and envelop oblique terms, introduced by the de-chirp processing, which shall be eliminated in the process of imaging. Referring to (10), it is known that the errors due to range error along LOS is

\[ H_{\text{los}} = \exp\left(\frac{4\pi k R^2(t, R_0)}{c^2}\right) \exp\left(\frac{8\pi k R^2(t, R_0)}{c^2}\right) \exp\left(\frac{16\pi k R^2(t, R_0)}{c^2}\right) \]

Where \( \beta \) is the off-nadir angle according to the center of scene.

As for the real-time processing of SAR, the approach of motion compensation and imaging are merged to pursue the maximum efficiency. Due to the fact that range doppler (RD) algorithm is effective and easy to realize, an available approach of motion compensation which is based on RDA is proposed in this paper, and the flow chart is shown in Fig. 3:

The flow chart of imaging process with motion compensation is similar with pulsed SAR exception for

\[ \Delta Rcm = \frac{c}{2k} \left[ V_y(t_n) \sin \beta + V_z(t_n) \cos \beta \right] \]

Where \( \beta \) denotes the off-nadir angle according to the center of scene.
the terms of new RCM compensation, and the
compensation of Doppler factor. The form of first
and second ordered motion compensation could quote from
[6][8], the two-step motion compensation method is
used widely in data focusing. It’s necessary to point out
that the Doppler factor is derived from the continuous
move inter-pulse and be considered as the main difference
between pulsed SAR and FMCW SAR.

IV. THE RESULTS OF SIMULATION AND EXPERIMENT

It is common for airplane to be disturbed by the air
current, especially, severe and high-frequency jitter would
make the velocity error obvious and increase the difficulty
for compensation. Firstly, partial parameters of simulation
are settled in Table 1.

Assuming the location error in Y dimension obey a
sine wave whose amplitude and period are 3m and 5s,
while the location error in Z dimension obey a cosine
wave whose amplitude and period are 4m/s and 5s. Then
the velocity error could be calculated by the difference of
location error. Both of them are shown in Fig. 4.

Based on (13) and the parameters in Table 1, we could
calculate the new RCM introduced by inter-pulse motion
effects, shown in Fig. 5.

There are two conclusions can be derived from Fig. 5;
for one thing, the absolute value of new RCM is about
0.2m which will has a obvious influence to imagery when
the range resolution is less than 0.2m. For the other,
referred to the parameter in Table.1, the difference of new
RCM from minimum range or maximum range to central
range are tiny, hence compensating the whole scene with
(14) is available.

Fig.6 shows the result of an analysis of simulated
FMCW SAR data. Fig.6 (a) shows the distribution of
scatters. Fig.6 (b) shows the motion correction result using
traditional compensation method, whereas Fig.6 (c) shows
the motion correction result using the proposed
compensation method. An array of point targets is used for
the qualitative analysis, especially for the points located in
the margin and center of scene. The proposed motion
compensation is compared to traditional motion correction
method in the presence of severe motion error. The better
results performed by proposed method represent the better
applicability for FMCW SAR.

From the theoretical analysis and simulated result, it
has been seen that the proposed method is valid to
compensate the motion errors when the resolution is very
high and the motion error is severe. Although when the
resolution is low and the trajectory is better, new range
migration derived from (13) is less than half of range
resolution, the new range migration derived from
inter-pulse motion error could be ignored.

The experiment was also carried out with our designed
FMCW SAR system, in which the SAR system is fixed on
the roof of a sport utility vehicle (SUV), the LFM
bandwidth is 600MHz, some corner reflectors were placed
in the scene.

<table>
<thead>
<tr>
<th>Range bandwidth</th>
<th>600MHz</th>
<th>altitude</th>
<th>700m</th>
</tr>
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<tbody>
<tr>
<td>Maximum range</td>
<td>1120m</td>
<td>Minimum</td>
<td>910m</td>
</tr>
<tr>
<td>Pulse duration</td>
<td>800us</td>
<td>velocity</td>
<td>30m/s</td>
</tr>
</tbody>
</table>

TABLE.1: PARTIAL PARAMETERS OF SIMULATION

Figure.4 Location and velocity errors in Y and Z dimensions

Figure.5. the new RCMs introduced by inter-pulse motion error

Figure.6. Simulated FMCW SAR data of an array of point targets and
different results are compared with two methods.
From Fig. 6, Fig. 7 and Fig. 8, it is obvious that the SAR is well focused, further, the proposed approach is valid. While more high-precise high-resolution large scene airborne based FMCW SAR data focusing is ongoing. It is confident that the high resolution SAR imagery will have a vast application in kinds of fields. Like remote sensing, monitor, cartography, disaster estimation, environment survey, etc.
V. CONCLUSION AND FUTURE WORK

The main difference between pulsed SAR and FMCW SAR is the approximation stop-go is invalid and the motion during inter-pulse should be considered. Hence the traditional motion compensation method which is suitable for pulsed SAR will no longer available for FMCW SAR. An available method for FMCW SAR motion compensation is proposed in this paper. The geometry and signal model of FMCW SAR in the presence of 3-D motion error is built; the influence introduced by inter-pulse motion error is analyzed while the accordingly approach is presented too. Simulated FMCW SAR data is generated and the proposed method is compared to traditional method. Simulation shows the proposed method results are much better than those of traditional method and the proposed method is particularly available for FMCM SAR. Eventually, the proposed approach is adopted on a laboratorial FMCW SAR system. Well-focused experimental data confirm the effectiveness of the presented method.

REFERENCES


A Novel DDS-PLL Hybrid Structure to Generate the LFM Signal

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Abstract—The signal generator with Direct Digital Synthesizer Phase Locked Loop (DDS-PLL) Hybrid structure generates the signal with the qualities of the wide bandwidth and high frequency resolution. The long acquisition time in the PLL limits its applications. A novel generator structure is proposed by adding the sweeping voltage circuitry to the classical DDS-PLL Hybrid in order to accelerate the acquisition speed. A test circuitry is designed to testify the novel structure. The measured results indicate that the acquisition time is reduced to be 2.175μs and 1.032μs corresponding to 1MHz and 2MHz loop bandwidth respectively; the pulse compression result is very good where the main lobe width remains as ideal, Peak Side Lobe Ratio (PSLR) is better than -38dB, and the Integrated Side Lobe Ratio (ISLR) is better than -9.5dB.

Keywords: DDS-PLL Hybrid; Frequency Sweeping; Fast-locking.

I. INTRODUCTION

With the developments of radar application, the requirements of fast frequency hopping increase. It has considerable benefits to improve the radar performances if the signal has large bandwidth, good frequency linearity and low harmonics [1].

Generally, a signal can be generated in three ways: Phase Locked Loop (PLL) [2], Direct Digital Frequency Synthesizer (DDFS) [3] and Direct Digital Waveform Synthesizer (DDWS) [4]. DDFS is named DDS for short usually. The PLL technique operates from audio frequencies up to millimeter waves and the bandwidth is large, it also has some drawbacks: poor linearity in the frequency modulation rate, long acquisition time, and large phase error. The DDS and DDWS both operate in digital way. Although they have the advantages of fine resolution, fast switching and continuous phase, the disadvantages of serious spur and harmonics and small bandwidth are obvious. For the most advanced commercial DDS circuit, the bandwidth is less than 1.4GHz, the phase noise is better than -128dBc/Hz, and the frequency resolution is 190pHz [5]. The system speed limitation causes the bandwidth of the signal generated by the DDWS to be relatively narrow, only 400MHz most [6].

In the DDS-PLL Hybrid structure, the DDS generates the narrow-band Intermediate Frequency (IF) signal with low intermediate frequency. The PLL up converts the IF to the required RF signal and expands the bandwidth. This structure has the benefits both from the DDS and PLL.

The acquisition time is the most serious problem in the DDS-PLL Hybrid structure. In the paper, we propose a novel structure by adding a sweeping voltage circuitry to the classical DDS-PLL Hybrid structure. It brings in the benefits of shortening the acquisition time. Meanwhile, the predistortion is employed to depress the static phase error and improve the pulse compression qualities.

