Frame Acquisition for Continuous and Discontinuous Transmission in the Forward Link of Ka-band Satellite Systems

Giovanni E. Corazza, Raffaella Pedone, and Marco Villanti

DEIS/ARCES - University of Bologna
Viale Risorgimento, 2 - 40136 Bologna, Italy
{gecorazza, rpedone, mvillanti}@deis.unibo.it

Abstract

This paper presents a novel frame acquisition design procedure for the forward link of Ka-band satellite systems. Multi- and single-dwell procedures are considered for continuous and discontinuous transmission, respectively. To cope with large frequency errors, the detector employs post detection integration and is based on the threshold crossing (TC) criterion. In particular, non coherent and differential post detection techniques are considered. The design procedure applies to TDM/TDMA networks in general, and it is applied here to the forthcoming DVB-S2 standard.

I. Introduction

The efficient provision of broadband interactive services and fast Internet access are two of the main objectives of the ACM (Adaptive Coding and Modulation) project [1] funded by the European Space Agency (ESA), in accordance with the ongoing development of the next generation TDM-TDMA standard Digital Video Broadcasting (DVB-S2/RCS), which is expected to have a very significant market impact in the medium term. To maximize the system capacity, the ACM air interface foresees a large set of possible modulation and coding schemes, which are selected adaptively according to the channel propagation conditions [2]. The enhanced robustness and flexibility in waveform selection carries as a natural consequence the fact that the effective signal-to-noise ratio can be extremely low, leading to the increased criticality in synchronization and parameter estimation, which then become key components in the critical path to the successful implementation of the standard.

This paper addresses the problem of frame synchronization, which for a TDM/TDMA system is a pre-requisite for correct data demodulation and decoding. To the best of our knowledge, this is one of the first instances in which this problem is tackled systematically for both continuous and discontinuous transmission. In particular, discontinuous transmission is experienced when beam hopping is employed, i.e. the transmission over a specific beam is interrupted after one or more frames (hop interval) in favor of different spot beams. Beam hopping can be useful to cope with non-uniform and dynamic traffic conditions. Several frame acquisition procedures have been studied, all based on the detection of an appropriate unique word (UW) marking the start of the frame. In this paper, the focus is on the threshold crossing (TC) criterion [3] [4], and in particular both single (1TC) and three-dwell (3TC) acquisition procedures are considered. 1TC is used in beam hopping for terminal start-up, i.e. for initial synchronization with discontinuous transmission, where one-shot acquisition must be performed. Differently, a multi-dwell procedure can be employed when continuous transmission is foreseen. In this case, the acquisition procedure can be completed over several frames in order to improve the overall performance, introducing a verification phase in cascade to the search mode. In particular, 3TC with immediate rejection (i.e. the acquisition procedure is immediately restarted when the UW is not detected during the verification phase) is considered. An additional scenario, beam hopping in hot start operating mode, has been addressed in [5]. In this case, only a limited uncertainty region needs to be explored, the UW presence is granted and its exact position must be detected.

A large frequency offset usually affects the frame synchronization process, as it is normally the first operation to be performed. Thus, to cope with the consequent degradation of the coherent integration, post detection integration (PDI) can be introduced [6], [7], [8]. Here, both non coherent (NCPDI) and differential post detection integration (DPDI) are considered. DPD1 can in general provide better performance over the classic NCPDI solution, for which we present a fully analytical design procedure. This completely analytical characterization
is one of the main achievements of the paper, as it allows for an effective parameter optimization to minimize
the overhead introduced by the UW.

The analytical model for both multi- and single-dwell procedures takes into account the frequency error
impact, residual out-of-sync autocorrelation, inter-symbol interference (ISI), and imperfect symbol synchroniza-
tion. Simulation results are also reported to validate the analytical model, and evaluate the DPDI performance.

