C/No Estimation in a GPS Software Receiver in the Presence of RF Interference Mitigation via Null Steering for the Multipath Limiting Antenna

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***Abstract*— Estimation of the Carrier-to-Noise ratio C/N_0 is an important aspect of the functionality of global positioning system (GPS) receivers. This estimate is used to weight the measurements obtained from each satellite in order to improve the accuracy of the position calculation. The Multipath Limiting Antenna (MLA) has shown promising results in combating ground multipath errors. Radio frequency interference (RFI) is a major error source for GPS receivers. Such interference can be reduced by using null steering at the antenna side. In this work, we present a novel null steering approach for the MLA, and demonstrate its effect on the estimation of the C/N_0 in a GPS software receiver. Two types of RFI sources are investigated; narrow band Continuous Wave (CW) and wide band Pulsed wave (PPS). Performance of two C/N_0 estimation algorithms is compared under different RFI and directions of arrivals.**

I. Introduction

Local Area Augmentation System (LAAS) is an augmentation between the satellite based Global Positioning System (GPS) service and ground based stations to provide accurate correcting information for aircraft precision approach and landing in airports. The initiative was proposed by the Federal Aviation Administration (FAA) in late 1990's. LAAS has to meet stringent requirements that will provide accuracy, integrity, continuity and availability of service for aircrafts during their final approaches to airports that require the greatest safety and reliability. This relys on local area differential GPS (DGPS) for horizontal and vertical position fixing [1], along with integrity checks.

Integrated Multipath Limiting Antennas (IMLA) are being used in Local Area Augmentation Systems (LAAS) to conform to the multipath rejection capabilities that is critical to meeting the integrity and accuracy of operation in fight control environments [2]. Such an antenna is a dual band antenna operating in the L1/L2 GPS bands [3]. It consists of two antennas, a High Zenith Antenna (HZA) covering from 90^o - 30^o in elevation, and a Multipath Limiting Antenna (MLA) covering from 35^o - 0^o. The Desired-to-Undesired (D/U) ratio which is a parameter that describes the ability of the antenna in rejecting ground multipath signals and has to be $\geq$ 30 dB [2], [3], [4].

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In the presence of intentional or un-intentional RFI, the received GPS signal might get blocked in the receiver, and the tracking of the satellites may be lost. In order to mitigate such type of interference, the use of beam-forming / null-steering techniques in baseband processing can be utilized [5] (although some pass band techniques have been proposed, but their feasibility is not always practical [6]). Beam-forming is performed by designing a planar patch antenna array that is to be placed under the MLA. By feeding the elements with the appropriate levels, the beam is steered in the direction of the interference. Then, in a digital signal processor (DSP) the formed beam is subtracted from that of the original MLA pattern to yield a null in that direction. This subtraction can also be performed at RF using power dividers/combiners. The array is designed for L1 GPS band (1575.42 MHz) with a bandwidth greater than 20 MHz to cover both standard positioning service (SPS) and precise positioning service (PPS).

The ratio of the received carrier power level to the noise power level in a 1Hz bandwidth is called the Carrier-to-Noise ratio (C/N_0)- a normalized measure to Signal-to-Noise ratio (SNR). It is an excellent measure for the estimation of errors in phase observables. It is an important parameter describing the performance of a GPS receiver, and its value determines the precision of the pseudo range and carrier phase calculations.

The novel null steering (NS) method (denoted as beam subtraction) described will have a minimum effect on the MLA phase center variation (PCV) and group delay (GD) compared to the conventional null steering based on feed level adjustments [7]. The original MLA PCV and GD curves are preserved outside the null direction, meaning that original MLA compensation curves are applied without any alteration. This is not the case if the null is steered by changing the weights of the antenna elements, where a new correction curve is needed for every null direction. The effect of applying this NS method on the C/N_0 estimation is investigated in this paper.

The paper is organized as follows. Section II will cover the MLA architecture and performance parameters. Section III will cover the design of the patch antenna array. Section IV presents the results, and section V concludes the paper.

