Reed–Solomon Decoding Algorithms for Digital Audio Broadcasting in the AM Band

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***Abstract*—Digital audio broadcasting (DAB) systems for the AM band are being developed to provide higher-quality radio broadcasts, broader coverage, and data services. Transmission is typically done by means of multistreaming with QAM/OFDM in the high bits/sec/Hz regime, in which Reed–Solomon codes, either alone or in concatenation with inner trellis coded modulation (TCM), are natural choices for error control. For the AM DAB application, Reed–Solomon codes with suboptimal decoders of the bounded-distance type offer error correction important for bringing down the error probability, but also offers error detection important for generating block detected-error flags for use in the error concealment, or error mitigation, algorithm in the audio decoder. For multistream systems, the block detected-error flags can also be used to select the component streams retained by the audio decoder. In this paper, we evaluate the performance of Reed–Solomon codes in terms of both detected (flag) and undetected error rates for DAB applications. We provide simulation results for a variety of decoding algorithms, including hard-decision decoding (HDD), successive-erasure decoding (SED), and fixed-erasure decoding (FED), and discuss the channel environments in which each of these algorithms may be appropriate. Among other results, for example, we address the issue of matching the blocklength of the Reed–Solomon code to the audio decoder with its variable frame structure. From our simulation results for HDD, we conclude that the increased error correction capability of longer Reed–Solomon codes more than compensates for the mismatch in terms of error mitigation in the audio decoder. We also observe that SED provides only small improvements for uniform interference channels, while FED offers considerable improvements for some partial-band interference channels. We describe an appealing two-mode adaptive configuration using HDD and FED for mitigating partial-band interference caused by some second-adjacent AM broadcasts.**

***Index Terms*—Bounded-distance decoding, digital audio broadcasting (DAB), error mitigation, generalized minimum-distance decoding, joint source-channel coding, Reed–Solomon codes.**

I. INTRODUCTION

ANALOG and digital Hybrid In Band on Channel (HIBOC), as well as all-digital In Band on Channel (IBOC), Digital Audio Broadcasting (DAB) systems for the AM frequency band are currently being proposed and developed as a migration path from legacy analog AM radio broadcasts [1], [2]. The digital portions of these systems provide higher quality audio, broader coverage, and broadcast data services using existing licensed frequencies by combining sophisticated audio compression algorithms with robust channel coding and transmission techniques.

The Perceptual Audio Coder (PAC) [4]–[9] represents one of the audio coders proposed for HIBOC and IBOC AM systems. PAC encoding uses a perceptually-based bit allocation scheme for vector quantization, followed by lossless Huffman coding, to produce variable sized frames, with the frame length statistics varying as a function of the average audio coder rate [4]–[6]. When operating over noisy channels, PAC can reduce the impact of frames lost due to channel errors by employing error concealment (or error mitigation) techniques. As a simple example of error concealment, a lost frame can be filled in by interpolating between its surrounding frames. In addition to audio-based error-detection mechanisms, it is useful for the channel decoding subsystem to provide a flag indicating block detected errors for triggering the error concealment algorithms [4]. Generally speaking, long blocklength channel error-detection codes can lead to multiple audio frames being lost for a given uncorrectable channel error, while short blocklength channel error-detection codes can allow more frames to pass to the audio decoder with undetected errors—both of these scenarios lead to reduced audio quality. Matching the channel error-detection blocklength to the audio coder frame statistics for a given rate thus becomes an important problem in fine-tuning the performance of the overall system.

Because AM DAB systems operate in very bandwidth limited settings, audio coding rates in the range of 16–64 kb/sec along with multilevel modulation of e.g., M-QAM type, have been suggested. In this regime, natural channel coding options include trellis-coded modulation (TCM) and multi-level coding (MLC) [3], possibly in concatenation with outer Reed–Solomon (RS) codes [2]. For low-complexity implementations, RS alone may be used, while, for improved power efficiency, inner TCM or MLC may be added [2]. In either case, the error-detection capabilities of RS codes decoded with bounded-distance-type decoders can be exploited to providing a flag mechanism for error concealment in the audio decoder.