II. NCPDI AND DPDI DETECTORS

The block diagrams for NCPDI and DPDI are reported in fig. 1. Post-detection integration serves to
counteract the effects of the residual frequency error, which can be large especially when initial synchronization
is performed upon terminal switch on. Indeed, the frequency offset requires to limit the coherent integration
length, and post detection accumulation is necessary to improve the signal-to-noise ratio.

In general, the received signal can be written as

\[ r(t) = s(t) + n(t) \]  

where \( s(t) \) is the transmitted signal and \( n(t) \) is the AWGN with two-sided power spectral density equal to \( N_0/2 \).

The transmitted signal is in the form

\[ s(t) = \sum_{k=-\infty}^{\infty} \sqrt{E_s} a_k p(t - kT_s)e^{j(2\pi \Delta f t + \phi)} \]  

where \( p(t) \) is the Squared Root Raised Cosine (SRRC) pulse waveform, \( T_s \) the symbol period, \( \Delta f \) the absolute
frequency error, and \( \phi \) the unknown phase. In TDM/TDMA transmission, each frame consists of a known UW
followed by data, so that we can also specialize \( a_k \) as

\[ a_k = \begin{cases} 
  c_k |k|_{L_F} & |k|_{L_F} \in [0, L_{UW} - 1] \\
  d_k & |k|_{L_F} \in [L_{UW}, L_F - 1] 
\end{cases} \]  

where \( L_F \) is the frame length, \( | \cdot |_\eta \) is the modulo \( \eta \) operation, \( L_{UW} \) is the UW length, \( c_k \) is the \( k \)th symbol of
the UW, and \( d_k \) is the \( k \)th random data symbol.

The received signal undergoes symbol matched filtering and sampling at the instants \((m + \Delta)T_s + \delta, m \in \mathbb{Z}\),
where \( \Delta \) is the integer timing misalignment between the transmitted signal and the received replica \((\Delta \in \mathbb{Z})\),
and \( \delta \) the fractional timing error \((\delta \in [-T_s/2, T_s/2])\). Therefore, we can express the \( m \)th sample as

\[ r_m = \sum_{k=-\infty}^{\infty} \sqrt{E_s} a_k e^{j(2\pi \Delta f [(m + \Delta)T_s + \delta])} R_p((m - k + \Delta)T_s + \delta) + n'_m \]  

where \( R_p(t) \) is the SRRC pulse autocorrelation function, and \( n'_m \) is the noise component at the output of the
filter, which is a complex Gaussian variable with zero-mean and variance equal to \( N_0 \).

Observe that the condition \( \Delta = 0 \) corresponds to the case of UW alignment (i.e. \( H_1 \) hypothesis), while the
condition \( \Delta \neq 0 \) corresponds to the opposite non-synchronous case (\( H_0 \) hypothesis).

Coherent correlation of the received samples over a locally generated UW segment produces

\[ x_i = \sum_{m = iM}^{(i+1)M-1} r_m c_{|m|_{L_F}}^{*} \]  

where \( M \) is the coherent integration length.

For NCPDI \([6]\) the detection variable, \( \Lambda \), to be compared with the threshold is given by

\[ \Lambda = \sum_{i=0}^{L-1} |x_i|^2 \]  

where \( L \) is the PDI length. In order to exploit the entire UW, the pair \((M, L)\) must obviously satisfy the design
constraint \( M \cdot L = L_{UW} \).

For DPDI \([7]\) the detection variable can be expressed as

\[ \Lambda = \sum_{i=1}^{L-1} |x_i x_{i-1}^*| \]
III. NCPDI DETECTION PERFORMANCE

Here, the analytical characterization is carried out for NCPDI, leaving to simulations the evaluation of the DPDI case. A fitting method is applied to the false alarm and missed detection probabilities of DPDI in order to optimize the overall performance in terms of mean acquisition time, as will be presented in sec.V.

