On the Use of PN Ranging With High-Rate Spectrally-Efficient Modulations in Satellite Payload Telemetry Links

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Abstract—Pseudonoise (PN) ranging is a ranging technique that has been recently introduced in near-Earth space research missions. For these, at the state-of-the-art, the PN ranging signal is combined with a high-rate telemetry stream binary modulated in phase. The coupling of the two, together with the 10 MHz bandwidth constraint imposed for this class of missions, translates into a data rate bound of approximately 10 Mb/s. The purpose of this article is to prove the feasibility of overcoming the current data rate limitation by defining a communication architecture that foresees the coupling of the PN ranging signal with a high-order modulated telemetry stream. To achieve the goal, we study the feasibility of pairing the PN ranging with filtered high-order modulations that are standardized for satellite payload telemetry links and investigate the simultaneous demodulation of the telemetry stream while tracking the ranging sequence. Accordingly, we design a receiver scheme capable of performing a closed-loop parallel cancellation of the ranging and the telemetry signal reciprocally. From our analysis, we find that the nonconstant envelope property, characterizing the considered modulation set, causes an additional jitter on the PN ranging timing estimation that, on the other hand, can be controlled and reduced by properly sizing the receiver loop bandwidth without limiting the timing synchronization dynamic. Our study proves the use of filtered high-order modulations combined with PN ranging to outperform the state-of-the-art in terms of spectral efficiency and achievable data rate while having comparable ranging performance.

I. INTRODUCTION

Every space mission relies upon the knowledge of the spacecraft’s position and velocity. Tracking a satellite’s orbit, along with its deviation from the estimated flight path, is known as orbit determination and depends on several types of measurements tailored for each space mission. Some examples are Global Navigation Satellite System data, distance and velocity measurements based on ranging techniques, and tracking data (including antenna angles) through an auto-track mode of the ground station antenna [6], [7].

Ranging techniques rely on a signal transmitted from the ground station to the satellite, and back to Earth. Then, a measure of the distance between the satellite and the ground station is retrieved from the round-trip light time of the ranging signal. Over the years, space agencies developed different ranging systems, such as the NASA sequential ranging and the ESA (European Space Agency) code ranging [8], [9]. Lately, to meet the demanded positioning accuracy while granting the cross-support among space agencies, the consultative committee for space data systems (CCSDS) standardized the pseudonoise (PN) ranging [1], [10]. This last is a state-of-the-art positioning technique that permits the on-board ranging regeneration, thus enhancing the signal-to-noise ratio (SNR) received on Earth. The standard, which envisions the simultaneous transmission of PN ranging with high-rate Gaussian minimum shift keying (GMSK)-modulated telemetry, ensures determining the spacecraft position while receiving the high-rate telemetry stream.

Future space research (SR) missions (e.g., JUICE [11]) will widely adopt PN ranging combined with GMSK: this was already implemented in the ESA missions BepiColombo and Solar Orbiter, and demonstrated an accuracy as high as a few centimeters during its validation [3], [12]. On the other hand, telemetry tracking and command (TT&C) and payload data transmission (PDT) for near-Earth SR missions usually take place in the X-Band (8.4–8.45 GHz). This frequency allocation constraints the occupied bandwidth to 10 MHz [4]. Thus, when using GMSK, SR missions are limited to a data rate of about 10 Mb/s, penalizing the payload data generation. Therefore, to overcome the current data rate bound, this class of missions requires efficient use of the available X-Band bandwidth. For this reason, one must resort to high-order modulation schemes. Their compatibility with PN ranging was investigated only with multi-h continuous phase modulations (CPMs) in [21]. However, this study focuses on...
optimizing the ranging shaping pulse for improving accuracy. Additionally, multi-
$h$ CPMs do not maximize spectral efficiency, and TT&C/PDT standards do not foresee them.

In this article, differently, we study the coupling of PN ranging with filtered (amplitude) phase-shift keying (PSK/APS K) signals by taking as reference the CCSDS standard 131.2 (see [13] and [14]). Namely, we define a receiver scheme that can simultaneously demodulate the PSK/APS K stream (with modulation orders as high as 64) and track the PN ranging sequence. This is done by performing a parallel closed-loop cancellation of the PN ranging sequence over the telemetry symbols and vice-versa. To evaluate the performance of the designed system, both under linear and nonlinear conditions, we assess the bit error rate (BER) and the ranging timing jitter. This last reveals the presence of a noise floor due to the nonconstant envelope of PSK/APS K modulations. By modeling this phenomenon, we derive a closed-form approximation for the ranging jitter.

