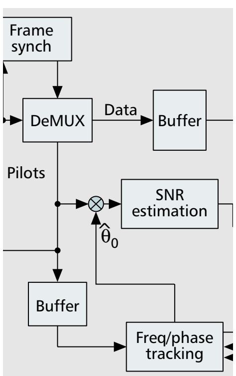
ON THE ADAPTIVE DVB-S2 PHYSICAL LAYER: DESIGN AND PERFORMANCE

GIANNI ALBERTAZZI, STEFANO CIONI, GIOVANNI E. CORAZZA, MASSIMO NERI, RAFFAELLA PEDONE, PAOLA SALMI, ALESSANDRO VANELLI-CORALLI, AND MARCO VILLANTI, ARCES, UNIVERSITY OF BOLOGNA



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ABSTRACT

The successful DVB standard has now evolved into the DVB-S2 standard, which promises to bring very significant capacity gains. The main DVB-S2 feature is its adaptive air interface, where coding and modulation techniques are varied flexibly to maximize performance and coverage. This article addresses the design of the entire DVB-S2 communication chain, considering practical algorithms for coding, modulation, predistortion, carrier and SNR estimation, frame synchronization, and data recovery. The design complexity is exacerbated by the fact that DVB-S2 foresees 28 different coding/modulation pairs, demanding specific optimization and variable frame length. The performance achieved considering all possible impairments is compared to the ideal performance achievable in the Gaussian channel in terms of integral degradation, which ranges from 0.4 to 2.5 dB in going from QPSK to 32-APSK.

INTRODUCTION AND MOTIVATION

The digital video broadcasting-satellite (DVB-S) standard is now the most widely adopted transmission protocol for broadcasting by satellite. This fact has prompted the development of the evolved standard identified as DVB-S2 [1], which promises to yield very significant system capacity gains, as well as the extension of the set of services and applications that can be delivered to an enlarged mass market. In particular, enhanced broadcasting will be paired to interactive communications for broadband access to the Internet and multimedia content distribution. In essence, DVB-S2 can be considered a valid alternative for solving the last mile dilemma and bridging the digital divide.

The necessary bandwidth resources for large throughputs can only be found at extremely high frequencies, such as Ka-band and above. Here, atmospheric effects on system capacity can become truly dramatic, and can only be maximized by abandoning the conservative worst case design approach in favor of flexible and adaptive transmission mode selection to efficiently exploit the available power and spectrum resources. This is the underlying rationale of the adaptive coding and modulation (ACM) air interface adopted within the DVB-S2 standard. The cost of adaptivity is the increased complexity represented by the large number of modulation/coding sets and the ever more critical parameter estimation and synchronization procedures.

The design of DVB-S2-compatible data recovery and synchronization procedures is an interesting open problem, calling for smart and effective solutions accounting for actual impairments, such as phase noise and nonlinearities, and operating at extremely low signal-to-noise ratios (SNRs), close to the Shannon limit. The design complexity is exacerbated by the fact that DVB-S2 foresees 28 different coding/modulation pairs, demanding specific optimization and variable frame length, which calls for optimized design of the frame synchronization subsystem. The latter can benefit from the use of advanced post-detection integration techniques described in the following. This work aims at proposing algorithms that embrace the overall degrees of freedom provided by adaptivity, while keeping complexity at a minimum.

These issues have been separately addressed for data recovery in [2] and frame synchronization in [3]. Differently, this article considers the entire transmission chain to assess overall performance, addressing at the same time the data decoding and synchronization issues. Performance is thus comprehensive of all nonidealities affecting the actual system, contrasted with the ideal additive white Gaussian noise (AWGN) case, to evaluate the so-called *integral degradation*.

This article introduces the reader to the essential elements of the DVB-S2 adaptive physical layer and guides him/her through a description of a possible algorithmic implementation. More in detail, the transmit-receive chain comprises high order modulation schemes, resilient to nonlinear effects; low density parity check (LDPC) codes; fractional predistortion techniques, operating after the shaping filter to

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counteract nonlinear effects and mitigate the intersymbol interference (ISI); nonlinear and linear distortions, introduced by the onboard high power amplifier (HPA) and input/output multiplexing filters (I/O muxes); frame synchronization and symbol timing estimation; frequency offset and phase recovery followed by data detection. Interestingly, the actual integral degradation of this transmission chain affected by all possible impairments can be limited by judicious design to rather small values, ranging from 0.4 to 2.5 dB going from quadrature phase shift keying (QPSK) to 32-amplitude and PSK (APSK), as shown in this article. To achieve these satisfactory results, predistortion techniques are essential for 16- and 32-APSK modulations. In the following, channel coding and modulation schemes as well as predistortion techniques are described. We illustrate the proposed DVB-S2 digital receiver, the performance of which is also discussed. Finally, overall conclusions are drawn.