II. C/No Estimation Methods

Two C/N_0 estimation algorithms are investigated in this effort, they are:

A. The Variance Summing method (VSM)

This method of C/N_0 estimation was described in [8]. For every value of the resulting I and Q accumulations (1 ms accumulation intervals are used in this effort), we find the sample of the time series Z, Z_k by

$$Z_k = (I_k^2 + Q_k^2) \tag{1}$$

the resulting time series Z has a mean $\overline{Z} = \frac{1}{K}\sum_K Z_k$ and variance $\sigma_Z^2 = \frac{1}{K-1}\sum_K (Z_k - \overline{Z})^2$. The average carrier power is given by

$$\left(\frac{NA}{2}\right)^2 = \sqrt{\overline{Z}^2 - \sigma_Z^2} \tag{2}$$

where N is the number of samples in the time series Z, and A is the signal amplitude at the front end output. Now, for the I and Q noise components, we can find the variance of the noise accumulation terms as,

$$\sigma_{IQ}^2 = \frac{1}{2}\left(\overline{Z} - \sqrt{\overline{Z}^2 - \sigma_Z^2}\right) \tag{3}$$

from which we can find

$$\frac{C}{N_0} = 10LOG_{10}\left[\frac{1000(NA/2)^2}{2\sigma_{IQ}^2}\right] \tag{4}$$

B. The Power Ratio method (PRM)

This method is also used to estimate C/N_0 and is described in [9]. For a detailed derivation the reader is referred to [9], [10]. It involves the comparison of the powers in two different bandwidths to estimate the SNR, and relate that to C/N_0. The wide-band power WBP_k at time t_k has a bandwidth of $1/T$, where T is the accumulation time for power measurement. Similarly, the narrow-band power NBP_k has a bandwidth of $1/MT$ for the accumulation time T, and M is the number of samples occurring in time T. Bit synchronization is necessary to compute the C/N_0 using this method. The navigation bit transition should be determined so that we would not sum over a bit transition and result with incorrect results.

Both wide-band and narrow-band powers are sums of I and Q samples coming out of the correlator output. The accumulation times used were 1 ms for both channels. The wide-band power over M Samples is given by:

$$WBP_k = \sum_{i=1}^{M}\left(I_i^2 + Q_i^2\right) \tag{5}$$

while the narrow-band power is given by,

$$NBP_k = \left(\sum_{i=1}^{M} I_i\right)^2 + \left(\sum_{i=1}^{M} Q_i\right)^2 \tag{6}$$

These power quantities are calculated using samples within the same navigation bit. With a correlator averaging time of 1 ms, this leads to $M = 20$ correlator values within a single navigation bit that has a frequency of 50 Hz. This gives possible values for M to be $\in [1, 2, 4, 5, 10, 20]$. The ratio of the narrow-band to wide-band power at time t_k yields an estimate of the noise power NP_k at the epoch t_k,

$$NP_k = \frac{NBP_k}{WBP_k} \tag{7}$$

In order to transform the noise power into a carrier-to-noise estimate, one must average the noise power. The expected noise power value (μ_{NP}) is the average noise power over $h = K/M$ values, where K is the averaging time interval passed to the algorithm, this gives

$$\mu_{NP} = \frac{1}{h}\sum_{k=1}^{h} NP_k \tag{8}$$

$$\frac{C}{N_0} = 10LOG_{10}\left[\frac{1}{T}\frac{\mu_{NP} - 1}{M - \mu_{NP}}\right] \tag{9}$$

The averaging should not occur over a navigation bit transition. If this happens, the energy estimation might vary significantly, and lead to an error in the power estimation for that interval (summing over a navigation bit transition), thus causing an error in the estimation of the C/N_0. This pitfall can be avoided by shifting the I and Q samples by a number of spaces corresponding to their navigation data transition (i.e. to start at a navigation bit transition). The number of possible K values lie between M to nM, where n is an integer (not all possible values of n will give a low standard deviation (σ)). For example, if we have a 2 s data set with 1 ms accumulations, we will have 2000 I and Q samples. Choosing $M = 5$, K can be between $5 - 2000$. If we choose $K = 20$ ms, then the number of C/N_0 values obtained from the algorithm would be $j = 2000/K = 100$, each is based on $h = K/M = 4$ NP values with $M = 5$ samples per NP estimate.