Motivated by a need to effectively match the RS channel code to the audio coder framelength for improved audio quality, this paper studies the performance of several Reed–Solomon decoding algorithms in terms of probability of flagging, or flag rate, as well as probability of undetected errors, or undetected error rate, for different blocklengths and code rates. Shortened codes are used to obtain codes of appropriate rate. The emphasis here is on low-complexity schemes for

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daytime transmission, for which multistream audio coders with orthogonal frequency division multiplex (OFDM) modulation have been proposed [2]. We model the channel in this situation with additive receiver thermal noise and, in some cases, partial band interference from first and second adjacent broadcasting stations. This interference may only occur in certain locations (certain transmitters in certain markets) and/or certain areas of the total coverage area for a given transmitter. Dramatic changes in channel conditions call for modified transmission modes for nighttime transmission. Specifically, skywave interference corrupts the reception, and for this case singlestream transmission at a lower audio coder rate is envisioned [2]. Detailed analysis of the nighttime transmission modes is beyond the scope of this paper.

We point out that similar HIBOC and all-digital IBOC developments are ongoing for transmission in the FM band [10]. More severe channel distortions such as multipath fading as well as substantially more available bandwidth mandate different audio coding rates and channel coding configurations for the HIBOC FM case. In particular, several system proposals employ an inner convolutional code and an outer high-rate cyclic redundancy check (CRC) code. For these systems, the outer code is employed only for the purpose of error-detection. More generally, certain digital cellular systems have employed similar coding configurations.

An outline of this paper is as follows. Section II describes our model for the characteristics of the AM band channel and the coding and modulation methods considered for daytime transmission. Section III summarizes three bounded-distance-type decoding algorithms, including hard-decision decoding (HDD), successive-erasure decoding (SED), and fixed-erasure decoding (FED), and provides simulation results of their performance in several channel scenarios. Finally, Section IV discusses the results and offers some concluding remarks.

II. System Model

The AM-band radio channel is currently not as well-modeled as the FM band channel. AM-band transmissions appear to be subject to mild fading, and significant distortions occur when mobiles travel near powerlines or through underpasses. Additionally, daytime and nighttime transmission environments differ dramatically due to strong nighttime skywave reflections, causing inter- and intra-channel interference. In this paper, we model the AM band channel as an additive white Gaussian noise channel with slowly-varying signal-to-noise ratio (SNR), a reasonable model for daytime transmission in the absence of underpasses and powerlines. We also consider potential interference from first and second adjacent AM broadcasts, that lead to, e.g., partial band interference at the band edges.

Fig. 1 illustrates example spectra of daytime transmission systems for multistream HIBOC AM as proposed in [2]; the allocation with $B = 15$ kHz can carry a higher data rate provided there is room for the wider bandwidth, and the allocation with $B = 10$ kHz may be required for more crowded markets. By using "multistream transmission"—audio encoding performed using multidescriptive and embedded techniques in conjunction with transmission in several separate frequency bands via, e.g., OFDM—the effects of noise in highly-variable partial band interference channels can easily be mitigated. Such partial band interference might arise, e.g., because the host analog AM signal is transmitted simultaneously with a digital stream, and second adjacent interference from other analog broadcasts most likely distorts the highest and/or lowest frequency components in the digital transmission, as depicted in Fig. 2. There are several other configurations that may be of interest for practical HIBOC AM systems; see [2] for details. We consider only the allocations shown in Fig. 1 for the purposes of evaluation of Reed–Solomon code performance.

Figs. 3 and 4 show conceptual block diagrams for the transmitter and receiver, respectively, of a multistream HIBOC AM system [2]. Multistream DAB systems partition the frequency band into several subbands for transmission, e.g., by suitably grouping together subcarriers in an OFDM modulation scheme. The audio coder may combine embedded and/or multidescriptive techniques, and different streams of the audio bitstream are channeled over certain of the subbands. Fig. 3 shows an example in which the audio information is partitioned into three

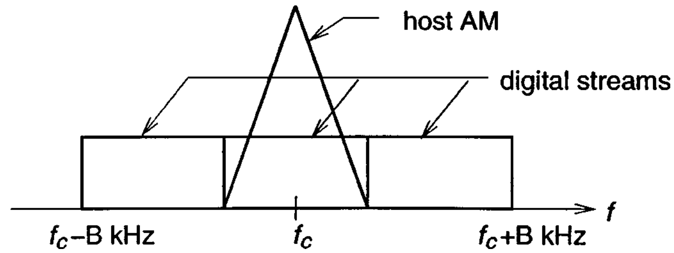

Fig. 1. Example multistream HIBOC AM system with three streams and a system bandwidth of $2B$ kHz, where $B = 10$ kHz or $B = 15$ kHz depending upon the available transmission bandwidth.