Under the H₁ hypothesis, we can rewrite equation (5) by separating the desired term from the disturbances as

\[ x_i = \sqrt{E_s} R_p(\delta) \sum_{m=iM}^{(i+1)M-1} c_{m,L}^* c_{m,L} e^{j(2\pi f(mT_s+\delta)+\phi)} + \gamma_i + \eta_i \]  

(8)

where the first term is the desired component, \( \gamma_i \) is the inter-symbol interference (ISI) term

\[ \gamma_i = \sum_{m=iM}^{(i+1)M-1} \xi_m c_{m,L}^* \]

(9)

and \( \eta_i \) denotes the AWGN sample at the output of coherent correlation. The complex AWGN samples, \( \eta_i \), are i.i.d. Gaussian distributed complex variables with zero-mean and in-phase and quadrature variance equal to \( \sigma_n^2 = \frac{M N_0}{2} \).

Now, the problem is to characterize the ISI term. Considering the received data \( a_k \) as zero-mean uniformly distributed random variables and invoking the central limit theorem, the ISI contribution \( \xi_m \) can be modelled as a Gaussian random variable. As a consequence, the ISI component \( \gamma_i \) turns out to be zero-mean Gaussian distributed, with branch variance \( \sigma_{ISI}^2 \) equal to

\[ \sigma_{ISI}^2 = M E_s \sum_{k=-\infty}^{\infty} R_p^2(kT_s + \delta) \]  

(11)

Hence, the samples at the output of coherent correlation \( x_i \) are complex Gaussian variables with mean

\[ \mu_i = \sqrt{E_s} R_p(\delta) e^{j(\phi+2\pi f \delta)} \sum_{m=iM}^{(i+1)M-1} e^{j2\pi f m T_s} \]

(12)

and branch variance \( \sigma_{ISI}^2 = \sigma_{ISI}^2 + \sigma_n^2 \).

Because \( \Lambda \) is the sum of \( L \) squared modules of non zero-mean complex Gaussian random variables, it is distributed as a non-central chi-square random variable with \( 2L \) degrees of freedom, with non-centrality parameter \( d \), and variance of the composing real Gaussian variates equal to \( \sigma_x^2 \) [1]. Considering \( \Delta f T_s \ll 1 \), it can be shown [6] that

\[ d = LM^2 E_s \text{sinc}^2(M \Delta f T_s) R_p^2(\delta) \]  

(13)

Thus, the correct detection probability \( P_d \) is

\[ P_d = \int_{-\infty}^{\infty} p_{\Lambda | H_1} (\Lambda) d\Lambda = Q_L \left( \frac{\sqrt{d}}{\sigma_x | H_1}, \frac{\sqrt{\xi}}{\sigma_x | H_1} \right) \]  

(14)

where \( Q_L(a,b) \) is the generalized Marcum-Q function of order \( L \) [9].

Under the \( H_0 \) hypothesis, we can rewrite equation (5) by separating the twofold nature of the received disturbance as

\[ x_i = \gamma_i + \eta_i \]  

(15)

\( \gamma_i \) in this case is the self-noise (SN) term, formally identical to (9), with

\[ \xi_m = \sqrt{E_s} e^{j(2\pi f [(m+\delta)T_s+\delta]+\phi)} \sum_{k=-\infty}^{\infty} a_k R_p(kT_s + \delta) \]  

(16)
and \( \eta_i \) again denotes the AWGN sample after coherent correlation. The complex AWGN samples, \( \eta_i \), are again i.i.d. Gaussian distributed complex variables with zero-mean and branch variance equal to \( \sigma_n^2 \). Applying the central limit theorem, the SN term can be modelled according to a Gaussian distribution, with zero mean and branch variance \( \sigma_{SN}^2 \) equal to

\[
\sigma_{SN}^2 = \frac{ME_s}{2} \sum_{k=-\infty}^{\infty} R_p^2 (kT_s + \delta)
\]  

(17)

Hence, the samples after coherent correlation \( x_i \) are Gaussian distributed with zero mean and branch variance \( \sigma_{x|H_0}^2 = \sigma_{SN}^2 + \sigma_n^2 \).