The remaining sections of this paper are arranged as follows. Section II analyzes the acquisition process. The novel structure is introduced in section III. Section IV lists the measurements to the signals generated by the test system in the fields of acquisition time and pulse compression. This conclusion and future work are in Section V.

II. ACQUISITION ANALYSIS

The classical DDS-PLL Hybrid structure is depicted in Figure 1. In the structure, the BPF is band-pass filter, the PD is Phase Detector, and the LF is low-pass filter. The DDS generates the signal, used as the PLL reference, with the frequency \(f_r\). The frequency of the system clock is \(f_c\).

The PLL outputs the signal with the frequency \(f_c\). The inherent filter in the PLL depresses the harmonics in the DDS output. The PLL multiplies the reference frequency by N to get the signal with the high radio frequency and large bandwidth [7][8][9].

The process of bringing a loop into lock is called acquisition. In Charge-Pump Phase-Locked Loops (CPPLL), the process of acquisition is divided to two parts, frequency pull-in process and phase acquisition process. Pull-in tends to be slow and often unreliable. When the signal was applied, a beat note at frequency \(\Delta \omega\) appears. The frequency pull-in process initiates and the beat note decreases. After the frequency pull-in process, the beat note reaches the pull-in limitation, and phase acquisition process starts. The time from unlocked state to locked state is named acquisition time. It is also divided into be frequency pull-in time and phase acquisition time. In theory, they are defined as follows for a 2\(^{nd}\) order type 2 PLL, respectively [10].

\[
T_p = \frac{\omega_0}{2\pi\omega_n} + \frac{2\xi}{\omega_n} \tag{1}
\]
Where \( \omega_r \) is the phase detecting frequency, \( \omega_n \) is the nature frequency, and \( \xi \) is the loop damping ratio. For example, when the phase detecting frequency is 20MHz, the loop damping ratio is 0.707, and the natural frequency is 500 kHz, then the frequency pull-in time and phase acquisition time are 2.48\( \mu \)s and 2.25\( \mu \)s respectively. In general, the phase acquisition process is faster than the frequency pull-in process.

**III. FAST ACQUISITION**

There are some drawbacks in the classical hybrid structure namely large phase error in the steady state and long acquisition time. We propose a novel structure by adding sweeping voltage controlling circuitry to the classical structure. The novel structure is shown in Figure 2.

**Figure 2. The block diagram of the novel topology**

The CLK is the system clock. The VCO controlling voltage \( V_s \) is a sum of the sweeping voltage \( V_c \) from Digital-to-Analog Converter (DAC) and the fine-tuned voltage \( V_{PLL} \) from the PLL. The sum is realized by a voltage summer.

\[
f_o(t) = K_{VCO} \cdot V_s(t) = K_{VCO} \cdot (V_c(t) + V_{PLL}(t)) = K_{VCO} \cdot V_c(t) + K_{VCO} \cdot V_{PLL}(t)
\]

From (3), we know that the sweeping voltage is a linear function of time. A short pull-in time can be achieved since the output frequency is directly set by the controlling voltage \( V_c \), therefore, the beat note between the divided output and reference input is smaller than the pull-in limitation \( \Delta \omega_p \). The sweeping voltage takes the role of frequency setting. The phase acquisition is accomplished by the PLL. The ideal sweeping voltage curve is shown in Figure 3(a) where the sweeping voltage increases from 1.6V to 9.1V in 100\( \mu \)s as a straight line, and decreases to 1.6V abruptly at the end of the signal pulse. The fine-tuned voltage output by PFD is shown in Figure 3 (b) where the fine-tuned voltage varies fast in a small range of -0.1V to 0.1V. Obviously, two oscillations occur in Figure 3 (b). The first one is the tracking process of the Linearity Frequency Modulation (LFM) signal from 0\( \mu \)s to 100\( \mu \)s. The second one is the process of tracking the falling frequency corresponding to the voltage falling at 100\( \mu \)s in Figure 3 (a). Although the instantaneous frequency migration is very large, thanks to the sweeping voltage which sets the output frequency of VCO primarily, the fine-tuned voltage converges to zero smoothly.

**Figure 3. Simulated transient settling behavior**

However, the actual voltage–frequency characteristic is nonlinear in practice. For instance, the VCO characteristic curve of ROS-2800-719+ is shown in Figure 4.

**Figure 4. Characteristic of VCO**

The dashed curve shows the ideal characteristic. The solid line figures out the actual VCO voltage–frequency characteristic. It approximates to a straight line in the range of 1V to 25V. The nonlinearity is serious when the voltage beyond 25V. We must note that the range with good linearity should be used when generating signals with the device ROS-2800-719+.

To sweep with the actual voltage characteristic we need to sample the actual controlling voltage and store them in the memory. The actual sweeping voltage is generated by DAC with the sample data. Synchronization of the sweeping voltage and the reference is most important. For this reason, the controlling signal SYNC is introduced as shown in Figure 2.

In the novel structure, the sweeping voltage takes the original beat note between the down-scaled frequency and the reference frequency into the pull-in limitation \( \Delta \omega_p \). The frequency pull-in time is reduced to be sufficient small that we can neglect it as compared to the phase acquisition time. So there is phase acquisition time only. The acquisition speed is increased substantially.

The loop filter and summer are shown in Figure 5. The in-phase follower and RC devices in the dash frame compose the active loop filter. The PLL is 3\textsuperscript{rd} order type 2. Different from the ideal 2\textsuperscript{nd} order type 2 PLL, there is an
adding high frequency pole. The influence to the transient induced by the additional pole is small. The PLL approximates an ideal 2\textsuperscript{nd} order type 2. The current $I_p$ output by the PFD is imported into inverter phase adder across the in-phase follower with the sweeping voltage across the RC filter. The summer outputs the final control voltage $V_c$.

The loop filter and summer

The values of the components in the loop filter are listed in Table 1. The software ADIsimPLL provided by ADI can be used to calculate the acquisition time based on listed values.

### IV. MEASUREMENT

We test the novel structure with four modules on hand. They are the DDS module, the DAC module, the filter and summer module, the PFD and VCO module. The connections and the actual circuit are shown in Figure 6.

- The DDS module: The DDS we used is AD9910 with 1GSPS internal clock speed and up to 400 MHz analog outputs. There is an integrated RAM which can be used as the storage of the phase predistortion data;
- The DAC module: The DAC5675 gets 14-bits data widths, up to 400MSPS updating speed. When the SYNC comes, voltage sample data is read out from the storage, and delivered to DAC to generate the sweeping voltage;
- The filter and summer module: This module contains an active filter and an inverse summer. The circuit is as in Figure 6. The amplifiers in the filter and summer are OP1177 and AD823 respectively;
- The PFD and VCO module: The PFD is ADF4113 with the features of 0.2GHz ~4.0GHz RF input frequency, 200MHz phase detect frequency, and max 5mA current output. VCO is ROS-2800-719+ with 1.4GHz~2.8GHz output.

By changing the RC values in the loop filter, we build PLLs with 1MHz and 2MHz loop bandwidth respectively. The signal is generated with 600MHz bandwidth, 100μs pulse width, and 6MHz/μs FM rate. And the output signal is in the range of 1350MHz~1950MHz, the frequency resolution is 5.82Hz. We record the waveform by the oscillograph of DPO71254 with 6.25GHz sample rate.

The measurement of the acquisition time is difficult because the lock-in point is not easy to be determined. In this paper, we fix the lock-in point with the phase error. If the phase error falls into the range of ±10°, starts to oscillate in a small range, and diminishes, we define the loop is locked in. Based on this assumption, the phase errors of the loop with 1MHz and 2MHz bandwidth are drawn in Figure 7 and Figure 8.