Channel Coding and Modulation Schemes

The most significant feature introduced in the DVB-S2 standard [1] is the capability to select the coding-modulation pair to optimize the system capacity in terms of achievable system throughput, under the hypothesis of a finite user population [4]. The coding and modulation schemes selected for DVB-S2 leverage the most recent and powerful advancements in the field of communication theory. In particular, powerful coding schemes that essentially achieve channel capacity are now available with simple implementation and affordable complexity. Specifically, the DVB-S2 forward error correction (FEC) encoding encompasses three components: an outer systematic Bose-Chaudhuri-Hocquenghem (BCH) code, an inner LDPC code, and a block bit interleaver.

The BCH [5] block codes are introduced to remove the possible error floor produced by LDPC undetected errors. However, the major error correcting capability definitely comes from the LPDC element. The LDPC code family was first proposed by Gallager in 1962 [6], and then rediscovered during the 1990s (e.g., [7]). LDPC codes are (k, n) linear block codes with a sparse parity check matrix and very large n (e.g., n in the thousands). LDPC codes can be represented by a bipartite Tanner graph [7], in which the ith check node is connected to the jth bit node if and only if the matrix element at position (i, j) is non-null. The sparsity of the graph is key to efficient algorithmic encoding and iterative decoding of these long codes. LDPC codes are defined as *regular* if the number of ones in every row, the so-called *check node degree*, and the number of ones in every column, identified as the *bit node degree*, are constant; they are identified as *irregular* otherwise. The DVB-S2 standard codes are irregular LDPC codes with two block length options, $n = 64{,}800$ and $n = 16{,}200$. The specified parity check matrices have a well defined structure in order to enable low-complexity systematic encoding and parallel decoding.

In order to cover an exhaustive range of spectral efficiencies, 11 different code rates (1/4, 1/3, 2/5, 1/2, 3/5, 2/3, 3/4, 4/5, 5/6, 8/9, 9/10) are specified in conjunction with four different modulation schemes for a total of 28 allowed combinations, as reported in [1, Table 12]. Besides classic QPSK and 8-PSK modulation schemes, a hybrid APSK modulation format is considered in order to have high bandwidth efficiency and cope with nonlinear satellite transponders. In essence, APSK is formed by concentric circles of constellation points. In particular, 16- and 32-APSK modulations are specified with two and three rings of points, respectively. The characteristic parameters of APSK are optimized as a function of the LDPC code rate [1].

Finally, the bit interleaving for 8-PSK, 16-APSK, and 32-APSK is performed through a block interleaver that acts after the LDPC encoder in order to increase code diversity.

Nonlinear Satellite Channel and Predistortion Techniques

A main source of degradation is given by the presence of saturation-driven HPAs, which are normally employed for satellite links to maximize output power and DC/RF conversion efficiency. The HPA is typically described by nonlinear AM/AM and AM/PM characteristics, and increases adjacent channel interference (ACI) due to spectrum sidelobes regrowth, as well as warping and clustering of the signal constellation. The complexity is enhanced by the adoption of variable high order modulations, while frame-by-frame adaptivity results in nondeterministic waveforms, which increase the system distortion sensitivity.

Several techniques have been proposed in the literature in order to mitigate the impact of saturation-driven HPAs. A straightforward solution is to introduce input backoff (IBO) to step back from saturation, with a consequent output backoff (OBO). For ground terminal amplifiers this may be acceptable (up to a reasonable level) because DC/RF conversion efficiency is not at a prime. For onboard amplifiers, the larger the OBO, the smaller the coverage and conversion efficiency; therefore, this approach must be used very sparingly, and other techniques are necessary.

Compensation techniques can be applied at both transmitter and receiver. Typical receiver techniques involve the use of equalizers, with the related drawbacks of increased receiver complexity and impossibility of eliminating the undesirable spectrum spillover. It is therefore preferable to use transmitter countermeasures to precompensate for the HPA input and achieve quasi-linear amplification of the signal. This approach is commonly referred to as *predistortion* [8], and can be classified into two main classes, analog and digital. Analog predistortion is performed at RF (or possibly at intermediate frequency [IF]), while digital predistortion is carried out at baseband, either before or after the pulse shaping filter, identified as *data* and *fractional* approaches, respectively. The main advantages of digital predistortion are flexibility, adaptivity, and ease of implementation.

The proposed predistortion method for DVB-S2 involves the use of a digital predistorter located after the pulse shaping filter, operating on an oversampled signal. The signal is processed by

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Following a sequential functional order from startup, the first operation after matched filtering is clock recovery, followed by frame synchronization, which exploits the physical layer frame header. Both operations must be performed in the presence of large carrier frequency errors.