Because \( \Lambda \) is the sum of \( L \) squared modules of zero-mean complex Gaussian random variables, it is distributed as a central chi-square random variable with \( 2L \) degrees of freedom and variance of the composing real Gaussian variates equal to \( \sigma_x^2 \). Thus, the corresponding false alarm probability is

\[
P_{fa} = \int_{\Lambda} p_{A|H_0}(\Lambda)d\Lambda = \exp \left( -\frac{\Lambda}{2\sigma_{x|H_0}^2} \right) \sum_{k=0}^{L-1} \frac{\left( \frac{\Lambda}{2\sigma_{x|H_0}^2} \right)^k}{k!}
\]

(18)

A. Coherent Integration Length Dimensioning

Considering the signal-to-noise ratio under the \( H_1 \) hypothesis at the output of the coherent integration, \( \gamma_{CC} \), a novel theoretical optimization of coherent correlation length is here proposed. In fact, from (8), it can be noted

\[
\gamma_{CC} \propto ME_s \sin^2 (M\Delta fT_s)
\]

(19)

Thus, deriving with respect to the coherent correlation length, \( M \), the maximization of \( \gamma_{CC} \) is obtained by solving the equation

\[
\cot(\pi M \Delta fT_s) = \frac{1}{2\pi M \Delta fT_s}
\]

(20)

yielding

\[
M \simeq \frac{3}{8} \frac{1}{\Delta fT_s}
\]

(21)

We identify equation (21) as the CHILD rule (CoHerent Integration Length Dimensioning).

The important result lies in the fact that once the frequency error affecting the system is known, the coherent correlation length that maximizes the signal-to-noise ratio is simply determined. However, this maximization procedure does not take into account the PDI effect, and therefore it does not necessarily achieve the absolute performance optimum, but can be considered as a solid starting point for the optimization process. In support of this consideration, exhaustive numerical evaluation campaigns have shown that the CHILD rule value is typically a very tight upper bound for the optimum value of the overall performance. Furthermore, its independence from the particular adopted PDI scheme represents one of its strengths because it is general and applicable to all identifiable PDI methods.

Note that, the CHILD rule-based design implies to know the frequency error. In general, the frequency offset is variable in a range and the design is based on the worst case corresponding to the maximum frequency error, which is typically known for the specific application environment (i.e., oscillators types, transmitter/receiver motion). The worst case design is furthermore justified as the selection of a coherent correlation length larger than the value given by the CHILD rule leads to rapidly performance degrading.

IV. GENERALIZED TC PROCEDURE

The multi-dwell procedure can in general be modelled as a Markov chain which can be represented by the associated flow graph, where nodes represent the states of the chain, and branches are the transitions between the states [3]. Here, for generality, the multi-dwell TC procedure characterization is developed for a generic number of dwells, \( N \). Oversampling has to be introduced if symbol timing synchronization is not available, to prevent the receiver from operating in unfavorable cases, e.g. \( \delta = T_s/2 \). We denote with \( h \) the number of samples in a symbol, equivalently addressed as hypotheses or cells in the following. If the detection variable in verification is below threshold, the procedure is restarted from the search mode (immediate rejection). In fig.2, the generalized flow-graph is depicted for \( N = 3 \) and \( h = 2 \). \( G^{(j)}_{CR,i}(z) \) and \( G^{(j)}_{FA,i}(z) \) indicate the correct rejection and false alarm branch gains in the \( i \)th dwell, testing the \( j \)th hypothesis

\[
G^{(j)}_{CR,i}(z) = (1 - P_{fa,i}^{(j)} z^i) \quad G^{(j)}_{FA,i}(z) = P_{fa,i}^{(j)} z^i
\]