<table>
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<th>R3(Ω)</th>
<th>R4(Ω)</th>
<th>R5(Ω)</th>
<th>R6(Ω)</th>
<th>C1(F)</th>
<th>C2(F)</th>
<th>C3(F)</th>
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<td>1k</td>
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<td>1.36n</td>
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</table>

Table 1. VALUE OF COMPONENTS IN THE LOOP FILTER

- The DAC module: The DAC5675 gets 14-bits data widths, up to 400MSPS updating speed. When the SYNC comes, voltage sample data is read out from the storage, and delivered to DAC to generate the sweeping voltage;
- The filter and summer module: This module contains an active filter and an inverse summer. The circuit is as in Figure 6. The amplifiers in the filter and summer are OP1177 and AD823 respectively;
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In the figures, the phase error is very large at the beginning, and then diminishes quickly. We find the lock-in point and mark it in Figure 7 (b) and Figure 8 (b). The acquisition times can be read out from the figures: 2.175μs and 1.032μs for the two loops of different bandwidth respectively. We calculate the acquisition time for the classical structure using ADIsimPLL. The phase acquisition time and frequency pull-in time are 3.43μs and 2.32μs for the loop of 1MHz bandwidth. And they are 2.175μs and 1.032μs for the loop of 2MHz bandwidth. Obviously, the total acquisition time is decreased to approach phase acquisition time. The novel structure proposed in this paper is effective to reduce the frequency pull-in time.

The pulse compression with Hamming weight is shown in Figure 9. The red curve is the pulse compression with ideal signal; the blue curve is the pulse compression with the actual signal. It has perfect...

Figure 7. Locking-in time in 1MHz loop

Figure 8. Locking-in time in 2MHz loop

Figure 9. Compression result
performance. The main-lobe width of the actual signal is the same as the ideal one, the PLSR is better than -38dB, and the ISLR is -9.5dB towards -9.9dB for the ideal.

V. CONCLUSION AND FUTURE WORK

A structure of fast locking signal generator is proposed in this paper, with adding sweeping voltage circuitry to the classic DDS-PLL Hybrid. Great decrease of the frequency pull-in time is achieved because the PLL is brought in the pull-in range by the sweeping voltage. To validate the structure, a test system is built to generate a LFM signal with 600MHz bandwidth, 100μs pulse width and 6MHz/μs FM rate. The acquisition time is reduced to be 2.175μs and 1.032μs with 1MHz and 2MHz loop bandwidth respectively. Based on this novel structure, signals with wider bandwidth and longer duration are expected to be generated to improve the qualities of radar.

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Research of Current Movements of the Earth's crust in the Territory of Almaty City

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Abstract—During the work carried out, measurements were taken by a local GPS network. It was the primary preprocessing and interpretation of the data network for the city of Almaty. The investigation had been carried to find the velocity field of modern movements of the Earth’s surface according to the GNSS-service. All the GPS-data were processed by the software package GAMIT/GLOBK.

Keywords-GNSS-network; IGS; SELE station; GNSS-measurements; velocity field; data interpretation.

I. INTRODUCTION

Deformation of the Earth’s surface on the territories of large cities occurs very often. The main reason is the growing demand of high-rise buildings, the intensive development of underground space, hydro-mode system violations and finally vibration impacts. In some areas of Almaty, the level of impact on the geological structures is already approaching a critical point. This means that the number of destructions will increase dramatically due to any deformations of the Earth’s surface.

Organization of a geodesic effective monitoring system of major cities is possible with high precision satellite methods. This allows almost continuous monitoring in a variety of geographical conditions to observe a certain strain down to millimeter level accuracy [1-4]. These studies were conducted for the first time in the territory of Almaty city.

II. GOALS AND OBJECTIVES

The goal of this work is to study the velocity of modern movements of the Earth’s surface by GNSS-measurements data for the region of Almaty.

On the basis of the known geological conditions, taking into account the differences, a priori data was compiled which preliminary was used for observing the network for the Almaty region. Research showed that it is the stochastic network with distribution points which is the most appropriate solution for this region. We have relied on the size of the foundation block structures that may have a major impact on the surface deformation processes in case of their mutual relative movements during strong earthquakes.

There are several kilometers between large faults when we consider the Earth’s crust depth is about 3 km.

The network is split over the Almaty region with insignificant deviations beyond its limits (Figure 1) and consists of 14 observation points (Table I). The expected results are maps of modern movements in the territory of Almaty city.

III. METHOD

Not being able to organize a network of GPS-synchronized observations in the city, there were a restricted series of successive observations using one receiving station. Measurements are held periodically by each of the points using a Leica GRX 1200 GG Pro receiver, twice a year. The duration of recording satellite signals at one point is 48 hours in a continuous mode with intervals of 30 seconds which was adopted by regulations. Per one session, each point worked about 48 hours and produced 5720 independent acts of registration. This gives an opportunity for further processing.
to increase the ratio of signal/random noise in the square root of this number (about 76 times).

GPS measurements were conducted since 2010 to the present time. GAMIT / GLOBK software package is used for pre-processing GPS data [7].

During calculations of a priori coordinates and velocities of points, a priori data of the Earth's rotation and orbital data generate a single integrated set that must be internally consistent. Imposition of severe restrictions on any of these parameters for a regional solution can deform the system. In order to prevent such action restriction, a priori coordinate, the speed, the parameters of orbit satellite and the Earth's rotation in the solution should not be rigid. However, that decision could shift or collapse the entire system of points and/or their veloctivity vectors. Stabilization allowing to define reference system by minimizing the deviation of the resulting solution which was obtained as the result of a decision using the a priori coordinates and velocities for a set of well-defined stations. The iterative scheme of stabilization automatically excludes measurements points, for which coordinates have large deviations. The velocity fields of the region are shown in Figs. 2 and 3. The rest of the time points are not very accurately measured [4-6].

The permanent station SELE, part of the global network IGS, has an error about 0.1 mm / year. The rest of the time points are not very accurate measurements. This shows that it is necessary to conduct special works to reduce the errors in the city network stations to the order of 1-2 mm/year.

IV. CONCLUSIONS

For the study of modern movements of the Earth's crust of Almaty, was built a local network and data collection from 2011 to 2012.

The first study which was conducted in the city of Almaty showed that within urbanized areas such work is possible and necessary. The main advantage of working with such a system is the possibility of rate parameters simultaneously in three mutually orthogonal components, characterized by significant variations during the year.

The errors of international IGS stations in the pre-processing were about 0.5-2 mm/year, and the local points showing the error of 5-20 mm/year.

Errors are likely due to the small number of independent observations in temporary settlements, due to the inability to organize continuous monitoring because of the absence in receiver numbers and transportation problems.

During two years of follow up in urban areas, set features significant movement of surface points about 3-25 mm/year. The conclusion shows that it is necessary to observe simultaneous points, then moving receivers to other items with overlapping observation systems.

To ensure the high quality of the primary data monitoring it is necessary to choose points of global network stations (IGS) with following control of the data quality.
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Cloud top Height Estimation from the Meteosat Water Vapor Imagery Using Computational Intelligence Techniques: Preliminary Results

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Abstract—This study investigates the cloud top height estimation using nonlinear methods to Meteosat imagery. The suggested approach aims to develop an integrated statistical methodology to estimate the cloud top height on a pixel basis using Meteosat Second Generation water vapor imagery. Radiosonde measurements are used as reference dataset and a spatio-temporal correlation with Meteosat images is performed in order to collect a representative sample for the statistical analysis. Here, we apply Multi Layer Perceptron (MLP) and Support Vector Machines (SVM) and we compare the results to the Linear Regression model. The best results are achieved using SVM for regression. The proposed approach is very promising as it can be used for future in-depth analysis so as to develop a robust approach for geometrical height estimation on a pixel basis of the operational data of Meteosat imagery. It is noted that an accurate estimation of cloud top height can help to eliminate geometric restrictions (e.g. Parallax phenomenon) of the Meteosat satellite imagery, improving its usefulness in a wide area of applications and especially in satellite-based weather forecast.

Keywords—Regression; SVM; MLP; Meteosat; Radiosondes; Geometric height

I. INTRODUCTION

Geostationary satellites nowadays provide many different data about the parameters concerning land, ocean and atmosphere and they are used in numerous applications not only in detecting but also in monitoring (on a real-time basis) and short-range forecasting [1], [2].

