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Non Data Aided Parameter Estimation for Multi-User ARGOS Receivers

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Abstract—In this paper, parameter estimators are analyzed in the context of Successive Interference Cancellation (SIC) receivers for the ARGOS system. A Non Data Aided (NDA) feed forward estimator is proposed for the amplitude and the carrier phase parameters. Time delays are assumed to be known. A Window Accumulator (WA) is used to reduce the influence of the additive noise. In the presence of frequency offset, the window length L cannot be chosen arbitrarily large but an optimal length L_{opt} can be determined. However, because the estimator induces a different optimal length for each parameter, a trade-off must be made. We show that a window length of around 35 samples induces mean square errors (MSEs) lower than 0.012 for both parameters. The MSE of the proposed estimator is also compared to the Modified Cramér Rao Bound (MCRB).

I. INTRODUCTION

ARGOS [1] is a global satellite-based location and data collection system dedicated for the study of the environment. User beacons [2], transmit data messages to ARGOS satellites that receive, decode, and forward the messages to ground stations. Each beacon has a different carrier frequency located within a 110 kHz band. Due to the relative motion between satellites and beacons, beacon signals are affected by both a different Doppler shift and a different propagation delay. The Doppler shifts are used to estimate the position of the beacons but they also have a negative impact on the system performance since they cause the overlapping of different beacon signals in the frequency domain. Thus, ARGOS satellites receive overlapping signals in both frequency and time domains. This kind of Multiple Access Interference (MAI) can be reduced using Multi-User Detection (MUD) techniques. When two asynchronous beacon signals are received at the ARGOS satellites [3], SIC receivers have been shown to offer a good optimality-complexity trade-off compared to Maximum Likelihood (ML) receivers. In an ARGOS SIC receiver, users are successively decoded, regenerated, and subtracted from the received signal.

In the context of SIC receivers, the estimation of signal parameters is a key issue since erroneous parameter estimates degrade the performance in terms of Bit Error Rate (BER). This impact has been studied in [4] but no parameter estimator has been studied in details. There are four signal parameters for each user: the time delay, the carrier frequency, the amplitude, and the carrier phase. In this paper, we assume perfect time synchronization and imperfect frequency synchronization. So we concentrate on the estimation the amplitude and the phase

in the presence of a frequency offset. This frequency offset is the residual carrier frequency at the output of the frequency estimator. A joint estimation of the amplitude and the carrier phase for multi-user receivers have been proposed in [5], [6]. The joint estimation is recursive, i.e., the estimation is performed for each symbol using previous estimates. The procedure is repeated until convergence and thus requires a large number of symbols [6]. This approach is not suited for our study since it would require the addition of overhead bits and it is not our goal to change the ARGOS frame structure. Instead, we propose an estimation technique that takes into account two design constraints. First, the estimation is performed without adding overhead bits to ARGOS messages. Second, parameter estimates are computed without any knowledge of data symbols. In this context, a NDA feed forward estimator is selected. Among NDA estimators for single user receivers [7], the Viterbi & Viterbi algorithm has been shown to offer a good optimality-complexity trade-off [8]. More interestingly, it provides a estimator variance that is close to the Cramér-Rao bound. This algorithm uses a WA, i.e., each data block is divided into smaller sub-blocks and an estimate is generated for each sub-block [9]. The impact of both frequency offset and window length on the performance of the proposed estimator is studied. For each frequency offset, the optimal length L_{opt} is determined. The window length is optimal in the sense that it minimizes the MSE. Then, the minimal MSE is compared to the MCRB [10], [11] and we show that the achieved MSE falls within the error margins that have been defined in [4].

The rest of the paper is organized as follows. Section II introduces the ARGOS SIC receiver, as well as the mathematical derivations. In section III, the Viterbi & Viterbi algorithm for the ARGOS system is presented. In section IV, simulation results are presented. Finally, we conclude in section V.