(22)
where $t_1 = T_d$, being $T_d$ the dwell time, equal to $T_s/h$ (assuming passive correlation with matched filtering), with $T_s$ equal to the symbol period, and $T_f = T_p$ for $1 < i < N$ being $T_p$ the frame duration; similarly, $G_{D,i}^{(j)}(z)$ and $G_{M,D,i}^{(j)}(z)$ indicate the correct detection and missed detection gains in the $j$th dwell, testing the $j$th hypothesis

$$G_{D,i}^{(j)}(z) = P_{d,i}^{(j)}z^j, \quad G_{M,D,i}^{(j)}(z) = (1 - P_{d,i}^{(j)})z^j. \quad (23)$$

False alarm is assumed to be a non-absorbing state with associated penalty time $T_p$, modelled here as a fixed value. $G_P(z)$ indicates the branch transfer function for recovering from false alarm

$$G_{P,i}^{(j)}(z) = G_{P,i}^{(j)}(z) = z^{N_i+1} \quad (24)$$

where $T_{i+1} = T_p$, and the index $N + 1$ denotes the tracking phase with $P_{fa,N+1} = 0$. We assume that after the penalty time the search phase restarts from the subsequent cell.

Uniform a priori probability distribution is assumed, i.e. $\pi_{n} = 1/N_c$, where $N_c$ is the total number of hypotheses constituting the uncertainty region. The equivalent flow-graph reported in fig.3 can be simply derived, grouping the synchronous cells into a single collective state $S$ with a priori probability equal to $h\pi_n$. Following a procedure similar to that described in [10], the associated transfer function results to be

$$p(z) = \frac{G_D(z) \sum_{n=0}^{N} \prod_{j=1}^{N} G_{o,j}^{(j)}(z)}{1 - G_M(z) \prod_{j=1}^{N} G_{o,j}^{(j)}(z)} \quad (25)$$

where

$$G_{o,j}^{(j)}(z) = \sum_{i=1}^{N} G_{o,j}^{(j)}(z) \prod_{k=1}^{i-1} G_{F,A,k}^{(j)}(z) + G_P(z) \prod_{i=1}^{N} G_{F,A,i}^{(j)}(z) \quad (26)$$

$$G_M(z) = \prod_{j=1}^{h} \sum_{j=1}^{N} G_{M,D,k}^{(j)}(z) \prod_{i=1}^{k-1} G_{D,i}^{(j)}(z) \quad (27)$$

$$G_D(z) = \sum_{j=1}^{h} \prod_{i=1}^{N} G_{D,i}^{(j)}(z) \prod_{i=1}^{N} G_{M,D,k}^{(j)}(z) \prod_{i=1}^{k-1} G_{D,i}^{(j)}(z) \quad (28)$$

and $Q = N_c - h$ is the number of non-synchronous cells in the uncertainty region. Notice that it is here intended that $\prod_{x=1}^{N} x = 1, \forall x$.

As known, the mean acquisition time can be obtained evaluating the derivative of (25) for $z=1$. Notice that the ACM 1TC and 3TC can be derived from the general characterization by taking $(N,h)$ as $(1,2)$ and $(3,1)$, respectively.

Here, we specialize the general multi-dwell procedure, assuming $h = 1$, which is the case of 3TC. In fact, assuming that timing has already been recovered by a blind timing acquisition circuit, frame acquisition can be performed assuming one hypothesis per symbol, and timing displacement $\delta = 0$. A residual limited timing error would not affect the validity of the following model, but only introduce a performance degradation. Therefore $N_c = L_F$.