There are some limitations because of the geostationary satellite instrument field of view and their scanning mode that usually lead to inaccuracies about the exact location of some cloud features (especially convective cloud systems) on the Earth’s surface. These cloud systems are usually vertically extended at several kilometers in the atmosphere, with the cloud tops reaching the middle and upper troposphere [3-5]. The above-mentioned inaccuracies are caused by the parallax phenomenon, [6] which is more evident in the convective nature cloud systems that are depicted at satellite images and appears at mid-latitudes growing towards the poles and the edges of the field of view of the satellite instruments.

The convective cloud systems have great importance because they are related to extreme weather phenomena and events like lightning, hail, strong winds and floods [5], [7] and [8].

As a result, the accurate estimation of Cloud Top Height (CTH) is appropriate in order to calculate and eliminate the parallax effect and it can provide the real location of the convective systems above the Earth’s surface, more accurately.

There are several studies that propose different approaches and methodologies to estimate the geometric CTH but they are mainly lacking in high accuracy or usually refer to small geographic areas [9-12].

Our effort is to propose a robust, generalized and accurate methodology to estimate the geometrical height of different cloud features (especially convective ones) at a broad range of latitudes and longitudes. This methodology aims to improve on pixel basis the accuracy of clouds’ location above the Earth’s surface, so as to eliminate the parallax phenomenon in real-time monitoring applications based on Meteosat operational data.

Section II provides all the relative information about the data used and the area of interest. In Section III, the methodology was followed in order to estimate cloud top height, is provided. The Section IV presents the accuracy results of the proposed methodology and the Section V provides the conclusions of the the study.

II. DATA AND AREA OF INTEREST

Two different datasets were used in this study. Firstly, radiosonde observations (provided by the National Oceanic and Atmospheric Administration/Earth System Research Laboratory: NOAA/ESRL) of the greater area of the Mediterranean region (Fig. 1). Radiosonde data are suitable to measure vertical profiles of many important parameters like temperature, pressure and geometrical height. Secondly, Meteosat satellite images of two different channels (5 and 6) at spectral ranges of 5.35 - 7.15 μm (spectral center: 6.2 μm) and 6.85 - 7.85 μm (spectral center: 7.3 μm) respectively.
It is mentioned that the specific Meteosat channels have been chosen because they belong to the water vapor absorption spectral region and are used to detect water vapor levels especially in the middle and upper troposphere. All the data used were collected for a period between 5/6/2012 and 14/6/2012. The area of interest covers the greater area of the Mediterranean basin (Fig. 1) because in this region the convective systems are developing frequently and especially during the warm period of the year [13] and [14] causing floods, strong winds and hail among others that affect importantly humans and properties. These cloud systems are vertical extended in several kilometers. The accurate detection of their spatial extent on the Earth’s surface of the field of view of a geostationary satellite like Meteosat, needs an accurate estimation of the CTH in order to eliminate the parallax phenomenon that is evident in such convective cloud systems [6], [12].

### III. METHODOLOGY

A representative training sample was needed in order to determine the non-linear functions that combine the Meteosat pixel values with the cloud geometrical height that is recorded by the radiosonde measurements (Fig. 2). For this reason, a spatiotemporal correlation between radiosonde measurements and satellite images was implemented. More specifically, the radiosonde balloons are usually released twice a day at times 00:00 UTC and 12:00 UTC. For the selected time period, a database was gathered with all the satellite images that have 90-min time distance from the point where radiosonde balloons were released. The 90-min time gap was chosen because our intention is to estimate optically thick and vertically extended cloud systems that are related to severe weather. These are generally called convective cloud systems where the cloud tops existed usually in the middle or the upper troposphere. The balloons reach these atmospheric levels at about 90 minutes after their release according to [15]. So, for each satellite image obtained 90 minutes after the radiosonde balloon release, a 5 x 5 kernel window (~25 x 25 km according to the Meteosat image pixel size at mid-latitudes) was created and implemented around the coordinates of every available Radiosonde point (Fig. 3). The mean Brightness Temperatures (BT) value of the two used channels and their mean value of BT differences of the kernel window pixels, are finally associated with the radiosonde measurement that was considered just above the cloud top (Fig. 2). In order to determine, if the radiosonde passed through a cloud and which is the height of its cloud top, we used the criterion of Relative Humidity level as it was proposed in [16]. If the value of the RH was higher than 85% (maximum value is the saturation level: 100%) then we consider that the balloon was in a cloud. More analytically, the higher (in km) measurement of a radiosonde balloon recordings that has RH lower than 85% and at least the three previous and consecutive measurements with RH higher than 85%, was considered as cloud top measurement (Fig. 2).

![Figure 1](image1.png)  
**Figure 1.** Locations of radiosonde stations that are used in this study.

![Figure 2](image2.png)  
**Figure 2.** Schematic representation of a representative cloud system and its cloud top height, as it is calculated with the proposed methodology.

![Figure 3](image3.png)  
**Figure 3.** The kernel window of Meteosat pixels that was implemented around the location of a radiosonde station.

It is mentioned that the size of the kernel window was chosen because the radiosonde balloons during their ascent can reach more than 20 km horizontal distance from their initial location as it was mentioned in [15].
way using an algorithm developed for this purpose in the VB.NET programming language.

Using the above mentioned spatio-temporal correlation methodology for the radiosonde and Meteosat data, a final dataset of 181 cases in the area of interest for the selected time period, was collected. Each case is a given radiosonde dataset that satisfies the RH criteria. For each case, the geometric height above the ground was recorded (as it was measured from the radiosonde) along with the mean BT in the channels 5 and 6 of the kernel window Meteosat pixels and their mean BT difference. These samples (of 181 cases) comprised our final data set for the regression analysis.

More specifically, we defined our dataset as \( \{ x_i^1, x_i^2, x_i^3, y_i \}, \ i = 1, \ldots, n \) where \( n \) is the number of measured samples (181 cases), the \( x_i^1 \) is the mean BT value of Meteosat kernel window pixels in the channel 5, \( x_i^2 \) is the mean BT value of Meteosat kernel window pixels in the channel 6, \( x_i^3 \) is the BT difference between them and \( y_i \) is the measured height of the balloon for each one of the collected cases.

This is a classical regression problem where the variable that we have to predict is \( y_i \). Here, we examine a Linear Regression model and two non-linear methods: Multi-Layer Perceptron (MLP) Neural Network and Support Vector Machines (SVM), in order to predict this value.

A. Linear Regression

A linear regression model assumes that the relationship between the dependent variable \( y_i \) and the \( p \)-vector of observed \( x_i \) is linear. Thus, the model has the general form:

\[
Y = \mathbf{\beta} X + \mathbf{\varepsilon},
\]

where 
\[
\mathbf{\beta} = (\beta_1, \beta_2, \ldots, \beta_p) \]: are the regression coefficients,

\[
X = \begin{pmatrix} x_1^1 & x_1^2 & \ldots & x_1^p \\
 x_2^1 & x_2^2 & \ldots & x_2^p \\
 \vdots & \vdots & \ddots & \vdots \\
 x_n^1 & x_n^2 & \ldots & x_n^p \end{pmatrix} \] : is the observation matrix and

\[
\mathbf{\varepsilon} = (\varepsilon_1, \varepsilon_2, \ldots, \varepsilon_p) \] : are the error terms.

The parameters \( \mathbf{\beta} \) are estimated using the Least Squares method and a standard solution is obtained [4].

B. MLP

Neural networks are used as a direct substitute for multivariable regression and other statistical analysis techniques. A neural network is used to represent a nonlinear mapping between input and output vectors [2, 18].

Multi-Layer Perceptron (MLP) neural network is the most widely used neural network architecture for classification and/or regression problems. MLP networks consist of an input layer, one or more hidden layers and an output layer (Fig. 4 represents an MLP architecture with one hidden layer). Each layer has a number of units and each unit is fully interconnected with weighted connections to units in the subsequent layer.