III. Patch Antenna Array

The L1 patch antenna was designed using a Method of Moments field solver (FEKO). The patch was designed on an FR-4 substrate with ϵ_r=4.2, and hight from GND plane of 1.58 mm. The patch was made of copper with σ=5.8$\times 10^7$ S and a probe feed coming through the GND plane. The feed location was optimized to match the patch impedance at resonance to that of the probe feed. The dimensions of the patch were derived from the equations in [11] and optimized using the FEKO package. The designed patch had the dimensions in Table I. The radiation pattern obtained from the field solver was formatted into azimuth and elevation complex components, and imported to MATLAB for processing. The single patch E-field was multiplied by the planar array factor to yield

TABLE I

Patch antenna dimensions.

Length	Width	Thickness	Probe Via
45 mm	45 mm	1 Oz. Copper	2.5 mm
GND plane		Probe feed location	
90 × 90 mm^2		x=12mm, y=22mm	

the antenna array radiation pattern, which was then used in a NS approach denoted as the beam subtraction method (BS). The planar array of $N \times M$ patch elements was simulated by utilizing the array pattern multiplication method [11]. The array factor was found from:

$$AF(\theta,\phi) = \left[\frac{1}{N}\frac{sin(\frac{N}{2}\psi_x)}{sin(\frac{\psi_x}{2})}\right]\left[\frac{1}{M}\frac{sin(\frac{M}{2}\psi_y)}{sin(\frac{\psi_y}{2})}\right] \quad (10)$$

where,

$$\psi_x = K_o dx sin\theta cos\phi + \beta_x$$

$$\psi_y = K_o dy sin\theta sin\phi + \beta_y$$

and K_o is the wave number, dx, dy are inter-element spacings in x and y directions respectively. They were set to $0.5\lambda_{L1}$. The beam is steered by changing the values of β_x and β_y. Uniform amplitude was fed to all patches.

The BS method is performed by subtracting the complex 3D pattern of the patch array from the actual 3D MLA pattern. Then the gain is obtained from the resultant $\overrightarrow{E}$ field via $Gain = 20 \times log_{10}(|\overrightarrow{E}|)$. Since the MLA is vertically polarized (VP), more rejection will be obtained if the antenna array was polarized the same way. Using circularly polarized (CP) elements will degrade the rejection capability since the total field will look like:

$$\overrightarrow{E}^{BS}_{TotalCP}(\theta,\phi) =$$

$$\left[\overrightarrow{E}^{MLA}_{v}(\theta,\phi) - \overrightarrow{E}^{array}_{v}(\theta,\phi)\right] - j\overrightarrow{E}^{array}_{h}(\theta,\phi) \quad (11)$$

where, $\overrightarrow{E}_v$ is the vertical electric field component, and $\overrightarrow{E}_h$ is the horizontal component. The horizontal array component will yield a residual error. Since the inter-element distance between two adjacent patches is $\lambda_{L1}/2$, then the effect of coupling would be minimal.

IV. Simulation Results

The stored GPS data was collected using a setup with an RF front-end gain of 90.6 dB, and a roof-top antenna with a gain $3dB$. It was sampled at 17.357737 MHz with an IF of 4.333211 MHz. PRN 22 is the one collected and processed (selected randomly). The data set contained 2 s of data samples ($\sim$ 55 MB of HD space). The planar antenna beam was steered into a specified arbitrary location in elevation and azimuth where the RFI is assumed. This is performed via changing inter-element phase excitation (β_x, β_y). Some techniques capable of identifying the location of the interferer using antenna arrays has already been proposed like the MUSIC algorithm. It is assumed that the RFI location is known in this effort. The output of the MLA feed network is combined in baseband (can be performed in RF by utilizing power dividers and combiners) with the output of the planar array to create the null in the desired elevation angle. The effect of the number of antenna elements along with the polarization were presented in [7].

Fig. 1 shows the simulation block diagram. The resultant pattern from NS is used to select the corresponding gain in the interference and signal directions, then the data and noise are combined (in digital IF) and passed to the software GPS receiver. Acquisition is performed to find the Doppler and code shifts. Tracking is performed using 1 ms accumulation intervals for the I and Q samples. The samples are then passed to the C/N_0 algorithms. Two types of interference sources are simulated; Continuous Wave (CW) and Pulses Per Second (PPS). The former is just a single tone interferer with a certain frequency and power. The later is a pulsed wave with a certain power and duty cycle. The NS mechanism will create a null in the interferer direction to reduce its effect on the receiver C/N_0 estimation. This is done for a certain elevation and azimuth directions. Fig. 2 shows the gain pattern in dB of the original MLA and the resultant pattern after creating a null in $\theta = 20^o$ and $\phi = 165^o$, which specify the interferer direction. The original signal is coming from $\theta = 20^o$ and $\phi = 45^o$. The polar plot shows a null in elevation and its opposite direction, this comes from the AF of the array. Although this will affect any desired signals in the vicinity of the nulls in both directions, the two nulls have a finite width in azimuth and elevation that makes this method superior to that of altering the complex feeds of the antenna, that will severely affect the availability of the system. This was discussed in [7]. Fig. 3 shows the original C/N_0 estimates