II. ARGOS SIC RECEIVER ARCHITECTURE

We follow the lines in [3], [4] and consider the transmission of two ($K = 2$) asynchronous ARGOS users. According to ARGOS traffic patterns, allowing the reception of two simultaneous beacons already significantly increases the number of beacons processed by unit of time. The satellite channel is assumed to be flat-fading, and beacon signals are affected by time delays and Doppler shifts. The base band received signal

$r(t)$ can be written as

$$r(t) = \sum_{k=1}^K r^{(k)}(t) + n(t) \quad (1)$$

where $k \in [1, K]$ and $r^{(k)}(t)$ is the signal with the k^{th} highest received power and is written as

$$\begin{aligned} r^{(k)}(t) &= \sum_{m=0}^{M-1} A^{(k)} s^{(k)}(m) h(t - mT - \tau^{(k)}) \\ &\times \exp[j(2\pi f^{(k)} t + \theta^{(k)})] \end{aligned} \quad (2)$$

where M is the number of symbols per user message, $A^{(k)}$, $\tau^{(k)}$, $f^{(k)}$, $\theta^{(k)}$, and $s^{(k)}(m)$ denotes respectively the amplitude (fading gain), the time delay, the carrier frequency, the initial carrier phase, and the m^{th} transmitted symbol of the signal with the k^{th} highest power, $h(t)$ is the unit energy waveform with a value of $1/\sqrt{T}$ over one symbol interval $[0, T]$ where T is the symbol period, and $n(t)$ is a circularly symmetric complex Gaussian noise with variance $\sigma^2 = 2N_0$. The symbols $s^{(k)}(m)$ are BPSK modulated in accordance with the ARGOS system requirements. The channels parameters are assumed constant over the duration of an ARGOS message. The receiver is assumed to be time synchronized to the signal with the highest power. In other words, we have $\tau^{(1)} = 0$ and $\tau^{(2)} = \Delta\tau$. So, the time difference $\Delta\tau$ between the two received signals can be positive or negative. The signal with the highest power, $r^{(1)}(t)$, is demodulated with a conventional detector provided that its power is higher than a detection threshold [4]. The output $y^{(1)}(m)$ of the matched filter, sampled at time $mT + \tau^{(1)}$, is given by

$$\begin{aligned} y^{(1)}(m) &= \int_{mT + \tau^{(1)}}^{(m+1)T + \tau^{(1)}} r(t) \exp[-j2\pi \hat{f}^{(1)} t] \\ &\times h^*(t - mT - \tau^{(1)}) dt \end{aligned} \quad (3)$$

where $\hat{f}^{(1)}$ denotes the estimated value of $f^{(1)}$ and the upper script * denotes the complex conjugate. Equation (3) is written as [3]

$$\begin{aligned} y^{(1)}(m) &= A^{(1)} s^{(1)}(m) \rho^{(1),(1)}(m, m) \\ &+ A^{(2)} s^{(2)}(m) \rho^{(1),(2)}(m, m) \\ &+ A^{(2)} s^{(2)}(m-1) \rho^{(1),(2)}(m, m-1) \\ &+ n^{(1)}(m) \end{aligned} \quad (4)$$

where $n^{(1)}(m)$ denotes the noise at the output of the matched filter, sampled at time $mT + \tau^{(1)}$. The coefficients $\rho^{(1),(1)}(m, m)$, $\rho^{(1),(2)}(m, m)$, and $\rho^{(1),(2)}(m, m-1)$ denote the time dependent cross correlation factors given by [3]

$$\begin{aligned} \rho^{(1),(1)}(m, m) &= \exp[j(2\pi \delta f^{(1)} mT + \theta^{(1)})] \\ \rho^{(1),(2)}(m, m) &= \frac{T - \Delta\tau}{T} \times \text{sinc}[\Delta f (T - \Delta\tau)] \\ &\times \exp[j\pi \Delta f (\Delta\tau + 2mT + T)] \\ \rho^{(1),(2)}(m, m-1) &= \frac{\Delta\tau}{T} \times \text{sinc}(\Delta f \Delta\tau) \\ &\times \exp[j\pi \Delta f (\Delta\tau + 2mT)] \end{aligned} \quad (5)$$