The output of MLP is defined as a linear combination of the outputs of the hidden layer nodes where every neuron uses a weighted average of the inputs through a function (e.g. sigmoid function).

\[
g_h = f(\sum_{i=0}^{n} g_h w_{ih}),
\]

where \( f() \) is the activation function, \( g_h \) is the activation of \( h \)-th hidden layer node and \( w_{ih} \) is the weight of the connection between the \( h \)-th hidden layer node and 0-th output layer node. The most used activation function is the sigmoid and it is given as follows:

\[
g_h = \frac{1}{1 + e^{-\sum_{i=0}^{n} w_{ih}}},
\]

The activation level of the nodes in the hidden layer is determined in a similar fashion. The error function is defined as the difference between the calculated value and the target value, as follows:

\[
E = \frac{1}{2} \sum_{i=0}^{l} (y_i - g_i(t))^2,
\]

where \( n \) is the number of pattern in data set and \( l \) is the number of output nodes.

The aim is to reduce the error by adjusting the interconnections between layers. The weights are adjusted using the Back Propagation (BP) algorithm. The algorithm
requires a training data that consists of a set of corresponding input and target pattern values \( y_i \). During the training process, MLP starts with a random set of initial weights and then training continues until the set of \( w_n \) and \( w_0 \) are optimized so that a predefined error threshold is met between \( g_0 \) and \( y_0 \) [3].

**C. SVM for regression**

Support Vector Machines are applied not only to classification problems but also to regression ones. SVM contain all the main features that characterize maximum margin algorithms: a nonlinear function is learned by linear learning machine mapping into high dimensional kernel induced feature space.

In SVM regression, the input \( x \) is firstly mapped onto a \( m \)-dimensional feature space using nonlinear mapping, and then a linear model is constructed in this feature space. The linear model is given by:

\[
    f(x, w) = \sum_{j=1}^{m} w_j g_j(x) + b ,
\]

(5)

where \( g_j(x) \), \( j = 1,...,m \) denotes a set of nonlinear transformations and \( b \) is the bias term.

SVM regression uses a new type of loss function called - insensitive loss function based on [5,6]:

\[
    L_\varepsilon(y, f(x, w)) = \begin{cases} 
        0, & \text{if } |y - f(x, w)| \leq \varepsilon \\
        |y - f(x, w)| - \varepsilon, & \text{otherwise}
    \end{cases} .
\]

(6)

SVM regression performs linear regression in the high-dimension feature space using \( \varepsilon \)-insensitive loss and tries to reduce model complexity by minimizing \( \|w\| \). This can be described by introducing slack variables \( \xi_i^-, \xi_i^+, i = 1, ..., n \), to measure the deviation of training samples outside \( \varepsilon \) - zone.

Thus SVM regression is formulated as minimization of the following functional [6]:

\[
    \minimize \frac{1}{2} \|w\|^2 + C \sum_{i=1}^{n} (\xi_i^- + \xi_i^+) \\
    \text{s.t.} \\
    y_i - f(x_i, w) \leq \varepsilon + \xi_i^+  \\
    f(x_i, w) - y_i \leq \varepsilon + \xi_i^-  \\
    \xi_i^-, \xi_i^+ \geq 0, i = 1, ..., n
\]

IV. RESULTS

We have tested and evaluated the above-mentioned regression models using a dataset of \( n = 181 \) samples. 80% of the dataset was randomly selected and used as the train set and the rest (20%) of the dataset was used as test set.

**TABLE I. THE AVAILABLE DATASET DIVIDED INTO TRAIN AND TEST SET.**

<table>
<thead>
<tr>
<th>Data Set</th>
<th>Number of Instances (%</th>
<th>Number of Instances (#)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Train Set</td>
<td>80</td>
<td>145</td>
</tr>
<tr>
<td>Test Set</td>
<td>20</td>
<td>36</td>
</tr>
<tr>
<td>Data Set</td>
<td>100</td>
<td>181</td>
</tr>
</tbody>
</table>

**A. MLP architecture**

In order to determine the best architecture for the MLP Neural Network, we tested different formulations consisting of one or two hidden layers, with 3 up to 20 neurons in each one. Table II, presents the best architectures for one and two hidden layers respectively in terms of R-coefficient and Mean Absolute Error.

**TABLE II. BEST MLP ARCHITECTURES FOR ONE AND TWO HIDDEN LAYERS.**

<table>
<thead>
<tr>
<th>Layers</th>
<th>Architecture (Input-Hidden-output) nodes</th>
<th>Correlation coefficient</th>
<th>Mean Absolute Error (MAE)</th>
</tr>
</thead>
<tbody>
<tr>
<td>One Hidden Layer</td>
<td>3-18-1</td>
<td>0.7272</td>
<td>1541.05</td>
</tr>
<tr>
<td>Two Hidden Layers</td>
<td>3-18-2-1</td>
<td>0.7372</td>
<td>1399.21</td>
</tr>
</tbody>
</table>

**B. SVM parameter selection**

In order to suggest the most appropriate values for the parameters \( C, \varepsilon \) of SVM, we applied a GridSearch method [7]. Table III, presents the results in terms of the correlation coefficient and MAE for the determined parameters that reported with the maximum correlation coefficient.

**TABLE III. SVM PERFORMANCE FOR THE PARAMETERS \( C = 25 \), \( \varepsilon = 0.1 \).**

<table>
<thead>
<tr>
<th>Layers</th>
<th>Correlation coefficient</th>
<th>Mean Absolute Error (MAE)</th>
</tr>
</thead>
<tbody>
<tr>
<td>SVM</td>
<td>0.7238</td>
<td>1286.97</td>
</tr>
</tbody>
</table>
C. Comparison of the Regression models

Table IV presents a comparison of the results of the three models used in our experiments including the Linear Regression method.

<table>
<thead>
<tr>
<th>Layers</th>
<th>Correlation coefficient</th>
<th>Mean Absolute Error (MAE)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Linear Regression</td>
<td>0.7200</td>
<td>1401.93</td>
</tr>
<tr>
<td>MLP</td>
<td>0.7372</td>
<td>1399.21</td>
</tr>
<tr>
<td>SVM</td>
<td>0.7238</td>
<td>1286.97</td>
</tr>
</tbody>
</table>

Fig. 6 illustrates the SVM predicted values and the actual values for the test set.

![Figure 6. The SVM predictions and the actual values.](image)

Fig. 7 displays the error of the predicted height versus the actual height in meters for the test set. As one can observe as the altitude rises from 8000 to 11000 km, the error becomes significantly large.

This is mainly due to the lack of a sufficient number of cloud tops (as they are defined with the RH criterion) in heights between 8000 and 1200 km in the selected time period. This fact led to a small number of collected cases and consequently to significant errors of the regression procedures at these heights. Increasing the time period of study, the number of the convective cases in the study area and consequently the number of convective cloud tops above 8000m will be increased. This goes beyond the scope of this study and it will be our first priority in our future work.

In this study, all three recorded features were used in order to estimate the height; it would be interesting to study the effect of each one. For this reason we conducted the same experiments with all the combinations of the features and we present the results in terms of the correlation coefficient and MAE in Table V.

![Figure 7. Absolute error of the prediction versus height (in meters).](image)

It is concluded that the most powerful feature is the mean BT value of Meteosat kernel window pixels in channel 5, then is the BT difference between channels 5 and 6 and last is the mean BT value of Meteosat kernel window pixels in channel 6.

<table>
<thead>
<tr>
<th>Corr. Coef.</th>
<th>BT-channel 5</th>
<th>BT-channel 6</th>
<th>BT-channel 5 - BT-channel 6</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.6979</td>
<td>0.3348</td>
<td>0.4017</td>
<td></td>
</tr>
<tr>
<td>MAE 1524.3</td>
<td>1975.2</td>
<td>1813.6</td>
<td></td>
</tr>
</tbody>
</table>

V. Conclusions and Future Work

In this study, some preliminary results regarding the cloud height estimation from Meteosat water vapor imagery were presented. An analytical and automated spatio-temporal methodology is suggested to investigate the correlation of radiosonde measurements and Meteosat Brightness Temperature values. For the estimation of geometrical cloud top height, three methodologies were examined: a linear one and two nonlinear algorithms, MLP and SVM.