In this work, we find the performance of the proposed scheme to be comparable to the classical GMSK approach while enabling much higher data rates, thus paving the way toward the next generation of near-Earth SR missions with more ambitious scientific objectives.

The rest of this article is organized as follows. Section II outlines the mathematical model that describes the proposed communication architecture inclusive of the transmission system and channel model. Section III mainly focuses on the description of the implemented receiver system. Then, Section IV analyzes and mathematically describes the impact on the ranging jitter when combining nonconstant envelope modulations with the PN ranging signal. Section V first shows the results coming from computer simulations, then presents examples of application to two space missions: the already deployed GAIA satellite and the possible future mission GAIA-NIR. Finally, Section VI concludes this article.

II. SYSTEM MODEL

This section describes the system model adopted by this article, namely, the transmission system and the channel model.

A. Transmission System

We consider the transmission system described by the block diagram reported in Fig. 1. The complex baseband modulated telemetry signal is defined as

$$x_{TM}(t) = \sum_{k} a_k p(t - kT)$$

where $p(t)$ is the square root raised cosine (SRRC) shaping pulse, and $\{a_k\}$ is the sequence of telemetry symbols, each transmitted at a symbol time $kT$, and belonging to a complex PSK/APS K constellation. Without loss of generality, we assume $\mathbb{E}[|x_{TM}(t)|^2] = 1$. Instead, the phase-modulated ranging signal $r_{RG}(t)$ is defined as $r_{RG}(t) = e^{j\Phi_{RG}(t)}$. Its phase can be expressed as

$$\Phi_{RG}(t) = m_{RG} \sum_{k} c_k h(t - kT_c)$$

where $m_{RG}$ is the ranging modulation index, $h(t)$ is a sinusoidal chip shaping pulse [1] that reads

$$h(t) = \sin \left( \frac{\pi t}{T_c} \right), \quad t \in [0, T_c]$$

and $c_k \in \{-1, 1\}$ is the ranging code chip transmitted at time $kT_c$ and generated according to a balanced Tausworthe scheme [1]. Note that, in line with CCSDS recommendation 2.4.22A [15] and considering that we target high bit rates, we imposed $T_c \neq nT$, with $n \in \mathbb{Z}^+$, and consider $T_c > T$.

B. Channel Model

We consider linear and nonlinear channel models. We adopt the additive white Gaussian noise (AWGN) channel for the linear case. The received signal reads

$$y(t) = x_{TM}(t) \cdot r_{RG}(t - r_{RG}) + w(t)$$

where $w(t)$ is white Gaussian noise with spectral density equal to $N_0$, and $r_{RG} \in [-\frac{T}{2}, \frac{T}{2}]$ is an arbitrary ranging timing delay. Differently, the nonlinear channel considers signal distortions introduced by a power amplifier. In this article, we use the nonlinear channel model described by the block diagram of Fig. 2. It is composed of a traveling wave tube amplifier (TWTA) followed by an RF filter, which compensates for the spectral regrowth. In Fig. 2, the signal output from the TWTA is modeled as

$$z(t) = f_{AM}(\cdot)e^{j\omega t} + f_{PM}(\cdot)$$

being $f_{AM}(\cdot)$ and $f_{PM}(\cdot)$ the normalized AM/AM and AM/PM characteristics shown in Figs. 3 and 4, respectively. Instead, the RF filter is implemented as an elliptic filter, having a passband of about 12.5 MHz.

III. RECEIVER FOR PSK/APS K-MODULATED SIGNALS AND PN RANGING

In this section, we provide the mathematical details of the receiver scheme. We consider a modified version of the...
receiver in [16] performing a closed-loop cancellation of telemetry and ranging signals from the received stream. Fig. 5 shows the receiver block diagram.