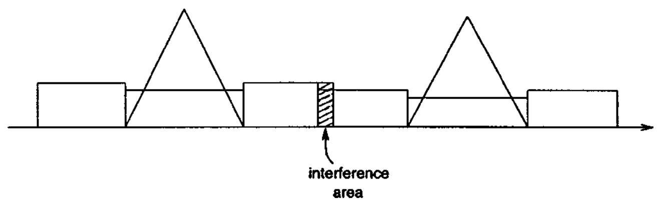

Fig. 2. Diagram illustrating second-adjacent interference between two stations.

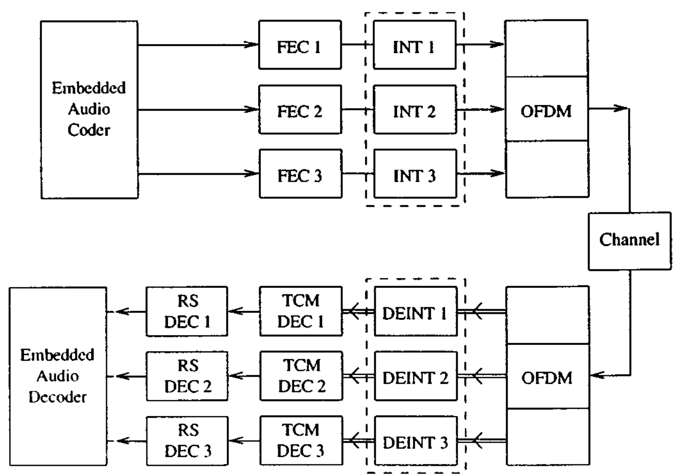

Fig. 3. Multistream system transmitter block diagram.

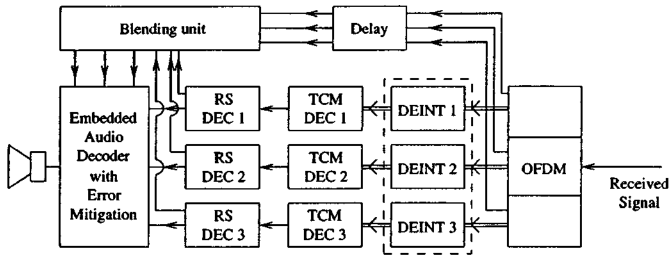

Fig. 4. Multistream system receiver block diagram. The OFDM demodulator also contains synchronization, training, equalization, timing, etc.

streams, with the "core" information transmitted in the middle subband and multidescriptive versions of the "enhancement" on the sidebands [2]. Because the broadcasting environment is varying with location but using only one standard transmission format, the main motivation for the multistream approach is flexible and adaptive interference management. The locations with very strong partial band interference lead to cases where certain parts of the frequency band is muted [1], [2]. A singlestream system with one audio bitstream and one transmission band may on the other hand be completely overwhelmed by the same partial band interference. In Fig. 3, FEC 1 to FEC 3 indicate in the general case a concatenated RS and TCM encoder; the decoders consist of a TCM and a RS decoder. In the general case there are also interleavers and deinterleavers present, although we do not consider interleaving in this paper. While the concept of multistreaming is important for the design of a viable AM (and FM) HIBOC system, the results presented in this paper apply to both singlestream and multistream systems, as we consider a single RS code that performs the functions of error-correction and error-detection for error concealment.

The narrowband channel environment and audio coder bit-demand for high-quality daytime transmission require either uncoded 16-QAM or rate $R = 4/5$ trellis coded 32-QAM signaling [2]. Outer Reed–Solomon codes [4] are good candidates for providing additional channel robustness via error-correction as well as a flag mechanism for error concealment in the audio decoder via error-detection. Using Reed–Solomon codes defined over GF(2^8), i.e., having 8 bits per symbol, we may map two, 4-bit modulation symbols into a single Reed–Solomon code symbol. We focus on the case of uncoded 16-QAM, the performance of which is shown in Fig. 5 for reference, with the understanding that $R = 4/5$ trellis coded 32-QAM should improve coverage at the cost of additional complexity. For 32-QAM without inner TCM, Reed–Solomon codes defined over GF(2^{10}) would be suitable candidate codes.