where $\delta f^{(1)}$ denotes the frequency offset between $f^{(1)}$ and $\hat{f}^{(1)}$. The estimation of $f^{(1)}$ is performed as follows. A 8192-point FFT is applied to the first part of the ARGOS signals. This part consists of a non-modulated carrier signal that lasts 40 ms. We consider that the frequency offset is approximately 8192 times less than the estimated value, so we have that

$$\begin{aligned} \delta f^{(1)} &\ll |f^{(1)} - \hat{f}^{(1)}| \\ \Delta f &= f^{(2)} - \hat{f}^{(1)} \approx f^{(2)} - f^{(1)} \end{aligned} \quad (6)$$

where Δf denoted the frequency shift between the two users. The symbols $\hat{s}^{(1)}(m)$ for $m \in [0, M-1]$ are decoded first. Then, the signal parameters $\hat{A}^{(1)}$ and $\hat{\theta}^{(1)}$ are estimated in order to obtain a replica $\hat{r}^{(1)}(t)$. The estimated signal $\hat{r}^{(1)}(t)$ is then subtracted from the received signal $r(t)$. The resulting signal $r(t) - \hat{r}^{(1)}(t)$ is fed into the second conventional detector in order to demodulate the signal $r^{(2)}(t)$ with the lowest power and generate the samples $y^{(2)}(m)$ with $m \in [0, M-1]$. These samples are then processed in order to estimate the symbols $\hat{s}^{(2)}(m)$.

III. FEED FORWARD NDA ESTIMATOR

The main contribution of this section consists of showing the ability of the Viterbi & Viterbi algorithm to operate in the multi-user context of the ARGOS system.

A. Amplitude and Carrier Phase Estimation

First, the samples $y^{(1)}(m)$ in (4) are squared in order to eliminate the influence of the data symbols [9]. Then, the impact of the additive noise is reduced by using an averaging filter over a WA. Two types of WA could be considered here: the Block Window Accumulator (BWA) and the Gliding Window Accumulator (GWA). When a BWA is used, data messages are divided into blocks and a carrier phase estimate is obtained for each block. The only unbiased estimate is obtained for the data symbol in the middle of the block. When a GWA is used, a sliding window is shifted by one symbol at a time. A carrier phase estimate is obtained for each step. When the data blocks are composed of M symbols and the GWA is L symbols long, then the estimates for the $M-L$ most central symbols are unbiased [9]. In the absence of a frequency offset, both WAs achieve the same performance in terms of MSE. However, when the received samples are subjected to non-zero frequency offsets, BWA-based estimates become biased compared to GWA-based estimates [12]. Thus, we limit our study to GWA-based estimators. The averaging filter designed with a GWA of length L computes the arithmetic mean $E_L(u_0, [y^{(1)}(u)]^2)$ for $1 \leq u_0 \leq M-L+1$, where

$$E_L(u_0, [y^{(1)}(u)]^2) = \frac{1}{L} \sum_{u=u_0}^{u_0+L-1} [y^{(1)}(u)]^2 \quad (7)$$

Knowing that the correlation coefficients $\rho^{(1),(2)}(u, u)$ and $\rho^{(1),(2)}(u, u-1)$ of (5) have sinusoidal shapes with a non zero frequency shift Δf , their time average over a WA of length L is close to zero. This property results from the fact that two ARGOS users are always separated by a non zero

frequency shift Δf , due to the Doppler shifts affecting each user. Thus, (7) is written as [9]

$$E_L(u_0, [y^{(1)}(u)]^2) \approx [A^{(1)}]^2 E_L(u_0, [\rho^{(1),(1)}(u, u)]^2) + E_L(u_0, [n^{(1)}(u)]^2) \quad (8)$$

where

$$E_L(u_0, [\rho^{(1),(1)}(u, u)]^2) = \frac{1}{L} \sum_{u=u_0}^{u_0+L-1} [\rho^{(1),(1)}(u, u)]^2$$

$$E_L(u_0, [n^{(1)}(u)]^2) = \frac{1}{L} \sum_{u=u_0}^{u_0+L-1} [n^{(1)}(u)]^2 \quad (9)$$