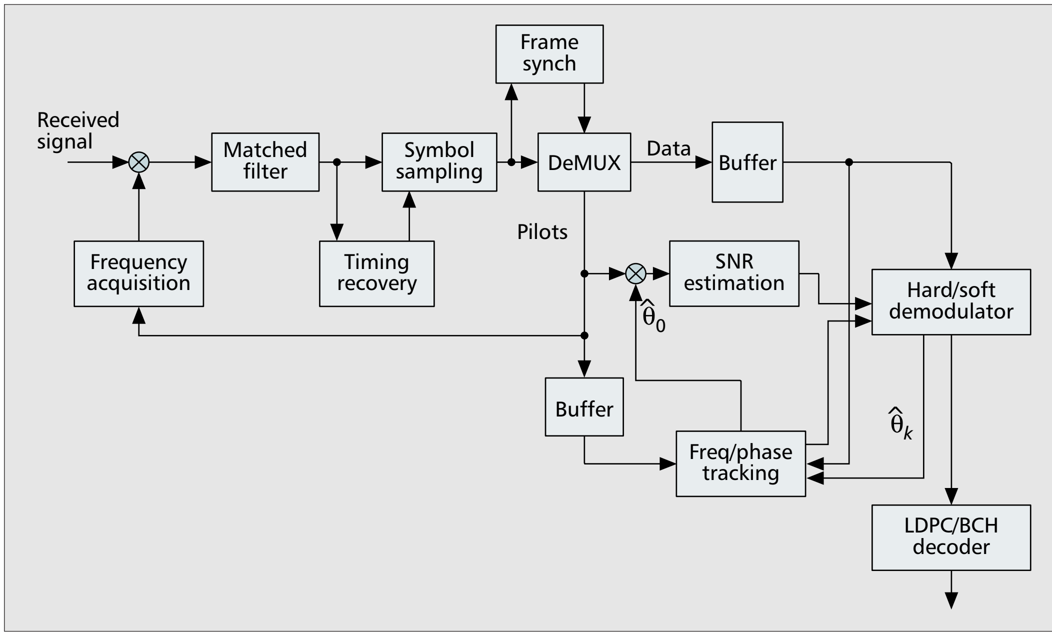

Figure 1. *Digital receiver block diagram.*

means of a lookup table (LUT), which stores the inverted HPA coefficients computed offline through analytic inversion of a proper HPA model. The steps needed to obtain LUT coefficients are:

- HPA model selection
- Parameter extrapolation
- Analytic model inversion
- LUT definition and construction

Regarding the HPA model, a simple yet robust empirical model is the classic Saleh model [9]. Here, a more sophisticated and accurate model is adopted, which involves the use of polynomial ratios with 10 parameters (five for the amplitude and five for the phase). Given the measured HPA characteristics, the model parameters can be extracted by minimizing the energy of the difference between the modeled and experimental HPA curves (minimum mean square error [MMSE] criterion). These parameters are then applied to the analytically inverted characteristics to obtain the analytical predistortion transfer function. The last step is the discretization of the analytical curve in order to store it into the LUT. The adopted strategy is linear in power indexing (i.e., table entries are uniformly spaced along the input signal power range), yielding denser table entries for larger amplitudes where nonlinear effects reside.

Digital Receiver Implementation

Having described the main baseband transmitter blocks, attention is now focused on the receiver. In Fig. 1 a possible DVB-S2 digital receiver architecture is depicted. Several subsystems are involved in the recovery of information. Following a sequential functional order from startup, the first operation after matched filtering is clock recovery, followed by frame synchronization, which exploits the physical layer frame (PLFRAME) header. Both operations must be performed in the presence of large carrier frequency errors. After this initial synchronization procedure, a demultiplexer is used to separate pilots from data symbols in a PLFRAME. The pilot symbol stream is used by the following three subsystems: the coarse frequency acquisition loop, the signal-to-noise level estimator, and finally a phase-locked loop (PLL) for tracking the residual frequency offset and carrier phase. Once all auxiliary parameters are recovered, the data symbols can be detected by the hard/soft demodulator. The hard decisions are fed back to the PLL, while the soft initial *a posteriori* probabilities (APPs) on the received information bits are passed on to the LDPC-BCH decoder. In the following sections each subsystem is described in more detail.

Symbol Timing Recovery

A suitable algorithm for timing recovery in DVB-S2 is the Gardner estimator [10]. In fact, this is a non-data-aided (NDA) circuit that is virtually insensitive to modulation format (from QPSK to 32-APSK) and performs efficiently even in the presence of quite large carrier frequency errors over the range of E_s/N_0 of interest. After the Gardner timing adjustment, the sampled symbol is indicated as

$$r_k = d_k e^{j\phi_k} + n_k, \tag{1}$$

where d_k is the transmitted symbol, ϕ_k contains the carrier frequency and phase offsets, and n_k represents the AWGN sample. In terms of achievable performance, with an oversample factor equal to 4, the overall acquisition transient can be completed in around 10^5 symbols (2–3 PLFRAMEs) independent from the modulation format, while at steady state the normalized residual timing jitter is less than 10^{-2}, which is more than satisfactory.