The channel and audio coder bit-rates in proposed systems allow for Reed–Solomon codes with rates of $R = 4/5$ and $R = 9/10$, for example. For codes defined over GF(2^8), the maximum blocklength is $2^8 - 1 = 255$ symbols, or 2040 bits. Codes with smaller blocklength can be obtained via shortening, i.e., encoding and decoding the full blocklength code with a given number of symbols s fixed at zero. To achieve rate R approximately for a given blocklength n, there two options for the information blocklength k, namely, $\lceil k \rceil = \lceil nR \rceil$, resulting in a slightly higher rate, or, $\lfloor k \rfloor = \lfloor nR \rfloor$, resulting in a slightly lower rate.

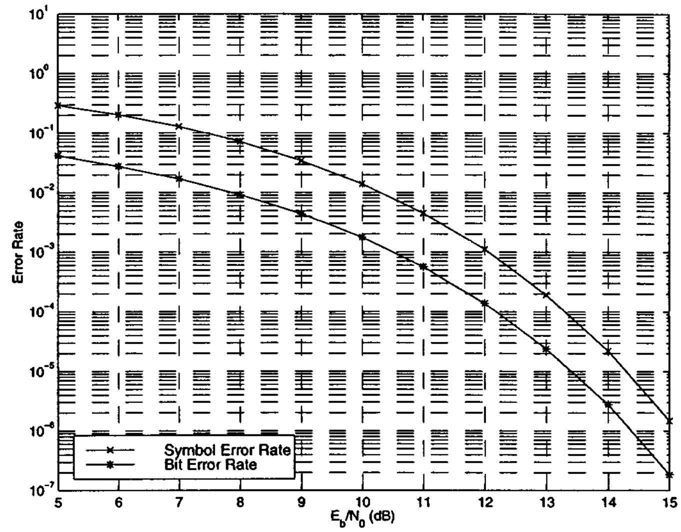

Fig. 5. Uncoded 16-QAM symbol- and bit-error rates on the AWGN channel.

TABLE I
CANDIDATE REED–SOLOMON CODES OVER GF(2^8) WITH RATES AROUND $R = 4/5$ AND $R = 9/10$

R	(n, k)	s	Rate	Blocklength in bits
4/5	(15, 12)	240	0.8	120
4/5	(30, 24)	225	0.8	240
4/5	(64, 51)	191	0.7969	512
4/5	(128, 102)	127	0.7969	1024
4/5	(255, 204)	0	0.8	2040
9/10	(20, 18)	235	0.9	160
9/10	(30, 27)	225	0.9	240
9/10	(67, 60)	188	0.8955	536
9/10	(128, 115)	127	0.8984	1024
9/10	(255, 229)	0	0.8980	2040

Table I shows candidate Reed–Solomon codes within 0.5% redundancy of the required rates and having blocklengths (in bits) roughly corresponding to those examined in [4], [10] for the FM system.

III. REED–SOLOMON DECODING ALGORITHMS AND PERFORMANCE

To discuss Reed–Solomon decoding algorithms in a common framework, we introduce some notation.[1] For a more thorough introduction to these codes, see [11], [12]. Given k information symbols, the transmitter chooses the appropriate codeword $\mathbf{x} = [x_1\ x_2\ \cdots\ x_n]^T$ from an (n, k) Reed–Solomon code over GF(2^8). This codeword is appropriately modulated, transmitted over the channel, and demodulated as discussed Section II. Due to channel distortion, the received signal $\mathbf{y} = [y_1\ y_2\ \cdots\ y_n]^T$ generally differs from the transmitted codeword, and the receiver must form an estimate $\hat{\mathbf{x}}$ of the transmitted codeword from the received signal.

[1]To differentiate between scalar and vector quantities, the latter is denoted in boldface. For example, the vector of scalars $x_1, x_2, \ldots, x_n$ is denoted by $\mathbf{x}$.