The first term on the right side of (8) denotes the useful component and the second term denotes a noise component at the output of the averaging filter. Equation (8) shows the similarity between the Viterbi & Viterbi algorithm in a single user context and the multi-user context of the ARGOS system, in the presence of a non zero frequency shift Δf [13]. Finally, the amplitude and the carrier phase estimates of the signal with the highest power $r^{(1)}(t)$ are given by [13]

$$\hat{A}^{(1)} = \sqrt{\text{Re}^2(Z) + \text{Im}^2(Z)}$$

$$\hat{\theta}^{(1)} = \frac{1}{2} \arg[\text{Re}(Z) + j \text{Im}(Z)] \quad (10)$$

where Z denotes the result number of (7), $\text{Re}(Z)$ and $\text{Im}(Z)$ denote respectively the real part and the imaginary part of Z , and $\arg(Z)$ denotes the argument function with

$$-\pi \leq \arg(Z) \leq \pi \quad (11)$$

Note that the parameter estimates $\hat{A}^{(2)}$ and $\hat{\theta}^{(2)}$ of the signal with the lowest power $r^{(2)}(t)$ are obtained in the same manner as in (10) by replacing $y^{(1)}(m)$ with $y^{(2)}(m)$. The samples $y^{(2)}(m)$ denote the matched filter outputs of the second conventional detector of the SIC receiver (see Section II). The carrier phase estimates $\hat{\theta}^{(1)}$ and $\hat{\theta}^{(2)}$ jump by $\pm\pi$ radians every time the phase angle trajectory of the averaging filter output samples crosses the modulo 2π boundary of the $\arg(z)$ function. Thus, the phase estimates are restricted to the interval $[-\pi/2, \pi/2]$ and they suffer from a phase ambiguity of $\pm\pi$. This ambiguity is correctly resolved by unwrapping the phase estimates [14].

B. Impact of Frequency Offset on the WA Length

We assume that the carrier phase $\theta^{(1)}$ is constant over the observation interval, i.e., the duration of an ARGOS data frame. We are searching for the optimal length L_{opt} of the WA. This optimal length minimizes the MSE of the phase estimator. According to (8) and when the frequency offset is zero ($\delta f^{(1)} = 0$), we choose the largest possible value for L_{opt} , i.e., the length of the data message. Hence, noise effects are reduced. Otherwise, when the frequency offset is considered to be zero ($\delta f^{(1)} \neq 0$), the correlation coefficient $\rho^{(1),(1)}(u, u)$ in (8) varies linearly with m . An optimal length L_{opt} is achieved when the averaged term $E_L(u_0, [\rho^{(1),(1)}(u, u)]^2)$ is zero. Thus, a different optimal length is achieved for each frequency offset $\delta f^{(1)}$.

IV. SIMULATION RESULTS

We consider the reception of two active users ($K = 2$). The Power Separation Ratio (PSR) between the two signals is defined as $PSR = E_{b(1)}/E_{b(2)}$ where $E_{b(k)}$ is the mean energy received per bit, in the single user case, for user k with $k \in [1, K]$. The performance of the Viterbi & Viterbi estimator is evaluated in terms of $E_{b(k)}/N_0$ denoted by SNR_k , the PSR, the relative frequency shift $\Delta f/R_b$ between the two received signals, the relative frequency offset $\delta f^{(1)}/R_b$ and the WA length L , where R_b denotes the symbol data rate. We consider a maximum value of $3 \cdot 10^{-3}$ for $\delta f^{(1)}/R_b$ at the output of the frequency estimator¹.