Our purpose is to gradually develop a robust and accurate algorithm for CTH estimation for different cloud features (and especially the convective ones) as they are depicted in Meteosat infrared imagery. Our final intention is to cover the operational needs of CTH estimation in large geographic areas, with a simple, accurate and easy to use dataset. Using accurate estimations of the CTH, we intend in the near future to estimate (and reduce) the parallax phenomenon which is evident in the geostationary orbit satellite imagery, mainly used in the satellite-based weather forecasting.

The first results showed that mean errors vary in general between 1000 m and 1500 m and there are significant errors especially after the height of 8000 m in the atmosphere. These errors could be explained due to the lack of data for height above 8000 m in the selected time period. These findings require further investigation which is beyond the scope of this study. Nevertheless, the correlation coefficients
were considered quite satisfactory and promising for all the used algorithms and their parameterizations.

Future work will concern a categorization of the different types of clouds in order to study separately every discrete category of the cloud features that are depicted in the Meteosat satellite imagery. The handling of high cloud tops that penetrate the lower stratosphere is an additional topic that we are planning to study, too.

What is of great importance also, is the collection of a larger dataset that will allow more robust results and the use of additional channels of Meteosat imagery in the infrared region in order to evaluate their performance in the geometric height estimations.

In addition to this, other approaches such as Random Forests [18], could be used to try to test and improve the accuracy of the cloud height estimations.

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The Two-Loop Tracking Strategy for Real-Time Multi-Antenna GNSS Receivers

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Abstract—A multi-antenna global navigation satellite system receiver accepts many radio-frequency inputs simultaneously. Then, many antennas can be deployed on a vehicle and the receiver can decide which one is useful for the reception of each satellite’s signal. As a consequence, navigation solutions can be obtained in more adverse conditions, but at the cost of an important increase of the receiver complexity. Particularly, tracking of the satellite signals is a challenging task in these receivers, especially when a variation in the vehicle’s attitude impose a change in the antenna used for a satellite signal reception. Carrier tracking is critical during these changes of antenna since the distance between them is larger than the signal wavelength. Thus, a direct antenna commutation can produce important tracking transients that degrade the navigation measurements, affect data demodulation, and can produce a loss of signal tracking. The strategy presented in this work allows the receiver to keep tracking during these antenna commutations avoiding the detrimental tracking transients. Since it is based on close-loop schemes, it is suitable for real-time receivers due to their low computational complexity. The proposed technique is tested in a simulated low-earth orbit satellite scenario assuming a constant change of attitude due to a roll movement along the orbit direction vector. It is shown that a four-antenna receiver using the proposed tracking strategy can properly track the satellites in view, in spite of the frequent antenna commutations imposed by the vehicle’s rotation.

Index Terms—GPS; GLONASS; Antenna Radiation Pattern; Carrier Tracking Loops; Antenna Commutation.

I. INTRODUCTION

Global navigation satellite system (GNSS) receivers allow the real-time computation of a vehicle’s position and velocity, generally referred as navigation solutions. For this purpose the receiver has to detect, track and demodulate at least four signals broadcasted from the satellites of the system constellation [1]. Currently, the two operative GNSS systems are the Global Position System (GPS), supported by the United States, and GLONASS developed by Russia that has been completed operational since 1995. Typical GNSS applications rely on a single antenna for satellite signals reception [2]. Since usual antenna radiation patterns are hemispherical, the antenna location in the vehicle must be properly chosen to maximize satellite visibility. Therefore, single-antenna receivers are useful only if the vehicle’s attitude does not depart significantly from their assumed values, as is often the case with cars, planes and ships. However, this is not the case for other applications where the vehicle’s attitude does change considerably, like during the flight of some satellites and rockets. Since vehicle dimensions are larger than the signal wavelength, the direct combination of multiple antennas creates interference patterns that preclude proper receiver operation. A multi-antenna receiver is required in these cases [3]. This kind of receiver can also be used for vehicle’s attitude determination [1], and currently they are also being proposed for usual GNSS applications because better interference and multipath mitigation techniques can be applied [4].

Direct sequence spread spectrum (DS-SS) signals are utilized in GNSS because they allow high time-resolution. Therefore, a correlation stage is needed at the receiver to de-spread them and obtain reasonable signal-to-noise ratios [1]. Tight code and carrier synchronization of the local replica is required to achieve the de-spreading. Typically, tracking loop schemes are adopted for code and carrier synchronization in real-time receivers due to their low computational complexity. We assume a multi-antenna receiver architecture where the signal from each antenna is correlated individually. The same code and carrier local replica is used for all the signals from different antennas when receiving a particular satellite signal, assuming the multi-correlator channel scheme of [4]. However, since the goal in this work is maximizing satellite visibility and not interference and/or multipath mitigation, the beamforming approach is not considered. Then, only the antenna that produces the highest signal-to-noise ratio will be used for the tracking of a given satellite.

Considering that the distance between the antennas deployed on a vehicle is in the order of meters and the carrier wavelength is approximately 0.2 meters, an antenna commutation can be unnoticed by a code loop but can severely affect a phase loop. The degradation caused by this effect to the GNSS satellite data demodulation and some techniques to cope with it has been recently reported on [5]. The sudden change in the input phase can be avoided if that phase difference can be properly measured before the commutation actually occurs. In this case, the radiation patterns of deployed antennas should have a significant overlapping to assure continuous satellite visibility under vehicle’s attitude changes. Then, there should be enough signal-to-noise ratio at the second best antenna to measure the phase difference accurately before the moment when this antenna becomes the best and the tracking loop should change its input and use the new signal. In this work, we propose to use an auxiliary carrier tracking loop to estimate the differences between the signals coming from
the best antenna and the second best one. In this way, the phase transient are completely avoided, the quality of phase measures preserved, and the data demodulation unaltered.

The rest of the paper is organized as follows. A digital model for the output of the correlation of the received GNSS signal from multiples antennas with the same code and carrier local replica is presented in Section 2. The proposed carrier tracking strategy is presented in Section 3. The proposed scheme is tested in a simulated low-earth orbit (LEO) satellite scenario in Section 4. The tracking transient effects found if a single carrier loop scheme is used, and their absence when the two-loop strategy is adopted, are verified. Finally, the conclusions and future work lines are given in Section 5.

II. DIGITAL MEASUREMENTS MODEL

Consider that \( N \) antennas have been deployed on a vehicle and connected to a multi-antenna GNSS receiver. According to the receiver architecture described in the previous section, the signal received at each antenna must be correlated with the same locally generated replica for each visible satellite. The complex correlations of the incoming signal at the \( i \)-th antenna, from a given satellite and for the \( k \)-th correlation interval of duration \( T \), can be expressed as [1]

\[
C_k^i = D_k \sqrt{\frac{T}{N_0}} e^{j \phi_k^i} R(\Delta \tau_k^i) e^{j (\pi \Delta f_k^i T + \Delta \theta_k^i)} + n_k^i
\]

where \( i = 1, \ldots, N \), \( C/N_0 \) is the carrier power to noise power spectral density of the signal from antenna \( i \), \( \Delta \tau_k^i = \tau_k^i - \hat{\tau}_k \) is the code delay estimation error, \( \Delta f_k^i = f_k^i - f^i \) is the frequency estimation error, both assumed constant during the integration time, and \( \Delta \theta_k^i = \theta_k^i - \hat{\theta}_k \) is the initial phase estimation error. The term \( n_k^i \) is a complex white Gaussian noise sequence with unit variance, assumed independent between antennas, \( R(\cdot) \) is the code correlation function, and \( \sin(x) = \sin(\pi x)/(\pi x) \). This expression assumes that the binary data bits \( D_k = \pm 1 \) are the same in all the antenna signals –given the slow data rate, e.g., 50 bps for GPS– and that correlations are computed within the same bit period. This type of binary data modulation is present in the GPS and GLONASS civil signals and in the data components of composite modernized GNSS signals. Notice that the estimated values, which corresponds to the local-replica, are the same in the correlations of different antennas, and that the carrier power to noise power spectral density \( C/N_0 \) is affected by the gain of each antenna according to the line-of-sight vector of the received satellite.