Maximum-likelihood decoding is optimal in terms of minimizing the probability that the receiver's estimate $\hat{\mathbf{x}}$ differs from the transmitted codeword $\mathbf{x}$; however, maximum-likelihood decoding has complexity that grows exponentially in the information blocklength and, importantly for audio applications, does not offer a flag mechanism for error concealment in the audio decoder. As practical alternatives to maximum-likelihood decoding, we consider decoding algorithms based on algebraic bounded-distance decoders such as, e.g., hard-decision decoding (HDD), successive-erasure decoding (SED), and fixed-erasure decoding (FED), of the Reed–Solomon codes listed in Table I, all of which offer block error flags.

These algorithms work with a decision vector $\mathbf{r} = [r_1 \ r_2 \ \cdots \ r_n]^T$, r_i being the maximum-likelihood *symbol* decision based only upon the received signal y_i. The decision vector $\mathbf{r}$ takes values in GF$(2^8)^n$, thereby allowing for reduced complexity algebraic error-correction algorithms to be applied to obtain an estimate $\hat{\mathbf{x}}$ of the transmitted codeword. When the number of symbol decision errors between $\mathbf{r}$ and $\mathbf{x}$ is less than the decoding radius $\lfloor (n-k)/2 \rfloor$ of the code, an errors-only decoding routine decodes to the correct codeword, so that $\hat{\mathbf{x}} = \mathbf{x}$. When the number of decision errors between $\mathbf{r}$ and some other codeword $\mathbf{x}'$ is less than the decoding radius of the code, an errors-only decoding routine decodes to the incorrect codeword $\mathbf{x}' \neq \mathbf{x}$, contributing to the undetected error rate P_u of the code. If neither of the above two conditions are met, an errors-only decoding routine flags a block error, contributing to the flag rate P_f of the code. Traditionally, the performance of errors-only decoding is evaluated in terms of the sum $P_e = P_f + P_u$, often referred to as the probability of not decoding correctly or simply block error rate, but for audio applications we are interested in the flag and undetected error rates individually. Error events are defined similarly for an errors-and-erasures decoding routine, but with a larger decoding radius depending upon the number of erasures.

We now summarize HDD, SED, and FED, and present simulation results for the candidate Reed–Solomon codes in Table I. Our simulations utilize the algebraic errors-only and errors-and-erasures decoding routines provided in [13] for decoding Reed–Solomon codes. Both routines can be used for shortened codes by simply padding the input words with zeros and removing these zeros at the output of the decoding routine. However, the decoding routines, particularly the errors-and-erasure decoding routine, may decode to a codeword without the leading zeros; instead of accepting this codeword, a decoding failure results.

A. Hard-Decision Decoding

Hard-decision decoding (HDD) applies an algebraic errors-only decoding algorithm to the symbol decision vector $\mathbf{r}$ to generate either a codeword estimate $\hat{\mathbf{x}}$ or a block error flag.

Fig. 6 shows flag rates P_f and undetected error rates P_u for HDD of the $R = 4/5$ Reed–Solomon codes over GF(2^8) listed in Table I. Fig. 7 shows similar results for the $R = 9/10$ codes. These simulation results suggest that using longer codes of the same rate with hard-decision decoding reduces the block error, flag, and undetected error rates at moderate to high SNR. These results are in contrast to the results obtained in [4] for CRC codes with error detection only: increasing the blocklength increased the flag rate and decreased the undetected error rate, leading to the choice of a medium-blocklength CRC code for better matching to the audio decoder.