Assuming an ideal lock and in the absence of nonlinear distortions, the ranging component is removed from the received signal by using a locally generated replica of the ranging signal. Mathematically,

$$ y(t) = x_{TM}^e(t) + w(t) r_{RG}^e(t - \hat{\tau}_{RG}) $$

(6)

being $r_{RG}(t)$ the ranging signal’s complex conjugate locally generated using the estimated time delay $\hat{\tau}_{RG}$. It is easy to see that, being $r_{RG}(t)$ a phasor, the resultant process $w(t) r_{RG}^e(t - \hat{\tau}_{RG})$ is still white Gaussian, and $x_{TM}(t)$ is perfectly recovered as long as $\hat{\tau}_{RG} = \tau_{RG}$. The telemetry stream, thus, obtained is demodulated to extract data and then remodulated to obtain the sequence of symbols. With it, we generate the complex conjugate $x_{TM}^c(t)$, necessary to suboptimally cancel the telemetry component from the received signal. We can write

$$ y(t) x_{TM}^c(t) = r_{RG}(t - \tau_{RG}) |x_{TM}(t)|^2 + \mathcal{W}(t) $$

(7)

where $\mathcal{W}(t) = x_{TM}^c(t) w(t)$ can be proved to be a white process with power spectral density equal to $N_0$. Since $x_{TM}(t)$ is SRRC filtered, $|x_{TM}(t)|^2 \neq 1$, and thus, perfect cancellation cannot be achieved. However, taking into account that $\mathbb{E}[|x_{TM}(t)|^2] = 1$, suboptimal cancellation can be performed by averaging the samples of the telemetry signal’s time-varying envelope, as shown later in Section IV.

Once recovered, the ranging sequence is input to the chip tracking loop (CTL), whose block diagram is shown in Fig. 6. Since the ranging chip sequence resembles a clock signal, we implement the CTL as a modified version of the data transition tracking loop (DTTL) [10], [18]. In this case, the mid-phase integrator operates according to an integrate and dump approach, integrating the PN sequence over a time interval $T_c$ between two consecutive chips. Then, the output of the mid-phase integrator multiplies the PN transition sequence to adjust the sign of the error at the input of the loop filter. Finally, the estimated CTL timing $\hat{\tau}_{RG}$ is fed back to the local PN generator in Fig. 5, thus closing the receiver loop.

Although the here-presented receiver is suboptimal, we found it capable of performing well. Additionally, the receiver modular design allows (potentially) reusing the acquisition and synchronization chains adopted in a classical PSK/APS modulator (as the one in [14]) and in a CTL, thus using the same approach currently implemented for GMSK with PN ranging [3], [16]. In particular, since the GMSK can be seen as a quaternary modulation, we expect the quadrature phase shift keying (QPSK) synchronization and PN acquisition to work similarly. Differently, the classical synchronization could be subject to higher losses when dealing with higher order PSK/APS modulations. However, the high SNR (driven by the high bit rate), or the use of pilots as foreseen by standards, can compensate for the additional losses. In light of this, the rest of this article entirely focuses on the performance in tracking conditions. The detailed study of the synchronization is left open for future implementation activities on this topic.

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1It is pointed out that the signal $x_{TM}(t)$ is actually a cyclostationary process, thus making $\mathcal{W}(t)$ cyclostationary as well. However, without loss of generality, in this article, we consider their stationary statistics by simply referring to the cyclic autocorrelation at null cyclic frequency [17].

2During our research activities, we investigated other cancellation methods, for instance $x_{TM}^c(t)/|x_{TM}(t)|^2$. However, since other methods are responsible for coloring the noise, the one adopted in this article was found to be the simplest suitable solution.

3For instance, CCSDS 131.2 foresees the use of $\pi/2$-BPSK pilots [13], [14].
IV. RANGING TIMING JITTER

In this section, we analyze the timing jitter of the receiver scheme presented in Section III.