Frame Synchronization

Frame synchronization for DVB-S2 can be divided into acquisition and tracking, as in best practice. Frame acquisition (i.e., the initial frame epoch detection) is the most critical phase and thus is investigated hereafter. DVB-S2 frame acquisition is performed by exploiting the autocorrelation properties of the PLFRAME header, which is a sequence of 26 $\pi/2$-binary PSK (BPSK) symbols, identified as start of frame (SOF). The received signal is correlated with locally generated SOF replicas shifted by discrete offsets. Therefore, the frame epoch estimation problem is translated into a detection problem that has to discriminate between *hypotheses* or *cells* in a discretized *uncertainty region*. In particular, we indicate as H_1 the hypothesis corresponding to the achieved synchronism, while H_0 marks all misaligned cells. A possible detection approach is the threshold crossing (TC) criterion [11], which amounts to declaring H_1 (H_0) if the correlation output does (does not) cross a properly designed threshold.

Frame acquisition in DVB-S2 suffers from two major impairments: the extremely low SNR, which can indeed assume negative values in dB, and the unknown carrier frequency offset and phase. In addition, at terminal startup the uncertainty region equals the entire frame length, T_F, which in the worst case is as large as 33,282 QPSK symbols. At first glance, one could hope to improve performance by exploiting multidwell procedures (i.e., collecting information from multiple frames before making a final decision). Unfortunately, this approach cannot be used in DVB-S2, where the frame length depends on the selected coding/modulation pair. In particular, the frame format is signaled to the receiver by a 64-symbol physical layer signaling (PLS) field. Since the PLS field cannot be decoded accurately prior to frame synchronization, the latter must be performed in single-dwell fashion. In any case, a genie-aided multidwell case can be considered as a performance benchmark. The single-dwell TC procedure is here identified as 1TC, while a benchmark three-dwell strategy is indicated as 3TC.

The detection of the SOF auto-correlation peak is heavily affected by the presence of the frequency offset, Δf, which bans the possibility of correlating coherently over the entire SOF. To cope with this problem, the classic noncoherent post-detection integration (PDI) approach could be adopted, but this is outperformed by the differential PDI (DPDI) scheme [12], sketched in Fig. 2, which performs coherent correlation over L SOF-segments of length M, followed by differential detection and combining. M and L should be optimized considering worst case Δf. A quick set for M is given by the coherent integration length dimensioning (CHILD) rule [13], according to which the value of M should approach

$$M \approx \frac{3}{8\Delta f T_s}, \tag{2}$$

where T_s is the symbol period, to maximize the SNR at the output of coherent correlation.

The frame acquisition performance is measured in terms of mean acquisition time, for which a practical performance specification of 2 s can be assumed at terminal startup. In the proposed design, a serial search procedure is considered where possible false alarms are recovered by a tracking circuit that restarts the procedure after a penalty time of T_p s. Passive correlation (SOF matched filtering) is used for coherent integration to reduce delay at the price of increased complexity. Assume that E_s/N_0 is in the typical DVB-S2 range [–2.3, 0.7] dB, the frequency offset is Δf = 5 MHz, and the baud rate is 27.5 Mbaud, and let the SOF length of 26 symbols be subdivided in M = 2 and L = 13. Thresholds are optimized in order to minimize the mean acquisition time at the worst SNR, –2.3 dB. Observing the performance results summarized in Table 1 for T_p in the range [0–2T_F], it can be noted that 1TC is indeed the optimal choice and is largely in spec. This is due to the fact that DPDI yields low false alarm and missed detection probabilities, avoiding the need for multiple dwells. As a further advantage, 1TC entails at the same time the smallest complexity, and does not require PLS decoding. Hence, DPDI is perfectly suited to frame synchronization for the DVB-S2 standard.