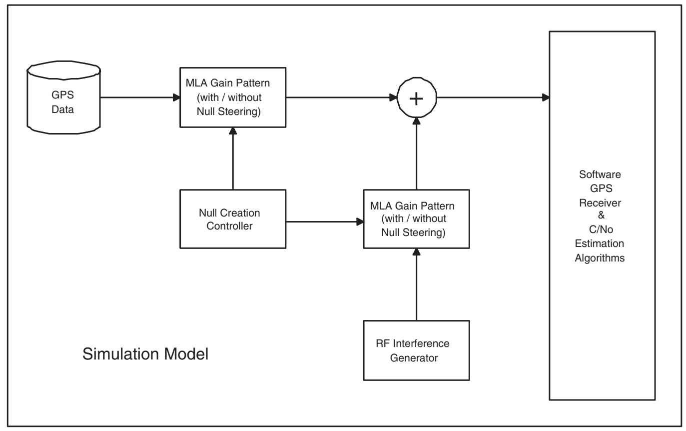

Fig. 1. C/No estimation simulation block diagram.

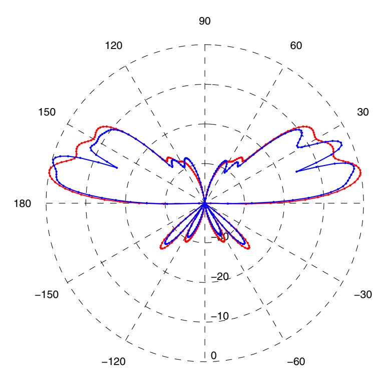

Fig. 2. MLA normalized gain pattern in dB before (red) and after (blue) null steering is incorporated. The null is created in the $\theta = 20^o$ elevation angle.

from both methods the VSM and PRM along with an actual measured level using a high-end survey grade GPS receiver. The measured/estimated C/N_0 values were recorded every

100 ms. The average measured value was 49 dB-Hz, which was lower by about 1 dB-Hz than the estimated means of both methods. This difference is believed to be attributed to added noise in hardware processing compared to SW code with high precision. In the same plot the estimates in the presence of a CW interference with a level of -84.6 dBW at the antenna input (amplitude of 4V at the baseband injection point in Fig. 1), and frequency of 4.49 MHz are shown. The CW is coming from $\theta = 20^o$ and $\phi = 165^o$ while the GPS signal is coming from $\theta = 20^o$ and $\phi = 45^o$. The presence of the RFI degrades the C/N_0 estimates for both the VSM and PRM by about 7 dB-Hz, with a mean difference of 0.8 dB-Hz between the two (red curves). When NS was incorporated, the estimates improved by about 5 dB-Hz for both, and then mean difference between the two estimates dropped to 0.2 dB-Hz (black curves). The dependence of C/N_0 estimation on the CW interference strength is illustrated in Fig. 4. The power of the CW is calculated at the baseband injection point (Fig. 1). Fig. 5 shows the effect of CW elevation angle variation on the estimation of the C/N_0 for both PRM and VSM. The angle variation was between $[5^o - 30^o]$ since the MLA covers from the horizon (i.e. 0^o) up to 35^o in elevation. The use of NS improves the estimates by a maximum of 11 dB-Hz when the CW is at $\theta = 15^o$. The reader should note that the narrow-band interference effect can also be reduced with the use of post-correlation analog-to-digital converters (ADC). Here, the interference will get spread in frequency, and its effect would be reduced. Direct conversion (zero IF) receivers will pass such narrow band interference, and its effect would be more profound. The dependance of the estimates on the frequency of the CW interference is shown in Fig. 6. The closer the CW frequency to that of the receiver IF, the stronger its effect and the lower the C/N_0 estimate becomes. Yet the use of NS boosts the C/N_0 by about 6 dB-Hz. More degradation would occur if the CW overlaps a C/A code spectral line. The second type of interference investigated is the PPS, with a duty cycle of 50%. Three factors are highlighted, the effect of power level, angle of arrival, and frequency. Fig. 7 illustrates the degradation in the estimate with increased PPS power levels.