A. Phase Discrimination

After the acquisition process has been completed, i.e., in tracking conditions [1], code and frequency estimation errors are sufficiently small so that the functions \( \sin(\cdot) \) and \( R(\cdot) \) can be approximated by 1. Hence, (1) becomes

\[
C_k^i = I_k^i + jQ_k^i = D_k \sqrt{\frac{T}{N_0}} e^{j \phi_k^i} + n_k^i
\]

where we have defined \( \Delta \phi_k^i = \phi_k^i - \hat{\phi}_k \), with \( \phi_k^i = \pi f_k^i T + \theta_k^i \) and \( \hat{\phi}_k = \pi f_k T + \hat{\theta}_k \), to model the input and output of the digital carrier tracking loops [6].

The phase estimation error present in the signal of each antenna is obtained from the angle of the corresponding complex correlation. To avoid the effects of binary modulation, a two quadrant discriminator should be utilized. Then,

\[
e_k^i = \tan^{-1} \left( \frac{Q_k^i}{I_k^i} \right) = \left[ \Delta \phi_k^i + n_\phi_k \right]_\pi
\]

where the notation \( [\cdot]_\pi \) indicates that its argument is kept within the interval \( (-\pi, \pi] \) by adding or subtracting \( \pi \) as many times as needed. The noise term \( n_\phi_k \) has zero mean and a complicated probability distribution. However, for high \( C/N_0 \) values, it can be approximated by a Gaussian distribution with zero mean and variance \( 1/(2TC/N_0) \).

B. Carrier Tracking Loops

The phase error estimates obtained with (3) are ambiguous due to the periodic nature of the phase. This ambiguity is responsible of many of the non-linear behaviours present in phase locked loops (PLL). Due to these non-linearities the PLL can have very long lock-in times and low resistance to high dynamics, and therefore many GNSS receivers add a frequency locked loop (FLL) in a scheme called FLL-assisted-PLL [7]. In previous works we have proposed an alternative solution, called the unambiguous frequency aided (UFA) PLL [6]. In the UFA-PLL, the ambiguous phase discriminator values, \( e_k^i \), are corrected by adding or subtracting an integer number of \( \pi \). The corrected phase errors, \( u_k^i \), are found such that the difference between two successive values is less than a quarter of a cycle in magnitude. Thus, starting with \( u_0^i = e_0^i \), it must be hold that

\[
u_k^i = e_k^i - I_x(e_k^i - u_{k-1}^i)
\]

where \( I_x(x) = x - [x]_\pi \) is like the integer part function, but with steps at the multiples of \( \pi \). The obtained sequence of phase errors \( u_k^i \) has unambiguous values as long as the frequency error is lower than \( 1/(4T) \). We have also shown that the UFA-PLL has the same noise resistance, and so the same tracking performance, that an equivalent PLL [8].

Through the rest of this work, the utilized carrier loop is the UFA-PLL shown in Figure 1 whose filter coefficients are \( p_1 = C = 0.5 \), \( p_2 = 0.105 \), and \( p_3 = 0.0123 \). For the correlation time, \( T = 5\)ms, the resulting PLL has an equivalent noise bandwidth of 75.6Hz. The two delays included in the loop model the real-time calculation of the correlations. One of them accounts for the time spent in the computation itself, and the second is present because the estimated values utilized to produce the local-replica used in the correlation have to be known before the calculations begin. That is, the estimated value \( \hat{\phi}_k \) is obtained with the loop filter output of the \((k-1)\)-th correlation interval, which in turn is calculated with the estimated value \( \hat{\phi}_{(k-2)} \). The loop filter is optimized for the tracking of acceleration steps, which produces a quadratic ramp of phase at the loop input. This loop design has been implemented in experimental GPS receivers.
[6], and more details and properties can be found in [9]. The loop models previously described can be used operating data-bits synchronously as typically done, but can also be extended for data-bit asynchronous operation as proposed in [10].

III. THE TWO LOOPS TRACKING STRATEGY

Assume the receiver is tracking a given satellite at the antenna \( l \) and that, due to the change in the vehicle’s attitude, the \( C/N_0 \) values are decreasing and the \( C/N_0 \) values, at the antenna \( j \), are increasing. For the sake of simplicity, assume that the rest of the antennas has no useful signal levels. The correlations are computed based on the loop estimates computed with the signal from antenna \( l \), but the error between the phase loop estimate and the phase of the signal from antenna \( j \) can also be obtained as long as the \( C/N_0 \) values are sufficiently high. This new phase error is computed with the signal from antenna \( j \).

\[
\hat{\phi}_k = \tan^{-1}\left(\frac{Q_k}{I_k}\right) = [\Delta \phi_k^j + n_{\phi_k^j}] \pi
\]

where \( \Delta \phi_k^j = \phi_k^j - \hat{\phi}_k \). The phase error needed to build a new loop to track the signal from antenna \( j \) should be calculated with respect to a different phase estimate. However, if we assume that the frequency estimates from both loops are sufficiently close, the new phase error can be obtained from (5) by correcting it with the difference between both loop phase estimates. If we call \( \hat{\phi}_k^M = \phi_k \), the phase estimates of the actual loop or master loop, the phase error of an auxiliary loop, with phase estimate \( \hat{\phi}_k^A \), is

\[
eA_k^j = e_k^j + \hat{\phi}_k^M - \hat{\phi}_k^A = \left[ (\Delta \phi_k^j + n_{\phi_k^j}) \pi + \hat{\phi}_k^M - \hat{\phi}_k^A \right] \pi
\]

The \( [\cdot]\pi \) operation is needed to obtain a phase error in the same range of values than the produced by a two quadrant inverse tangent. Notice that in the previous cases this operation was used to model the ambiguity of the measured phase error, but now it has to be actually implemented as part of the signal processing needed for the new auxiliary loop, which has input \( \hat{\phi}_k \) and output \( \hat{\phi}_k^A \). The loop filter used for the auxiliary loop is exactly the same that the one used in the master loop. The reason for this choice will be clear in the following subsection.

A. Master Loop Actualization

Assume that the auxiliary loop has already been started and its initial transient, due to the initial phase and frequency errors, is already extinguished. Also assume that at the present moment the measured power at the antenna \( l \) becomes lower than the one at the antenna \( j \). At this point the loop with the best estimates is the auxiliary one, and there is no reason to keep using the same loop estimates as parameters for the local-replica. Then, the master loop internal variables are overwritten by the ones from the auxiliary loop, and from now on its input is taken from the \( j \) antenna. In this way, the master loop is again tracking the best antenna, and the change has not produced any transient. Therefore, the logic for the master loop actualization is simple: each time that the power on the auxiliary loop becomes higher than the power on the master, the input of the master loop has to be changed by the auxiliary one. Depending on the auxiliary loop actualization logic, both loops can be tracking the signal from the same or different antennas. When both loops are tracking the signal from the same antenna no master loop actualization is possible, and the correction term of (6) becomes zero, and then \( eA_k^j = eM_k^j \).