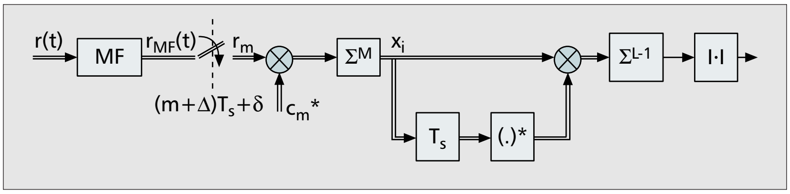

Figure 2. *Differential post detection integration block diagram.*

Ndwells	E_s/N_0 (dB)	Penalty time	Normalized threshold	Mean acquisition time (ms)
3	–2.3	2 TF	8, 1, 1	578
1	–2.3	2 TF	8	244
1	0.7	2 TF	8	42
1	–2.3	1 TF	8	162
1	0.7	1 TF	8	32
1	–2.3	0	4	3.2
1	0.7	0	4	1.3

Table 1. *Frame synchronization performance results.*

An alternative frame acquisition design approach is reported in [3]. There, SOF differential combining is performed without coherent accumulation, which can lead to inferior performance, especially for smaller Δf. This requires the exploitation of the Reed-Muller encoded PLS field to aid frame acquisition, introducing considerably increased complexity in the detector. In summary, the adoption of the DPDI scheme with optimized coherent correlation length allows the requirements to be satisfied with reduced complexity.

Coarse Carrier Frequency Recovery

Coarse frequency recovery is also a critical step due to the fact that DVB-S2 mass market terminals will typically incorporate low-cost oscillators, which introduce large initial frequency offsets (e.g., 5 MHz at 27.5 Mbaud). Fortu-

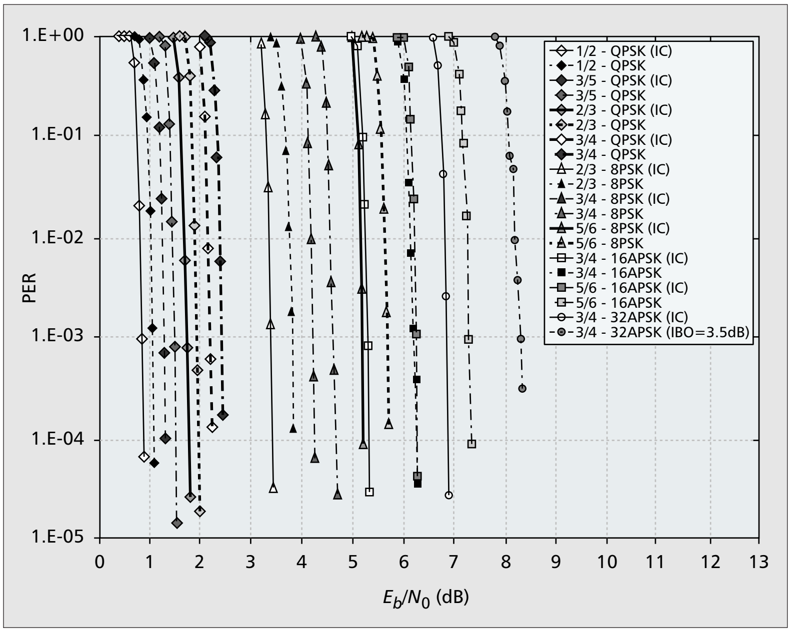

Figure 3. *PER comparison between AWGN channel (IC) and non linear distorted channel, IBO = 2 dB, with channel parameter estimation.*

nately, after frame synchronization is achieved it is possible to exploit the pilot fields introduced in the DVB-S2 PLFRAME. In particular, the specified length for the pilot field is $N_p = 36$. In these conditions, a pilot-aided Mengali and Morelli (M&M) algorithm [14] appears to be a valid candidate, because it allows to have low estimation error variance (approaching the Cramer-Rao bound) and a sufficiently large pull-in range, according to the requirements. The M&M frequency estimate is computed as

$$\hat{f}_0 = \frac{1}{2\pi T} \sum_{m=1}^{N} w_m \arg\{R(m)R^*(m-1)\}, \qquad (3)$$

where N is a design parameter, and

$$w_m = 3\frac{(N_p - m)(N_p - m + 1) - N(N_p - N}{N(4N^2 - 6N\,N_p + 3N_p^2 - 1)}, \qquad (4)$$

N_p being the observation length in symbols. The optimum choice for N is $N_p/2$. The auto-correlation function, $R(m)$, is defined as

$$R(m) \equiv \frac{1}{N_p - m} \sum_{k=m}^{N_p - 1} (r_k d_k)(r_{k-m}^* d_{k-m}^*). \qquad (5)$$

The frequency estimate can be further improved by averaging the auto-correlation function over M_P consecutive pilot fields over several PLFRAMEs. In this case the auto-correlation function is replaced by its average

$$R(m) = \sum_{i=1}^{M_P} R_i(m), \qquad (6)$$

$R_i(m)$ being the auto-correlation contribution of the ith pilot field defined in Eq. 5. As a practical example, consider $M_P = 660$, which corresponds to 30 QPSK PLFRAMEs, or equivalently to an estimation window of 36 ms. The normalized residual frequency root mean square (RMS) error for the averaged pilot-aided M&M estimator is less than 10^{-4} for E_s/N_0 larger than 1 dB. Interestingly, this result shows performance improvements over the scheme proposed in [2].