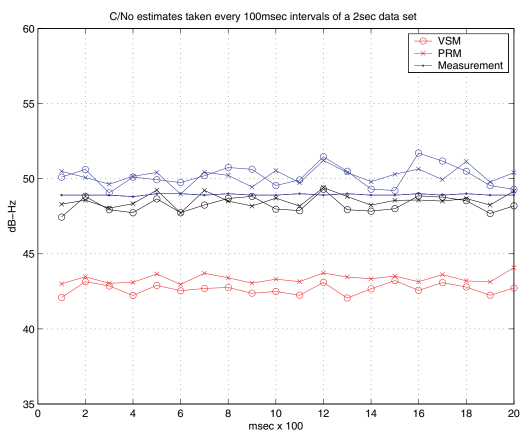

Fig. 3. C/N_0 estimation, without (blue)/with (red) CW interference (P_{cw}=9 dB,f_{cw}=4.49 MHz). The estimation with null steering incorporation is in (black).

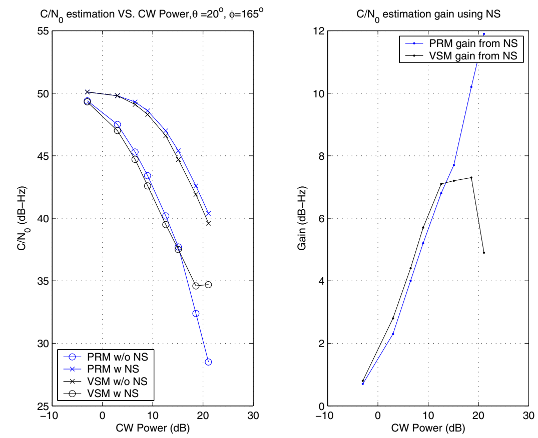

Fig. 4. C/N_0 estimation vs. CW power levels (f_{cw}4.49 MHz), with (o) and without (x) NS for both PRM and VSM.

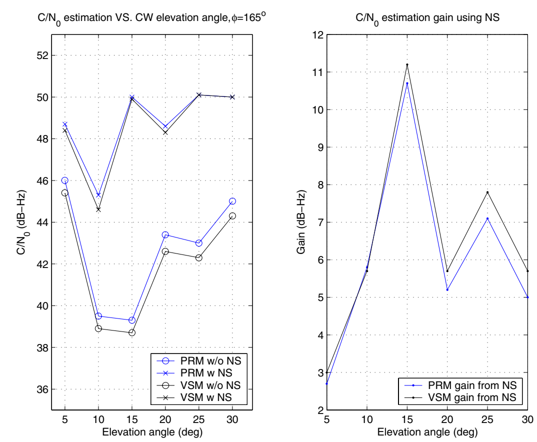

Fig. 5. C/N_0 estimation vs. CW elevation angles (P_{cw}=9 dB,f_{cw}4.49 MHz), with (o) and without (x) NS for both PRM and VSM.

Fig. 8 shows the dependance on the PPS angle of arrival in elevation, and Fig. 9 shows the degradation due to various frequency values. The figures show that more degradation occurs at lower elevation angles along with closer frequencies to IF. Although it seems that closer to IF PPS frequencies gives more degradation, the reader should note that the PPS in an actual receiver will be filtered, and for such an IF, only the fundamental component would be the one to survive (here PPS was injected at digital IF without filtering, only was scaled with antenna gain). Lower frequency PPS signals with higher powers are the ones of concern for GPS receiver jamming (i.e. radar signals). The use of NS will improve the C/N_0 estimates by at least 7 dB-Hz at the points of low estimates. This gives NS a great advantage for use in such an architecture to combat RFI. If the RFI comes from the same direction as that of the desired GPS signal, then NS would suppress both the RFI and the desired signal. But even with such a situation, because of the structure of GPS, usually there are more than 6 satellites visible at most locations, and suppressing one, will not affect the accuracy much. The presence of multiples interferes would dictate a more complex structure to create multiple beams simultaneously [12].