Let the overall probabilities of correct detection, missed detection at $i$th dwell, and correct rejection at $i$th dwell be defined as $P_D = P_{d,1}P_{d,2}...P_{d,N}$, $P_{M,i} = (1 - P_{d,i})\prod_{j=1}^{i-1} P_{d,j}$ for $1 \leq i \leq N$, and $P_{R,i} = (1 - P_{fa,i})\prod_{j=1}^{i-1} P_{fa,j}$ for $1 \leq i \leq N + 1$. The test duration until the $i$th dwell can be computed as

$$T_i = \sum_{j=1}^{i} T_j \quad i = 1, 2, \ldots, N + 1 \quad (29)$$

Thus, the average time periods to overall missed detection, correct rejection, and correct detection are

$$\bar{T}_M = \sum_{i=1}^{N} T_i P_{M,i}, \quad \bar{T}_R = \sum_{i=1}^{N+1} T_i P_{R,i}, \quad \bar{T}_D = T_N P_D \quad (30)$$

Finally, the mean acquisition time, $\bar{T}_A$, is [11]

$$\bar{T}_A = \frac{\bar{T}_D}{P_D} + \frac{\bar{T}_M}{P_D} + \frac{2 - \bar{T}_D}{2P_D}(N_c - 1)\bar{T}_R \quad (31)$$
The single-dwell TC (1TC) procedure is used for beam hopping, where symbol timing has not been recovered up-front, thus at least two hypotheses per symbol must be considered \((h=2)\) with in general fractional timing displacement \(\delta\). The uncertainty region spans over the entire hop period. In this case, the number of cell in the uncertainty region, \(N_c\), is equal to \(2L\_Hop\). The hop period is assumed equal to the frame duration of continuous operation, i.e. \(L\_Hop = L_F\). The corresponding flow-graph is obtained by grouping the 2 synchronous cells into the collective state \(S\). The number of \(H_0\) cells is therefore \(Q = N_c - 2\).

In this case, the mean acquisition time results to be

\[
T_A = \frac{1}{P_D} \left\{ T_d \left[ 1 + \frac{Q}{2} (2 - P_D) \right] + T_p \frac{Q}{2} P_{fa} (2 - P_D) \right\}
\]

where \(P_D = P_d (2 - P_d)\) is the overall correct detection probability associated to the collective synchronous state, being \(P_d\) the single cell correct detection probability (we omit the index \(i\) for convenience).

\section{VII. DPDI PERFORMANCE ANALYSIS}

Because DPDI performance characterization is not trivial, the evaluation of the overall performance improvement achievable with respect to NCPDI has been based on a semi-analytical approach, that have been validated with simulations. This semi-analytical method employs the simulated single cell probabilities (i.e. missed detection and false alarm probabilities) for the computation of the analytical mean acquisition time obtained in sec.IV. Notice that the same technique is applicable to NCPDI, too.

In order to appropriately select the dwell threshold, minimizing the mean acquisition time, the fitting procedure proposed in [12] has been adopted to characterize the behavior of missed detection and false alarm probability of DPDI varying the decision threshold. The minimum mean square error optimization has provided very tight match between fitting and simulated curves, so allowing accurate performance characterization.

\section{VI. NUMERICAL RESULTS}

Performance evaluation of the proposed procedures follows a threefold approach: the fully analytical characterization, a semi-analytical approach which uses the simulated single cell performance and the analytical mean acquisition time, and the simulated approach, where also the multi-dwell procedure is simulated. The first approach is applied only to NCPDI for which analytical characterization of correct detection and missed detection probabilities have been detailed above. The penalty time is fixed equal to twice the frame duration, and the normalized frequency error is set to 0.056 (i.e. 5MHz at 90 MBaud) [1]. At this frequency offset value, the CHILD rule indicates to set \(M\) equal to 6. \(\delta\) is set to zero in 3TC, and is assumed equal to \(T_s/4\) in 1TC, corresponding to the worst case. The roll-off factor is assumed equal to 0.3.

First of all, the single cell Receiver Operating Characteristic (ROC) performance, i.e., \(P_{md}\) vs. \(P_{fa}\), is presented in fig.4 for NCPDI and DPDI. The UW length is equal to 96 symbols, with \(M=6\), and \(L=16\). For NCPDI, the solid lines identify the analytical performance, while markers refer to simulation results. For all considered \(\delta\) values \((0, T_s/4, T_s/2)\), a good match can be observed. DPDI simulation results are also reported. A considerable improvement with respect to NCPDI can be noted.