We define $P(t) = |x_{TM}(t)|^2 - 1$ as a zero-mean signal. Thus, (7) can be rewritten as

$$y(t) = m_{TM}(t) = r_{RG}(t - \tau_{RG}) + W(t) + r_{RG}(t - \tau_{RG})P(t)$$

(8)

where the last term can be understood as additional noise arising from the nonconstant envelope of the considered modulation formats. To assess the effects generated by this new noise term, we analyze the output of the CTL. As shown by Fig. 6, the CTL extracts the imaginary component of the ranging signal, which reads

$$\alpha N_{RG}(t - \tau_{RG}) + \mathcal{N}(t) + \mathcal{D}(t)$$

(9)

where $\mathcal{N}(t)$ is the imaginary component of the Gaussian noise, $\mathcal{D}(t) = \alpha N_{RG}(t - \tau_{RG})P(t)$, and $\alpha \approx 2J_1(m_{RG})$. Thus, the ranging signal power at the CTL input is $P = \alpha^2/2$. Therefore, we define $n_k \triangleq \int_{kT}^{(k+1)T} N(t) dt$ and $\xi_k \triangleq \int_{kT}^{(k+1)T} D(t) dt$, i.e., the output of the mid-phase integrator when $N(t)$ and $D(t)$ are the inputs, with autocorrelation $\mathcal{R}_n[k]$ and $\mathcal{R}_\xi[k]$, respectively. While it is easy to see that $\mathcal{R}_n[k] = (N_0 T_c/2) \delta_n$, a closed-form solution for $\mathcal{R}_\xi[k]$ is not easy to obtain. However, as shown later, it can be numerically characterized.

At this stage, we are ready to derive a linearized model of the CTL, as reported in Fig. 7, in which both the discretized noise terms at the kth time ($n_k$ and $\xi_k$) are added to the phase detector output $\xi_k$.

The loop error is defined as

$$\eta_k = \xi_k + \frac{n_k}{K_c} + \frac{\xi_k}{K_e}$$

(10)

where $K_e = 2\sqrt{2F}$ is the mid-phase integrator gain [10]. Therefore, considering that $n_k$ is the input to the loop filter having transfer function $H(f)$, the relation that describes the ranging timing jitter $\sigma^2_\tau$ reads

$$\sigma^2_\tau = T_c \int_0^{1/T_c} \left| \frac{K_c H(f)}{1 + K_c H(f)} \right|^2 S_\eta(f) df$$

(11)

being $S_\eta(f)$ the power spectral density of the input $\eta_k$. Since the integrated loop bandwidth is usually very narrow w.r.t. the ranging signal rate, (11) can be rewritten as

$$\sigma^2_\tau \approx T_c \int_0^{1/T_c} \left| \frac{K_c H(f)}{1 + K_c H(f)} \right|^2 df S_\eta(0).$$

(12)

Considering that the integral in (12) is the definition of the loop bandwidth $B_L$ [19], and that the random variables $n_k$ and $\xi_k$ are uncorrelated, we can finally approximate the ranging timing jitter as

$$\sigma^2_\tau \approx \frac{2B_L T_c}{K_e^2} \sum_k \mathcal{R}_n[m] + \sum_k \mathcal{R}_\xi[k]$$

(13)

$$= \frac{2B_L T_c}{K_e^2} \left[ \frac{N_0 T_c}{2} + \alpha^2 T_c^2 \sigma^2_\epsilon \right]$$

(14)

$$= T_c^2 \left[ \frac{1}{8} \frac{N_0 B_L}{P} + \sigma^2_\epsilon \right]$$

(15)

where, with an abuse of notation, it has been defined $\sum_k \mathcal{R}_\xi[k] = \alpha^2 T_c^2 \sigma^2_\epsilon$, explicitly reporting its dependence on the time and power factors $T_c^2$ and $\alpha^2$.

Interestingly, $\xi_k$ can be seen as an additional white noise process. Namely, we observe that the first right-hand term in (15) coincides with the theoretical PN ranging jitter [10] and decreases as $P/N_0 B_L$ grows. However, due to the nonconstant envelope of the modulation, the second term appears, acting as an additional interference. Since this interference is proportional to $B_L$, the only way to reduce it is by narrowing the noise loop bandwidth, i.e., averaging the samples of the varying envelope.

The term $\sigma^2_\epsilon$, representing the normalized sum (with respect to $T_c^2 \alpha^2$) of the coefficients of the autocorrelation function, depends on the modulation scheme, roll-off factor, and telemetry-symbol-rate to chip-rate ratio. Therefore, finding a closed-form solution can be cumbersome. However, as Section V will show, $\sigma^2_\epsilon$ can be easily calculated through computer simulations, estimating the autocorrelation of the mid-phase integrator output when $D(t)$ is given as input.