A. Performance of the Amplitude Estimator

Fig. 1 presents the amplitude MSE as a function of the relative frequency shift $\Delta f/R_b$. The proposed estimator offers good performance for values of $\Delta f/R_b$ higher than 0.025. For values of $\Delta f/R_b$ smaller than 0.025, the time average of the coefficients $\rho^{(1),(2)}(u, u)$ and $\rho^{(1),(2)}(u, u-1)$ are not negligible and the approximation in (8) does not hold. Subsequently, we assume that $\Delta f/R_b \geq 0.025$.

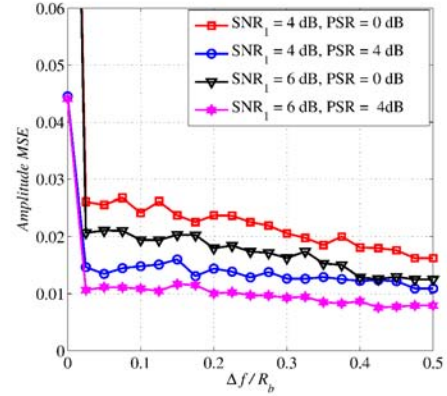


Fig. 1. Amplitude MSE for the signal with the highest power $r^{(1)}(t)$ as a function of $\Delta f/R_b$ for $L = 40$ and $\delta f^{(1)}/R_b = 1.10 \cdot 10^{-3}$.

Fig. 2 presents the amplitude MSE as a function of both the WA length L and the relative frequency offset $\delta f^{(1)}/R_b$. An optimal length L_{opt} minimizing the MSE of the amplitude estimator exists for each value of $\delta f^{(1)}/R_b$. In particular, the optimal length L_{opt} is equal to 32 for $\delta f^{(1)}/R_b = 3 \cdot 10^{-3}$, and 85 for $\delta f^{(1)}/R_b = 1.10 \cdot 10^{-3}$. When $\delta f^{(1)}/R_b = 0$, the MSE for the amplitude estimator decreases monotonically with the length of the window. Now, we compare the MSE of the amplitude estimator with the MCRB. We use the MCRB because analytical expressions of this bound are available. The results show that the performance of the amplitude estimator approaches the MCRB (see Fig. 3 and Fig. 4). However,

¹The bandwidth is 110 kHz, the spectral analysis performed by the ARGOS receiver uses a 8192-point FFT, and $R_b = 400$ bits/s.

²The relative frequency shifts are limited to $[0, 4]$ since the power of a bi-phase signal is concentrated in the $[-4R_b, 4R_b]$ band.

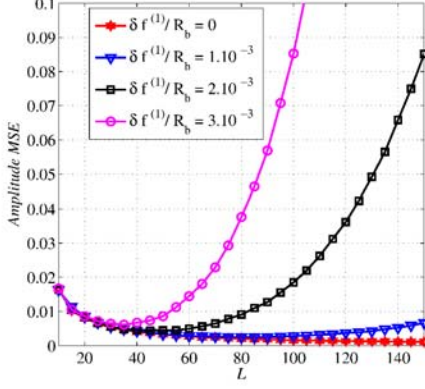


Fig. 2. Amplitude MSE for the signal with the highest power $r^{(1)}(t)$ as a function of the window length L for $SNR_1 = 8$ dB, $PSR = 0$ dB, and $\Delta f/R_b = 0.75$.