SNR Estimation

Adaptive physical layer receivers require accurate estimation of the received SNR, for two basic reasons: first, this estimate can be fed back to the network to determine the most suitable coding/modulation pair according to the channel conditions experienced by the specific user; second, this information is needed by the soft demodulator to compute the APPs on the received symbols. A suitable algorithm to this purpose is the SNR Estimator (SNORE) [15], which is here performed in a data-aided fashion over the pilot field after frequency and timing corrections have taken place. Considering a possible DVB-S2 system at Ka-band, it can be assumed that channel propagation impairment (atmospheric attenuation) is a very slowly varying process compared to SNR estimation time; thus, quasi-stationary channel conditions can be considered. Briefly, the useful power, P_S, can be estimated as follows:

$$\hat{P}_S = \left[\frac{1}{N_P} \sum_{k=1}^{N_P} \mathrm{Re}\left\{r_k d_k^*\right\}\right]^2, \qquad (7)$$

where N_p is the number of known symbols for coherent accumulation, and d_k is a QPSK pilot symbol. On the other hand, the estimate of the total received power, P_R, is given by

$$\hat{P}_R = \frac{1}{N_P} \sum_{k=1}^{N_P} |r_k|^2 \qquad (8)$$

so that the noise (plus possible interference) power estimate is $\hat{P}_N = \hat{P}_R - \hat{P}_S$. Finally, the SNR estimate is computed as the ratio $\hat{P}_S$ over $\hat{P}_N$. Exploiting the slowly varying channel conditions, the SNR estimate can further be averaged over the last W values to reduce the estimation error standard deviation. Assuming $N_p = 36$ and $W = 50$, the estimated SNR is unbiased and the RMS error is on the order of 3×10^{-2}, which is perfectly adequate for both uses outlined above.

Fine Carrier Frequency and Phase Tracking

The residual frequency offset after coarse acquisition and phase estimation (penalized by strong phase noise) is performed through a second-order PLL that exploits all possible aiding mechanisms, using a hybrid data-aided and decision-directed approach. The initial phase estimate, $\hat{\theta}_0$, is obtained by a maximum likelihood (ML) feedforward estimator over a pilot sequence, and is employed by a decision-directed second-order PLL.

Within a PLFRAME, whenever a new pilot field occurs, the PLL operates according to a data-aided approach exploiting the known pilot symbols (Fig. 1).

LDPC Decoding

LDPC decoding is performed through the iterative sum-product algorithm, which updates the APP values after each iteration, exploiting the Tanner graph. This is a message passing algorithm, where-by at the jth iteration the message sent from a bit node v to a check node c is denoted $\Lambda^{j,v\to c}$, and the message in the opposite direction $\Lambda^{j,c\to v}$. As shown in [7], these messages are computed as

$$\Lambda_k^{j,c\to v} = 2\cdot\tanh^{-1}\left\{\prod_{\substack{i=1\\ i\neq k}}^{d_c}\tanh\left\{\frac{1}{2}\Lambda_i^{j,c\to v}\right\}\right\} \quad (9)$$

$$\Lambda_k^{j,v\to c} = \Lambda_k^0 + \sum_{\substack{i=1\\ i\neq k}}^{d_v}\Lambda_i^{j-1,c\to v}, \quad (10)$$

where Λ_k^0 is the initial APP value given by the soft demodulator related to r_k. This simple decoding algorithm proceeds iteratively until the code parity check constraints are all verified, or a maximum number of iterations is reached. As shown in Eq. 9, the check node updating rule takes into account the tanh(·) function, which introduces significant computation complexity. For this reason, proper approximations of this function are recommended for use in actual hardware implementations.

Performance Results

The entire DVB-S2 transmission chain has been simulated using a purposefully developed C++ software tool. The final results are reported in terms of packet error rate (PER) in steady state conditions, after successful frame synchronization and coarse carrier frequency recovery, where the residual offsets and uncertainties are continuously tracked by the receiver. The user packet length is assumed to be equal to 1504 bits (MPEG format), while the long LDPC coded block (64,800 symbols) is selected. A reference PER target of 10^{-4} is specified here, which serves as a benchmark for all E_b/N_0 performance losses.