V. NUMERICAL RESULTS

In this section, we first present the performance of the designed telemetry ranging system from a physical layer perspective. Namely, we report the BER and timing jitter (derived through numerical simulations) in Section V-A. Then, using these results, in Section V-B, we analyze the performance of the proposed communication architecture at a satellite system level, by taking as a reference the near-Earth SR mission GAIA and the still ongoing study of the ESA GAIA-NIR.
A. Physical Layer Results

We simulate the transmission of a SRRC-filtered telemetry signal, PSK/APSK modulated (with order ranging from 4 to 64), having channel symbol rate $R = 1/T = 4.2$ Msymbol/s, and together with a PN ranging signal of rate $R_c = 1/T_c = 3$ Mchip/s. For the sake of simplicity, at the receiving end, we consider a CTL of the first order. However, the extension to higher order loops, for better tracking effects as Doppler and Doppler rate, can be easily done employing standard design techniques [19]. First, we carry out simulations under linear conditions with the AWGN channel. These serve as a benchmark for the nonlinear results later provided.

1) Linear Channel: We first estimate $\sigma^2_\zeta$ of (15) by characterizing the autocorrelation function $R_\zeta[k]$. For the sake of clarity, Fig. 8 shows the example for a QPSK-modulated telemetry stream, with roll-off 0.2: we derive the most significant components of the autocorrelation function, such as the ones provided in the picture, via Monte Carlo simulations. Then, $\sigma^2_\zeta$ is approximated as a sum of these elements. We repeat this approach for different modulation formats and roll-off factors and obtain the estimated values reported in Table I. Here, we can observe a common trend regardless of the modulation order: the higher the roll-off factor is, the smaller $\sigma^2_\zeta$ becomes. In line with the expectations, higher roll-off factors correspond to a more constant envelope of the telemetry signal and, thus, lower power fluctuations. Additionally, by fixing the roll-off factor to 0.2, we analyze $\sigma^2_\zeta$ when the telemetry-symbol-rate to chip-rate ratio, $R/R_c$, changes. The findings, presented in Fig. 9, highlight an almost descending and linear behavior of $\sigma^2_\zeta$ for increasing values of $R/R_c$ for QPSK and 8-PSK, which results in a decrease of the ranging timing jitter floor. In fact, by increasing the number of symbols per chip input to the CTL, there is a better averaging of the samples, thus reducing the telemetry power fluctuations. Whereas, we cannot draw the same conclusion for 16-, 32-, and 64-APSK for which there is no sharp reduction of $\sigma^2_\zeta$ as $R/R_c$ increases. Additionally, these curves do not express a linear behavior.

Now, we analyze the ranging jitter in the ideal condition of perfectly demodulated telemetry symbols. In this way, we assess how the SRRC filtering impacts the timing recovery regardless of erroneous demodulation events. To do so, we implement and simulate a genie-aided receiver provided with the correct sequence of symbols $\{a_k\}$. We then, evaluate the timing jitter $\sigma^2_\tau$, normalized to $T^2_c$, via Monte Carlo simulations, as a function of the SNR $P/N_0B_L$ and for different values of $B_L$. Figs. 10 and 11 present the simulation results for SRRC filtered (roll-off = 0.2) QPSK and 64-APSK modulations with the genie-aided receiver. For comparison, the figures also include the curves derived by means of (15) (by using the estimated $\sigma^2_\zeta$ in Table I) and the jitter of the PN ranging with GMSK-modulated telemetry signal. As foretold, the jitter has a floor that decreases with $B_L$, and thus the performance gets closer to the GMSK jitter. In particular, $B_L$ equal to 150 Hz provides almost an ideal performance without limiting the timing

Fig. 8. Discrete-time autocorrelation function $R_\zeta[m]$, for a QPSK-modulated telemetry stream (roll-off = 0.2).

Fig. 9. $\sigma^2_\zeta$ as function of $R/R_c$ for different modulation schemes and roll-off factor 0.2.

Fig. 10. Impact of SRRC-filtered QPSK modulation (roll-off = 0.2) on ranging timing jitter, computed by using a genie-aided receiver.
Fig. 11. Impact of SRRC-filtered 64-APSK modulation (roll-off = 0.2) on ranging timing jitter, computed by using a genie-aided receiver.