B. Auxiliary Loop Actualization

Now is clear that the aim of the auxiliary loop is to track the signal from the second best antenna whenever it is possible. Then, two possible situations have to be managed. The first is when the master loop is tracking the best antenna and the auxiliary one is properly tracking another antenna, which is not the second best antenna any more. That means that there is a better signal to be tracked with the auxiliary loop. Then, a direct commutation between these signals should be done at this loop. Of course, this change will produce a transient due to phase and frequency variations caused by the different physical locations and velocities of these antennas. However, since this loop is not generating the local-replica, the actual signal tracking will not be affected. The second situation is that the tracked signal in the auxiliary loop is actually the second best, but its power is too low to guarantee a satisfactory tracking. Then, the only possible option for the auxiliary loop is to switch its input to the same antenna that the master one. When this is the case, both loops should track the same signal until the second best antenna reaches a highly enough power level. At this point, the auxiliary loop should commute to this new antenna. Again, this change will cause a transient in the auxiliary loop, but it will not affect the master one.
TABLE I
ALGORITHMIC DESCRIPTION OF THE LOOP ACTUALIZATION.

if \( P[a] > P[m] \) then
\[ m \leftarrow a \]
Master loop variables \( \leftarrow \) Auxiliary loop variables
end if
if \( m = b \) and \( P[s] > P[b]/10 \) then
\[ a \leftarrow s \]
else
\[ a \leftarrow b \]
end if

C. Power Estimation

The strategy presented in this work needs a periodic estimation of the received power level for each satellite in view and for each antenna. This information can be obtained almost directly from the already computed correlations, (1), assuming that no interferences or severe signal obstructions or reflections are considered. A raw estimate of the received signal power from antenna \( i \) can be obtained as

\[ \hat{P}_k = |C_{ki}|^2 = (r_k)^2 + (Q_k)^2 = T \frac{C}{N_0} \cdot k + \text{noise}. \]  

(7)

Based on these values, better power estimates can be obtained by temporal filtering accordingly to the desired trade-off between noise reduction and the maximum rate of change of the filtered estimates, which is in turn established by the vehicle’s dynamics. One simple option is to take the average of each of these values during a selected time-window.

D. Summary of the Algorithm

A commutation period should be selected according to the expected vehicle’s dynamics, namely \( L = KT \), with \( K \in \mathbb{R} \). Then, every \( L \) seconds, a vector of estimated power at each antenna should be computed,

\[ \bar{P} = [P[1]P[2] \cdots P[N]] \]  

(8)

where each components is obtained as described in the previous subsection. Then, this vector has to be ordered and the two highest values and their corresponding indexes stored. If we called \( b \) the index of the first best antenna, and \( s \) to the index of the second best antenna, then \( P[b] > P[s] \). Now consider that \( m \) and \( a \) are variables that indicate the number of the antenna selected for the master and the auxiliary loop respectively. Then, assuming that after the acquisition process of the satellite signal the loops are initialized with the best and the second best antenna at that moment, the algorithmic description—in pseudo code—of the proposed two loops tracking strategy is presented in Table I. In this code, it was considered that 10 dB less than the power of the signal in the master loop is the maximum difference allowed to the auxiliary loop to track a different signal.

IV. APPLICATION TO A LEO SATELLITE SCENARIO

In order to test the proposed tracking strategy, a LEO satellite scenario was simulated assuming a constant change of attitude due to a roll movement along the orbit direction vector of 5 revolutions per minute (rpm). This kind of rotation can be found during the first moments of the satellite life due to the spin stabilization used by the launching vehicles. In this case, the signals from the GNSS satellites can arrive from any direction. Then, the tetrahedron geometry shown in Figure 2 (a) was adopted for the placement of four antennas. The use of such non-aligned antenna array for this type of space application has been proposed in [11]. A typical GPS satellite constellation was considered and the position and velocity of each visible satellite were calculated. The same values were obtained for each antenna attached to the LEO satellite, whose trajectory is plotted in Figure 2 (b). Thirteen GPS satellites were found in view during some part of the 10 minutes of simulated trajectory. Notice that given the high velocity of the LEO satellite, \( \simeq 7 \text{ km/s} \), the time in view of a GPS satellite is much shorter than the found in earth applications. Based on the position and velocity values obtained, line-of-sight vectors, distances, and rate of change in these distances were computed every 10 ms to generate the input signals for the following digital processing. The change of the power level received at each antenna is a critical aspect in this analysis and therefore a realistic radiation pattern was considered. Since the GPS signals are radiating using right-handed circularly-polarized (RHCP) signals, the gain pattern of a typical patch antenna at the GPS L1 band (1575 MHz) was considered. The antenna gain pattern adopted is shown in Figure 2 (c).

A. Carrier tracking loop simulation

The loop simulations made are based on the correlation output model of (1). Then, the inputs required for each antenna and visible satellite are the carrier power to noise power spectral density, \( C/N_0 \), the instantaneous carrier frequency \( f_k \) and the instantaneous input phase, \( \theta_k \). Since the code loops are not considered in this work, perfect code estimation is assumed. Neglecting second order terms, the instantaneous phase of the received signal can be obtained directly dividing the distance from the GPS satellite to the considered receiver’s antenna by the carrier wavelength, which is \( \lambda \simeq 19 \text{ cm} \) for the GPS L1 frequency. In the same way, the frequency changes can be obtained considering only the Doppler effect, i.e., dividing the rate of change of the distance by \( \lambda \). Finally, the gain factor \( G_{RHCP} \), obtained with the line-of-sight vector and the gain pattern of the antennas, is converted into carrier power to noise power spectral density, according to

\[ C/N_0[k][dB] = G_{RHCP}[dB] + 42[dB/Hz] \]  

(9)

where the value 42 dB/Hz, was obtained considering 2 dB of antenna losses, equivalent noise temperature of the antenna 130K, cable losses of 0.4dB and receiver noise figure of 1 dB.

B. Simulation results

The results presented in the first place correspond to a single loop tracking strategy. It simply consists on changing the loop input each time that the receiver detects that a better signal is available from another antenna. Only the results obtained with the tracking of one of the GPS satellites in view are presented due to space limitations. In Figure 3 (a), the \( C/N_0 \)
removed to appreciate the tracking error in a proper scale. These ambiguities can be easily compensated by the receiver, allowing continuous and reliable data demodulation.

The results obtained in the same situation but using the proposed two-loops tracking strategy are presented in Figure 4. In Figure 4 (a) the indication of the antennas selected as the input for the master and the auxiliary loops are plotted. It can be noticed that the master loop uses as input practically the same antenna that it would be used by the single loop. However, since its signal has been already tracked by the auxiliary loop, at the time of commutation the master loop can be properly initialized, and correspondingly a transient-free response is obtained as depicted in Figures 4 (b) and (c). Finally, it can be also appreciated that the estimation error levels change accordingly with the variation of the $C/N_0$ values of the tracked signal, as expected.

V. CONCLUSION AND FUTURE WORK

An efficient tracking strategy devised for real-time multi-antenna GNSS receivers has been presented. The considered receiver architecture computes the correlations of the signal from each antenna with a common local-replica, for each tracked satellite. The use of a single carrier tracking loop in this kind of receiver will cause transient errors each time a commutation of the antenna used as loop input is done. These transients degrade the navigation measurements, affect data demodulation, and can produce a loss of signal tracking. The two-loop proposed strategy avoids these transients by properly modifying the internal loop variables at the moments of the commutations. The values of these variables are obtained by a second, twin loop, which does not require the calculation of extra correlations. The proposed strategy has been described in detail and tested in a simulated four-antenna LEO satellite scenario verifying that all the GPS satellites in view can be continuously tracked in the adverse situation of a rolling movement of 5 rpm.

The computational cost of the proposed strategy is minor assuming that in a multi-antenna receiver the power estimation and antenna selection algorithm is mandatory. Given the inherent increase of complexity of managing many radio-frequency inputs coherently, their digitalization, and the computation of the multi-correlations simultaneously, the cost of the extra loop seems negligible. The authors are now working on that multi-correlator architecture to be implemented in field programmable gate arrays (FPGA), and in the design of an experimental multi-antenna GNSS receiver.

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REFERENCES

Fig. 3. Tracking of SV 29 with a single loop strategy.

(a) $C/N_0$ values at each antenna and antenna selection.

(b) Phase estimation error (commuted phase input).

(c) Frequency estimation error (commuted frequency input).

Fig. 4. Tracking of SV 29 with the two loops strategy.

(a) $C/N_0$ values at each antenna and antenna selection.

(b) Phase estimation error (commuted phase input).

(c) Frequency estimation error (commuted frequency input).