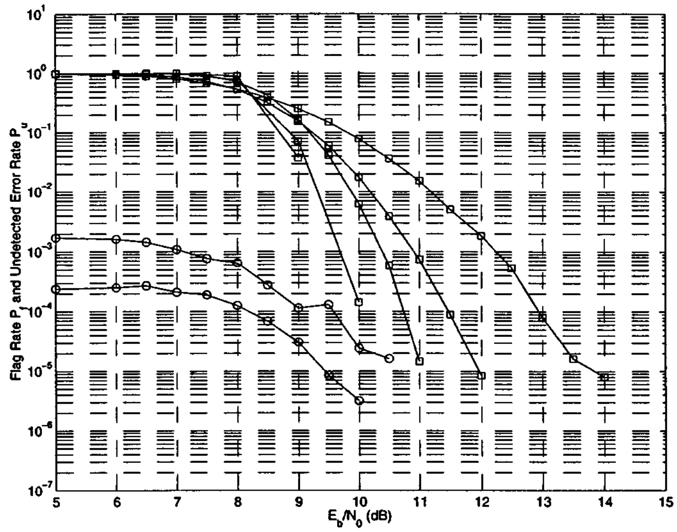

Fig. 6. Flag error rate P_f (squares) and undetected error rate P_u (circles) versus E_b/N_0 in dB for the $R = 4/5$ Reed–Solomon codes with hard-decision decoding (HDD). The successively lower curves correspond to the longer blocklength codes listed in Table I.

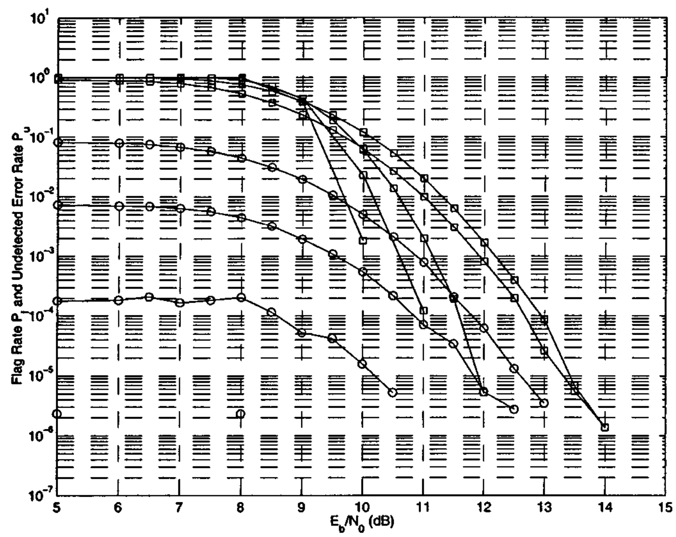

Fig. 7. Flag error rate P_f (squares) and undetected error rate P_u (circles) versus E_b/N_0 in dB for the $R = 9/10$ Reed–Solomon codes with hard-decision decoding (HDD). The successively lower curves correspond to the longer blocklength codes listed in Table I.

B. Successive-Erasure Decoding

Increasing the blocklength of the Reed–Solomon code can reduce the undetected error probability beyond that required by the audio decoder. For example, for the $R = 9/10$, $(67, 60)$ Reed–Solomon code, there appears to be nearly four orders of magnitude between the relative frequency of block error flags and undetected errors, while the audio decoder only needs roughly two or three orders of magnitude separation [2]. Audio quality can be significantly improved by reducing the flag rate as much as possible, while maintaining reasonable undetected error rates. Successive-erasure decoding (SED) [14]–[16]

should reduce the flag rate at the expense of increasing the undetected error rate, and potentially allow for a better match to the error concealment algorithm in the audio decoder. SED applies an algebraic errors-and-erasures decoding algorithm to the symbol decision vector $\mathbf{r}$ with increasingly more of the symbol decisions erased, thereby generating a list of candidate codewords from which the "best" candidate can be selected.

In our SED algorithm, successive symbol decision erasures are generated from continuous-valued reliability information obtained from the symbol decision device. In the most general setting [16], the reliability information is captured by a matrix, $\mathbf{A} = [\boldsymbol{\alpha}_1 \; \boldsymbol{\alpha}_2 \; \cdots \; \boldsymbol{\alpha}_n]$, where $\boldsymbol{\alpha}_i = [\alpha_{i,1} \; \alpha_{i,2} \; \cdots \; \alpha_{i,2^8}]^T$ is a vector indicating the reliability of the all possible symbol decisions. Specifically, for the received sample y_i, $\alpha_{i,j}$ indicates the reliability, e.g., likelihood or (generalized) distance, of making the hard-decision that symbol j was transmitted in slot i. We note that the earliest methods [14], [15] fuse the information in each vector $\boldsymbol{\alpha}_i$ into a single scalar value indicating the reliability of the hard-decision r_i. The method in [16] maintains separate reliabilities for all possible symbol decisions in each slot, and, in particular, appears to be more effective for nonbinary signaling alphabets as employed in our application. SED reduces computation over maximum-likelihood decoding by never combining the reliability information in the matrix $\mathbf{A}$ into 2^k codeword reliabilities.