Fig. 12. AWGN end-to-end ranging timing jitter for QPSK and 8-PSK modulations (B_L = 1.5 kHz, roll-off = 0.2).

Fig. 13. AWGN end-to-end telemetry BER for 16-APSK, 32-APSK, and 64-APSK modulations (B_L = 1.5 kHz, roll-off = 0.2).

Fig. 14. AWGN end-to-end ranging timing jitter for 16-APSK, 32-APSK, and 64-APSK modulations (B_L = 1.5 kHz, roll-off = 0.2).

Fig. 15. AWGN end-to-end telemetry BER for 16-APSK, 32-APSK, and 64-APSK modulations (B_L = 1.5 kHz, roll-off = 0.2).

TABLE II
IBO and (corresponding) OBO Values Adopted for PSK/APSK Modulations

| Modulation | OBO [dB] | IBO [dB] |
|------------|----------|----------|
| QPSK       | 0.33     | 0        |
| 8-PSK      | 0.31     | 0        |
| 16-APSK    | 2.02     | 6        |
| 32-APSK    | 4.34     | 10       |
| 64-APSK    | 7.23     | 14       |

synchronization dynamic. Additionally, the closed formula expression well predicts \( \sigma^2 \).

Finally, we evaluate the end-to-end performance of the receiver scheme of Fig. 5 from the telemetry and ranging points of view by simulating the complete communication chain for all SRRC filtered PSK/APSK modulations. Figs. 12, 13, 14, and 15 present the ranging timing jitter as a function of \( P/N_0 B_L \) and the telemetry’s BER versus \( E_b/N_0 \) (being \( E_b \) the energy per bit). For comparison, the BER figures also show the theoretical curve in AWGN and in absence of ranging. Figs. 12 and 14, showing timing jitter performance, reveal the same trend for the modulation sets. Despite an initial deviation (more pronounced as the modulation order grows), the simulated ranging jitter curves converge to the genie-aided one. Moreover, although the telemetry and ranging performances are inherently intertwined, the worsened jitter performance for low values of \( P/N_0 B_L \), with respect to ideal cancellation, is not reflected in the BER. In fact, the end-to-end BER is comparable to the theoretical one. In light of this, one could use channel coding (e.g., the serially concatenated convolutional codes foreseen by CCSDS 131.2 [13]) and obtain the performance results found in [14] without a significant implementation loss.

2) Nonlinear Channel: To consider a more realistic scenario that includes power amplification, we rerun the simulations under nonlinear conditions, with the channel model of Fig. 2, using the input and (corresponding) output back-off (IBO and OBO) values reported in Table II.
TABLE III
Occupied Bandwidth for PSK and APSK (SRRC Filtered With Roll-Off 0.35) and GMSK, Over the Nonlinear Channel, as a Function of the Ranging Modulation Index $m_{RG}$

| $m_{RG}$ | PSK  | APSK | GMSK  |
|---------|------|------|-------|
| 0.111   | 7.4 MHz | 7.1 MHz | 7.22 MHz |
| 0.222   | 7.6 MHz | 7.4 MHz | 7.31 MHz |
| 0.444   | 8.4 MHz | 8.1 MHz | 7.72 MHz |
| 0.666   | 8.7 MHz | 8.7 MHz | 8.23 MHz |

Fig. 16. Normalized spectra for 8-PSK (SRRC filtered with roll-off 0.35), over the nonlinear channel, as a function of the ranging modulation index $m_{RG}$.

First, we analyze the occupied bandwidth, as 99% of the signal power, of GMSK- ($B_{T} = 0.25$), PSK-, and APSK-modulated telemetry streams with different ranging modulation indexes. For a fair comparison, we consider a GMSK signal having a bit rate equal to the QPSK-modulated telemetry stream. Instead, we compare the other modulations using the same channel symbol rate.

Trivially, the spectral occupancy changes as a function of the ranging modulation index $m_{RG}$. As shown in Table III, the higher $m_{RG}$ is, the more the spectrum broaden. This effect can be seen very well in Fig. 16, that provides as an example of the spectra of 8-PSK at the output of the RF filter, for different values of $m_{RG}$. Interestingly, despite the spectral regrowth caused by power amplification, the PSK/APSK scheme has an occupied bandwidth that is comparable to the one of the GMSK scheme.