The performance results reported in Fig. 3 show the comparison between DVB-S2 FEC code performance in an AWGN channel (labeled ideal channel, IC), and a channel with nonlinear distortion and parameter estimation inaccuracies. In the latter case, the onboard HPA works with an IBO = 2 dB that corresponds to OBO = 0.43 dB for both QPSK and 8-PSK modulations, and 0.63 dB for the 16-APSK case. Furthermore, a carrier phase noise mask compliant with the DVB-S2 standard [1] has been adopted. With the adoption of the estimation blocks described above, actual values for the estimation inaccuracies have been considered. In particular, the normalized symbol timing offset has a standard deviation of 10^{-2}, and the residual normalized frequency error has a standard deviation of 10^{-4}, 5×10^{-5}, and 3×10^{-5} for QPSK, 8-PSK, and 16-APSK, respectively. The degradation with respect to the IC case ends up being on the order of 0.25 dB for QPSK, 0.4 dB for 8-PSK, 1.0 dB for 16-APSK, and finally 1.45 dB for 32-APSK. These very small losses with respect to IC are achievable thanks to the efficient predistortion technique that has been purposefully designed. This fact is confirmed by the results of Fig. 4, where the performance with and without predistortion is compared for IBO = 2 dB. Apart from QPSK and 8-PSK schemes, in which the degradation is limited, predistortion introduces a gain of about 2 dB for 16-APSK, while 32-APSK is not reported as it is completely unreliable without compensation techniques for IBO values lower than 4 dB.

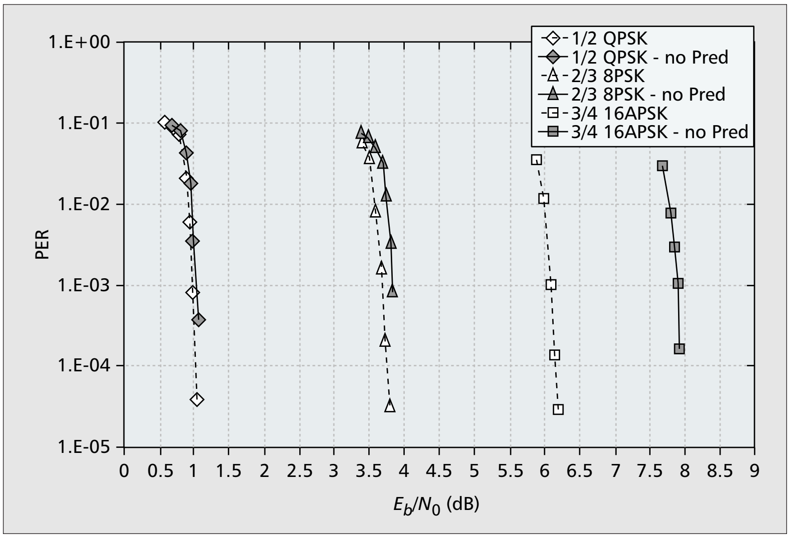

Figure 4. *PER comparison in presence of nonlinear distorted channel, IBO = 2 dB, with and without predistortion technique.*

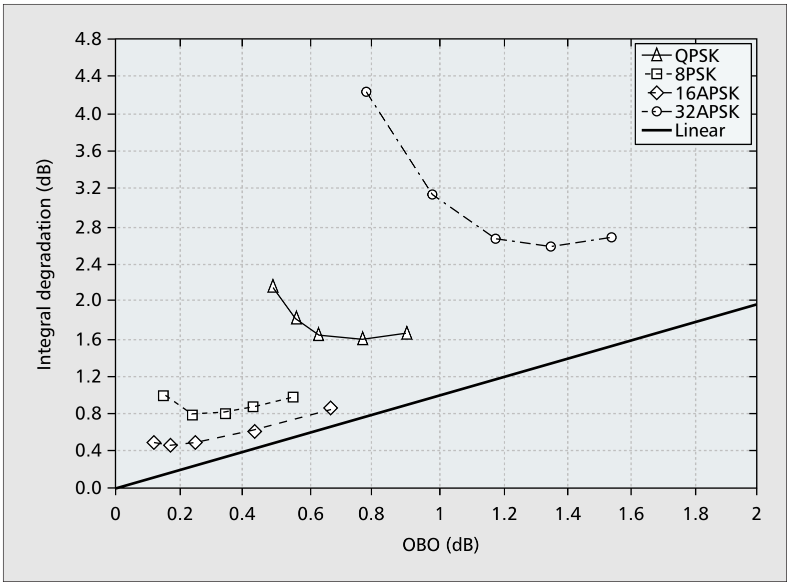

Figure 5. *Integral degradation vs. OBO for several modulation formats.*

The most significant overall achievements are summarized in Fig. 5, which reports the integral degradation for all modulation formats as a function of the HPA OBO. The integral degradation is a novel figure of merit that includes the overall E_b/N_0 loss with respect to the ideal channel case due to both linear/nonlinear distortion and channel parameter recovery imperfections (time, phase/frequency, and SNR). In the figure it can be noted that the minimum integral degradation is in the order of 0.47 dB for QPSK, 0.78 dB for 8-PSK, 1.61 dB for 16-APSK. and 2.59 dB for 32-APSK, respectively. These results confirm that the proposed design solutions are able to

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provide performance compliant with the DVB-S2 specifications.