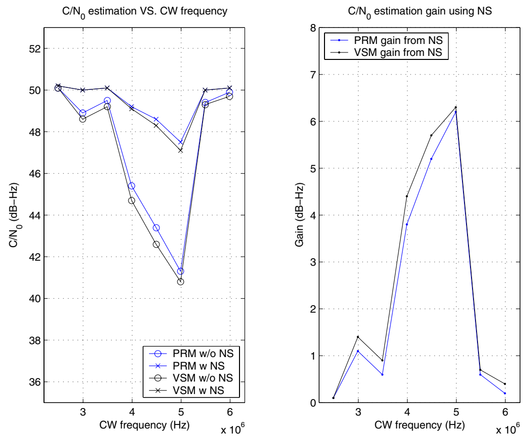

Fig. 6. C/N_0 estimation vs. CW frequency (P_{cw}=9 dB), with (o) and without (x) NS for both PRM and VSM.

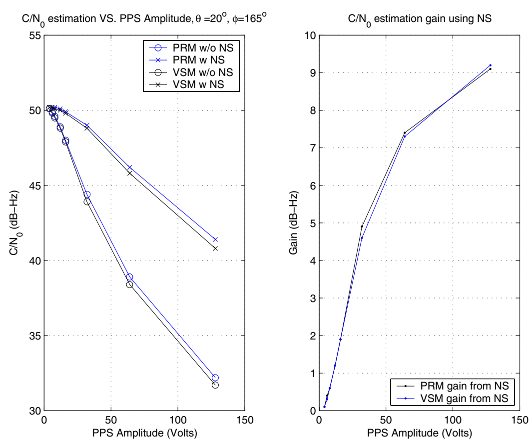

Fig. 7. C/N_0 estimation vs. PPS voltage (f_{PPS}=1500 Hz), with (o) and without (x) NS for both PRM and VSM.

V. Conclusions

The estimation of C/N_0 with the presence of two RFI source types in a GPS SW receiver was investigated. The effect of utilizing NS using a planar patch antenna array on the C/N_0 estimation with RFI presence was demonstrated via simulations. This NS approach will have minimal affect on the MLA PCV and GD curves. For both RFI types, the higher the power the more the impact on the C/N_0 estimates and more degradation. Lower elevation angles and closer to IF frequencies had more impact on the C/N_0 estimates. The utilization of NS improved the C/N_0 estimates by $\sim$11 dB-Hz for low elevation and high RFI power levels.

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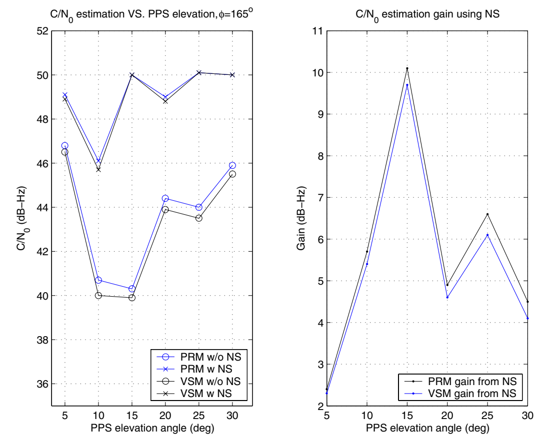

Fig. 8. C/N_0 estimation vs. PPS elevation angle (V_{PPS}=32V,f_{PPS}=1500 Hz), with (o) and without (x) NS for both PRM and VSM.

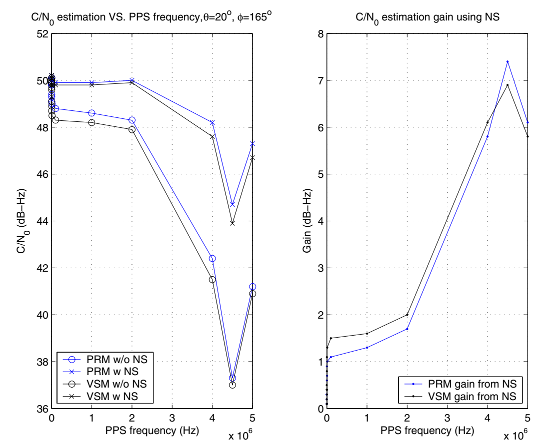

Fig. 9. C/N_0 estimation vs. PPS frequency (V_{PPS}=4 V), with (o) and without (x) NS for both PRM and VSM.

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