To briefly summarize the algorithm of [16], candidate codewords are generated via successive erasures as follows:

1) Each Reed–Solomon symbol in GF(2^8) is transmitted as two consecutive 16-QAM symbols, so for each RS-symbol i, the 16-QAM demodulator computes squared Euclidean distances $d^2_{1,i,k}$ and $d^2_{2,i,k}$, $k = 0, \ldots, 15$ between the two complex-valued channel outputs and all of the symbols in the constellation. All possible pairs $(d^2_{1,i,k}, d^2_{2,i,k'})$ of distances are added to form the Reed–Solomon reliability vector $\boldsymbol{\alpha}_i$.
2) For each RS-symbol, construct the index set
$$j_{i,1}, j_{i,2}, \ldots, j_{i,2^8}$$
such that
$$\alpha_{i,j_{i,1}} < \alpha_{i,j_{i,2}} < \cdots < \alpha_{i,j_{i,2^8}}.$$
Then Reed–Solomon symbol $j_{i,1}$ is the minimum-distance (maximum-likelihood) symbol decision.
3) Let $\beta_i = \alpha_{i,j_{i,2}} - \alpha_{i,j_{i,1}}$, so that β_i represents the reliability of the second-most likely symbol decision relative to that of the most likely symbol decision. Small β_i implies that the decision in RS-symbol i is unreliable, because the received signal is nearly equidistant to at least two possible symbols. Construct the index set
$$i_1, i_2, \ldots, i_n$$
such that
$$\beta_{i_1} < \beta_{i_2} < \cdots < \beta_{i_n}.$$
Then RS-symbol i_1 contains the most unreliable Reed–Solomon symbol decision, RS-symbol i_2 contains the second-most unreliable Reed–Solomon symbol decision, and so forth.
4) Set $e = 0$.

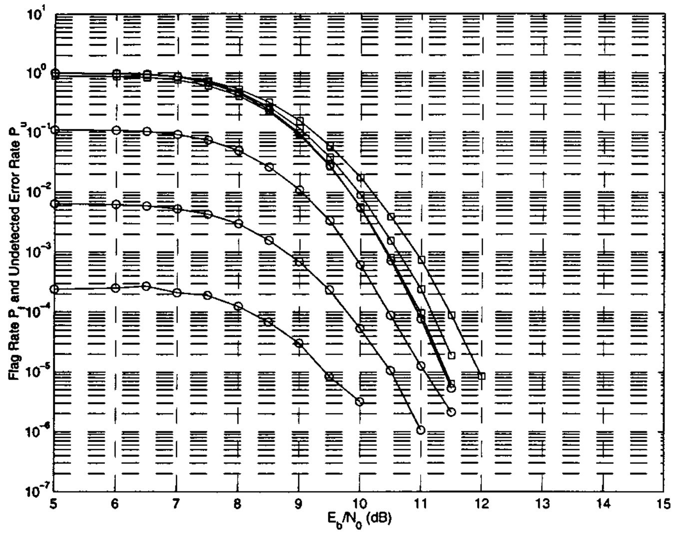

Fig. 8. Flag error rate P_f (squares) and undetected error rate P_u (circles) versus E_b/N_0 in dB for the $R = 4/5$, $(30, 24)$ Reed–Solomon code with successive-erasure minimum-distance decoding (SEMDD). The successively lower squared curves and successively higher circled curves correspond to growing the list of candidate codewords by one.

5) Erase RS-symbols $i_1, i_2, \ldots, i_e$ and apply an errors-and-erasures decoding routine to generate a potential candidate codeword $\hat{\mathbf{x}}_e$ for the list.
6) If $e < (n-k)$, set $e = e + 2$ [14] and goto step 5); otherwise, select from among the candidate codewords on the list.

SED may select from the list of candidates the "best" codeword according to one of several criteria. For example, a successive-erasure bounded-distance decoding (SEBDD) algorithm chooses the unique codeword within the generalized decoding radius of the code [14]–[16], if this codeword can be found on the list; otherwise, SEBDD flags a block error. As another example, successive-erasure minimum distance decoding (SEMDD) chooses the codeword on the list with smallest Euclidean distance from the received signal [15], if any codewords can be found on the list; otherwise, it flags a block error. In this paper, we focus on SEMDD because it offers lower flag rates and more closely approximates maximum-likelihood decoding than SEBDD on an additive white Gaussian noise channel.