CONCLUSIONS

This article provides general guidelines for the implementation of an adaptive physical layer compliant with the DVB-S2 standard. These guidelines can be summarized as follows. Frame synchronization can be performed through a single-dwell TC strategy coupled to a robust post detection integration scheme, such as DPDI, which exploits the SOF only, with no introduction of further complexity. Timing acquisition can be performed through the Gardner estimator, while coarse frequency estimation can be achieved through the Mengali and Morelli algorithm. Adaptivity is enabled by a SNORE algorithm to estimate the experienced SNR conditions. Fine carrier phase and frequency tracking is maintained through a hybrid PLL. By coupling the LDPC capabilities with the advanced predistortion techniques based on LUTs, data recovery performance shows surprisingly small integral degradation for all coding/modulation pairs. The interesting result is that by considering an actual Ka-band satellite link severely affected by strong linear and nonlinear distortion, and taking into account nonideal parameter estimation, it is still possible to fully exploit the adaptive physical layer potential, and to effectively achieve the capacity gains foreseen by the DVB-S2 standard.

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BIOGRAPHIES

GIANNI ALBERTAZZI (galbertazzi@arces.unibo.it) received a Dr.Ing. degree in telecommunications engineering in 2002 from the University of Bologna, Italy, defending a thesis on turbo coding techniques. In January 2002 he joined the Advanced Research Center for Electronic Systems (ARCES) of the same university as a Ph.D. student. His research activity is concerned with digital transmission, mobile and satellite CDMA systems, and especially channel coding techniques, with particular interests in turbo and LDPC code construction.

STEFANO CIONI (scioni@arces.unibo.it) received a Dr.Ing. degree in telecommunication engineering and a Ph.D. from the University of Bologna in 1998 and 2002, respectively. From March 2002 to October 2002 he was a visiting researcher at the European Space Agency (ESA) on ACM techniques for future broadband satellite networks. His research activities and interests include synchronization techniques, medium access control resource allocation algorithms, OFDM, and iterative decoding techniques joint with channel parameter estimation.

GIOVANNI E. CORAZZA (gecorazza@deis.unibo.it) received a Dr.Ing. degree (summa cum laude) in electronic engineering in 1988 from the University of Bologna and a Ph.D. in 1995 from the University of Rome "Tor Vergata," Italy. He is a full professor in the Department of Electronics, Computer Science, and Systems (DEIS) of the University of Bologna. He is Chairman of the Advanced Satellite Mobile Systems Task Force (ASMS-TF).

MASSIMO NERI (mneri@deis.unibo.it) received a Dr.Ing. degree in telecommunications engineering from the University of Bologna in 2002. Since October 2002 he has been with the Department of Electronics, Computer Science and Systems (D.E.I.S.) at the University of Bologna, where he started his Ph.D. study in 2003. His research activities and interests include nonlinearity mitigation techniques, multi-user detection, OFDM, and coding theory.

PAOLA SALMI (psalmi@ieee.org) received a Dr.Ing. degree in telecommunications engineering and a Ph.D. in electronics and computer science from the University of Bologna in 1996 and 2001, respectively. From 2001 to 2004 she was at the ARCES of the University of Bologna. During 2001 she was a summer manager at AT&T Shannon Laboratory, New Jersey. She currently works with the Optical Research Team of Andrew Wireless Systems. Her research activities include digital communication theory, 3G systems, synchronization techniques, and nonlinear effects on optical links.

ALESSANDRO VANELLI-CORALLI (avanelli@deis.unibo.it) received a Dr.Ing. Degree (cum laude) and a Ph.D. in electronics engineering from the University of Bologna in 1991 and 1996, respectively. In 1996 he joined the University of Bologna, where he is currently an assistant professor. During 2003 and 2005 he has been a visiting scientist at Qualcomm Inc., San Diego, California. He is responsible for the R&D group of the ASMS-TF.

MARCO VILLANTI (mvillanti@deis.unibo.it) received a degree in telecommunications engineering (summa cum laude) in 2001 and a Ph.D. in telecommunications in 2005 from the University of Bologna, where presently he is a post-doctoral researcher with DEIS and ARCES. He was a visiting scholar at the University of California, San Diego, in 2003 and 2005. His research interests are in communications theory, acquisition and synchronization for CDMA, TDMA, and UWB systems.

RAFFAELLA PEDONE (rpedone@deis.unibo.it) received a degree in electronic engineering (summa cum laude) from the University of Bologna in 2002, where she is presently a Ph.D. student in telecommunications with DEIS. Her interests are in digital communications, with particular focus on acquisition and synchronization for spread spectrum and TDMA systems.