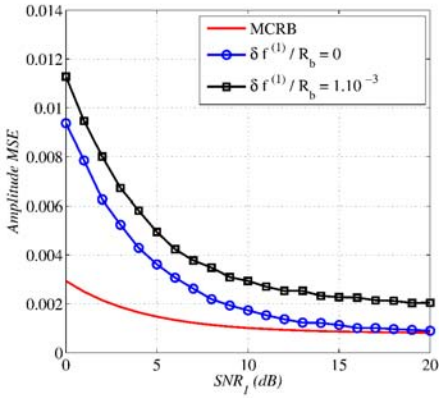


Fig. 3. Amplitude MSE for the signal with the highest power $r^{(1)}(t)$ as a function of SNR_1 for $L = 50$, $PSR = 4$ dB, and $\Delta f/R_b = 0.75$.

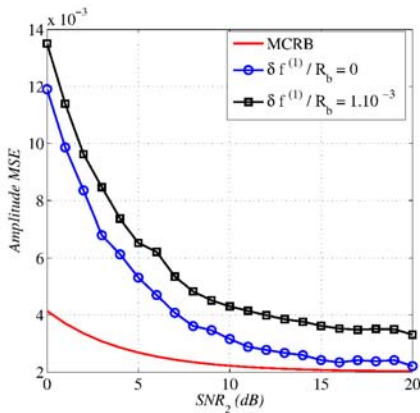


Fig. 4. Amplitude MSE for the signal with the lowest power $r^{(2)}(t)$ as a function of SNR_2 for $L = 50$, $PSR = 4$ dB, and $\Delta f/R_b = 0.75$.

it is also shown that the amplitude estimator is biased for

$\delta f^{(1)}/R_b \neq 0$. Further studies should provide a means to remove this bias.

B. Performance of the Carrier Phase Estimator

The carrier phase estimate $\hat{\theta}^{(1)}$ is unwrapped after the estimation procedure [14], so that the resulting phase $\check{\theta}^{(1)}$ is in $[-\pi, \pi]$. A phase ambiguity of $\pm 2\pi$ occurs but this ambiguity is not critical since it does not affect the symbol decision. Similarly to amplitude estimation, Fig. 5 presents the effects of the WA length L and the relative frequency offset $\delta f^{(1)}/R_b$ on the carrier phase MSE. The results show the existence of an optimal length L_{opt} for a given value of the relative frequency offset. However, we note that this optimal length is twice the one obtained for the amplitude estimator. For example, for a $\delta f^{(1)}/R_b$ of 3.10^{-3} , the optimal length to minimize the amplitude MSE is 32, however it is of 63 to minimize the carrier phase MSE. This result comes from the fact that the

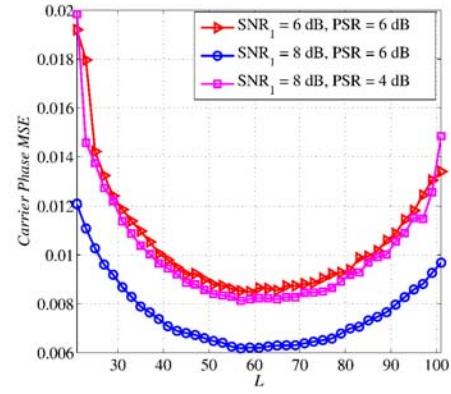


Fig. 5. Carrier phase MSE for the signal with the highest power $r^{(1)}(t)$ as a function of the window length L for $\delta f^{(1)}/R_b = 3.10^{-3}$ and $\Delta f/R_b = 0.25$.

signal amplitude is estimated once for all the window of length L . The presence of a relative frequency offset $\delta f^{(1)}/R_b$ affects each sample m of the window with the value $(\delta f^{(1)}m)/R_b$. Thus, the maximum offset value is $(\delta f^{(1)}L)/R_b$ at the edge of the window. However, for the phase estimation, the estimates holds for the sample in the middle of the window. This sample is affected by a maximum offset value of $(\delta f^{(1)}L)/2R_b$. This explains our results. Finally, the MSE of the carrier phase estimator is compared to the MCRB. The results are shown in Fig. 6 and Fig. 7. The results show that the performance of the phase estimator approaches also the MCRB values. Thus, the Viterbi & Viterbi estimator can be used for carrier phase estimation in the ARGOS system.