Next, we simulate the end-to-end communication system. As for the linear case, we present results for QPSK and 64-APSK modulations, for which the ranging timing jitter is reported in Figs. 17 and 18, respectively. The figures show both linear and nonlinear performances, revealing a gap of about 2 dB between the two due to distortions caused by the power amplification and the filtering process.

Finally, for assessing the telemetry performance, we analyze the BER degradation due to the presence of ranging. Figs. 19 and 20 show an example for QPSK and 64-APSK modulation, respectively. The obtained results imply no additional losses with respect to a scenario free of the ranging signal.

$^a$BT is the normalized filter bandwidth.

B. System Analysis Results

In this section, we adopt the results of the physical layer simulations for analyzing the performance of two cases of study: GAIA and GAIA-NIR. GAIA [22] is an ESA mission launched in 2013 that achieved 8.7 Mb/s in X-Band using a GMSK signal. GAIA-NIR, instead, is an ongoing ESA study that could improve the data return by using a higher equivalent isotropically radiated power (EIRP). Both mission scenarios need to maximize the achievable telemetry bit rate, while keeping a ranging accuracy below
TABLE IV
GAIA/GAIA-NIR Link Budgets

|            | GAIA | GAIA-NIR |
|------------|------|----------|
| EIRP [dBW] | 8500 | 8500     |
| Sat. altitude [km] | 1368000 | 1760000 |
| Elevation angle [deg] | 50   | 50       |
| Slant range [km] | 1369486 | 1765259 |
| Path loss [dB] | 234 | 234      |
| Atmospheric losses [dB] | 0.27 | 0.27     |
| G/T [dBK] | 53.1 | 50.2     |
| Rec. SNR [dBHz] | 81.7 | 75.8     |
| Margin [dB] | 3 | 3        |
| Modcod | 13 | 7         |
| $E_b/N_0$ [dB] | 4.45 | 2.41     |
| Demodulation losses [dB] | 2.90 | 0.95     |
| $S/N_0B_L$ [dB] | 39.7 | 33.9     |
| Timing jitter $\alpha^2/T_e^2$ [dB] | -44.27 | -39.92 |
| Jitter [m] | 0.61 | 1.00     |

5 m. Consequently, this section aims to prove the gain in the data rate of the here-proposed communication architecture for these two mission scenarios with respect to the use of the state-of-the-art GMSK and PN ranging scheme.

In pursuit of our goal, we carry out a link budget analysis, for which we choose the Cebreros site [23] as the receiver station. For modulation and coding formats (modcods), we consider the 27 modcods provided in CCSDS 131.2 (see [13] and [14]), which foresees PSK/APSK modulations as those considered in this article, with different coding rates. Starting from the satellite EIRP and the best and worst case satellite altitude, we select the modcod that maximizes the bit rate, while complying with the 10-MHz bandwidth limitation [4]. Then, we assess the ranging jitter (in meters) when using a CTL bandwidth $B_L$ of 1.5 kHz, which does not limit the timing tracking in presence of Doppler.7

For instance GAIA and GAIA-NIR could adopt a loop bandwidth as low as 1 mHz.

Table IV summarizes the findings of the link budget. According to this, GAIA, in the best case, would employ the modcod 13 (16-APSK), leading to a bit rate of 15.2 Mb/s. In the worst case, it can achieve up to 8.8 Mb/s by means of modcod 7 (8-PSK). GAIA-NIR, on its own, can reach a bit rate of 27.4 Mb/s by means of modcod 25 (64-APSK) and 18.9 Mb/s by means of modcod 17 (16-APSK). In all cases, the ranging jitter was below 1 m. Consequently, the results in Table IV highlight how the proposed system overcomes the current bit rate limit while achieving the required ranging performance.

VI. CONCLUSION

In this article, we investigated the possibility of combining the PN ranging with high-order, spectrally efficient PSK/APSK modulations. At the receiver side, we designed, implemented, and simulated a receiver scheme for simultaneous demodulation of high-order telemetry symbols and tracking the received ranging sequence for orbit determination. The obtained results show that it is possible to achieve a ranging performance comparable to the current state-of-the-art receiver architecture by simply lowering the noise loop bandwidth $B_L$. Furthermore, the use of such modulations in telemetry-ranging systems enables a data rate way higher than 10 Mb/s.

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