Cross-validation of the three different methodologies is presented in fig. 5 and 6 for 3TC and 1TC respectively, with a frame length of 1000 symbols in a \(E_s/N_0\) range of \([-6, -3]\)dB. The evident result is the good match among the three approaches for NCPDI and between simulations and the semi-analytical approach for DPDI. Note that the importance of this agreement lies in the fact that numerical simulation results cannot practically be achieved because of the long required simulation time for large frames, as that of the ACM project. The semi-analytical approach allows to accurately evaluate practical cases considering all impairments that can affect single cell performance. Therefore, we consider only analytical and semi-analytical performance evaluation in the actual ACM case with \(L_F = 2880000\) symbols (corresponding to a 32 ms frame duration at 90 MBaud), and with \(E_s/N_0 \in [-2.3, 0.7]\)dB.

The thresholds employed in each dwell are optimized for the worst case, i.e. \(E_s/N_0=-2.3\)dB, in order to minimize the associated mean time to acquisition both for NCPDI and DPDI. The CFAR (constant false alarm rate) normalization ensures that this choice is close to the optimum also for larger \(E_s/N_0\). Fig.7 reports the minimum mean acquisition time vs. \(E_s/N_0\) for 3TC and 1TC. The relevant result is that with 3TC outperforms 1TC when NCPDI is employed, while the inverse behavior characterizes DPDI. Thus, in the specified study case, the employment of DPDI results in a complexity reduction (1TC instead of 3TC) and performance improvement at the same time.

Fig. 8 presents the conservative \(L\_UW\) design based on NCPDI at -2.3dB. The mean acquisition time specification is 2 s. The result is that with 3TC a minimum length of 84 symbols is enough to ensure the desired performance, while with 1TC an extended length of 108 is needed.
A novel analytical characterization of frame synchronization has been provided for continuous and discontinuous transmission, based on the threshold crossing criterion. In particular, three-dwell and single-dwell procedures have been presented. Non coherent and differential post detection integration schemes have been considered to cope with frequency error, and a novel and powerful analytical rule for optimal coherent correlation length selection has been proposed. The good agreement between analytical and simulated results allows to obtain accurate parameter optimization and fast performance evaluation. The problem of frame synchronization has been addressed through a general and complete methodology, that can embrace all actual scenarios in the DVB-S2 context and beyond. Numerical results show that DPDI can significantly reduce the mean acquisition time, together with the acquisition subsystem complexity, since it allows the adoption of the single dwell procedure instead of the three-dwell needed with NCPDI. Open interesting points are the detailed system optimization employing DPDI, and possible alternative UW designs.

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REFERENCES


Fig. 1. Non coherent and differential post detection integration block diagrams
Fig. 2. Generalized TC flow-graph, $N=3$, $h=2$

Fig. 3. Equivalent TC flow-graph, $N=3$, $h=2$

Fig. 4. NCPDI and DPDI receiver operating characteristic, $\Delta f=5$ MHz, 90 MBaud, $E_s/N_0=-2.3$dB

Fig. 5. 3TC procedure; analytical, semi-analytical and simulated performance. $\Delta f=5$ MHz, 90 MBaud, $L_{\text{Frame}}=1000$, $\delta=0$

Fig. 6. ITC procedure; analytical, semi-analytical and simulated performance. $\Delta f=5$ MHz, 90 MBaud, $L_{\text{Frame}}=1000$, $\delta=0.25$

Fig. 7. Mean acquisition time vs. $E_s/N_0$, $\Delta f=5$ MHz, 90 MBaud, $L_{\text{Frame}}=2880000$

Fig. 8. UW length design based on NCPDI. $\Delta f=5$ MHz, 90 MBaud, $L_{\text{Frame}}=2880000$, $E_s/N_0=-2.3$dB