The optimal length L_{opt} is not the same when considering the amplitude estimator or the carrier phase estimator. So another approach is needed to choose the most appropriate window length L . When L is in the interval $[30, 40]$, the amplitude MSE is lower than 0.01 and the carrier phase MSE is lower than 0.012, for any value of the relative frequency offset $\delta f^{(1)}/R_b$. An MSE value of 0.01 for the amplitude estimator has been

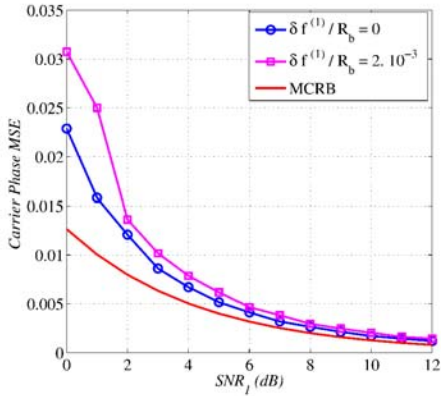


Fig. 6. Carrier phase MSE for the signal with the highest power $r^{(1)}(t)$ as a function of SNR_1 for $L = 51$, $PSR = 4$ dB, and $\Delta f/R_b = 0.75$.

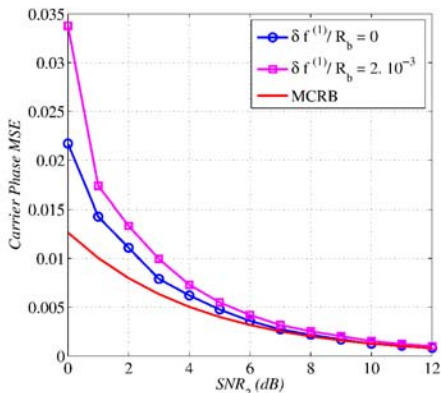


Fig. 7. Carrier phase MSE for the signal with the lowest power $r^{(2)}(t)$ as a function of SNR_2 for $L = 51$, $PSR = 0$ dB, and $\Delta f/R_b = 0.875$.

shown to induce an additional loss on the received SNR that is lower than 0.05 dB, i.e., the received SNR is degraded by a factor lower than 0.05 dB when the MSE of the amplitude estimator is lower than 0.01 [4]. Similarly, an MSE value of 0.012 for the carrier phase estimator induces an additional loss on the received SNR that is lower than 0.05 dB [15]. These SNR losses are much lower than the ones induced by imperfect time delay estimation [4].

V. CONCLUSION

In this paper, we propose an NDA feed forward estimator for the ARGOS system. The proposed estimator is based on the Viterbi & Viterbi algorithm. This algorithm has been originally designed for a single user scenario, and it has been adapted here to the multi-user case of the ARGOS system. The proposed algorithm estimates the amplitude as well as the carrier phase of each user, in the presence of a frequency offset. A GWA is used to obtain unbiased estimates of the symbols. The estimator uses a WA to reduce the influence of the additive noise. However, in the presence of

a given frequency offset, increasing the WA length reduces also the useful component, so an optimal length L_{opt} should be determined. Because the two estimators provide different L_{opt} , a trade-off is made: the window length is chosen in the interval [30 40] so that MSEs on both estimators are lower than 0.012 and the induced degradation on the received SNR is lower than 0.05 dB. Simulation results have been obtained assuming perfect estimation of the time delays. The results show that the performance of the estimator in terms of MSE approaches the MCRB. Thus, the Viterbi & Viterbi estimator can be implemented for both amplitude and carrier phase estimation for the ARGOS system.

Time delay and carrier frequency estimators should now be considered in order to complete this study. This step is critical since it should ensure the good operation of the estimator proposed in this paper. In particular, the carrier frequency estimator should ensure that frequency errors are in the range of the frequency offset values given in this paper. This part is left for future work.

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