The SED algorithm outlined above can be readily modified to accommodate a concatenated coding scheme using outer Reed–Solomon codes over GF(2^8) and inner $R = 4/5$ trellis coded 32-QAM; however, such a system requires a soft-output Viterbi algorithm (SOVA) for TCM, adding to the decoder complexity in a concatenation approach. We also point out that, although in our simulations we execute the errors-and-erasures decoding routine to generate each possible candidate codeword, several authors have developed more computationally efficient SED algorithms that may be useful for implementation purposes [17]–[19].

Fig. 8 shows the flag rate P_f and undetected error rate P_u for SEMDD of the $R = 4/5$, $(30, 24)$ Reed–Solomon code over GF(2^8) listed in Table I. Fig. 9 shows similar results for the $R = 9/10$, $(30, 27)$ code. These results suggest that, for 16-QAM modulation and these Reed–Solomon code rates, SED

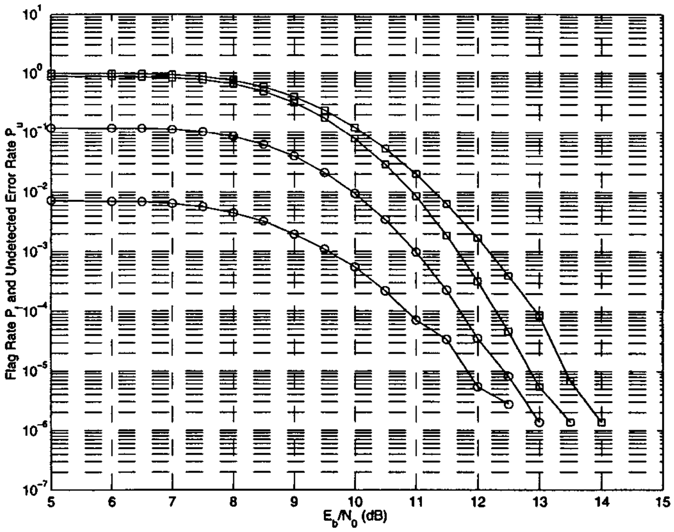

Fig. 9. Flag error rate P_f (squares) and undetected error rate P_u (circles) versus E_b/N_0 in dB for the $R = 9/10$, $(30, 27)$ Reed–Solomon code with successive-erasure minimum-distance decoding (SEMDD). The successively lower squared curves and successively higher circled curves correspond to growing the list of candidate codewords by one.

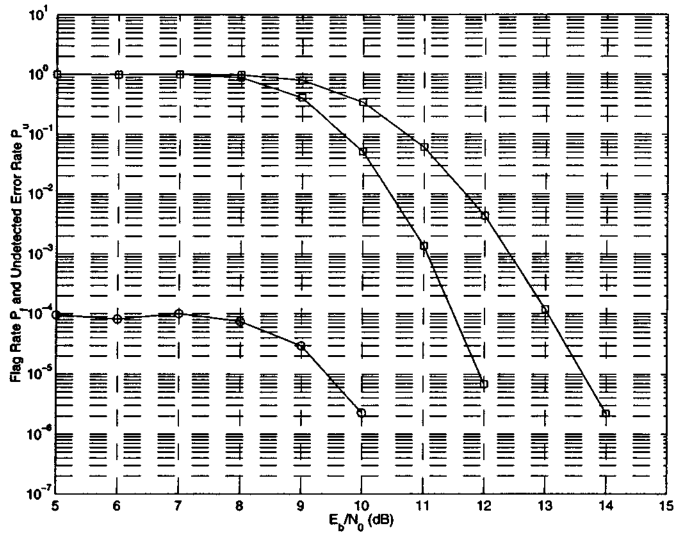

Fig. 10. Flag error rate P_f (squares) and undetected error rate P_u (circles) versus E_b/N_0 in dB for the $R = 4/5$, $(64, 51)$ Reed–Solomon code with hard-decision decoding (top curve) and fixed-erasure decoding (bottom curves) in the presence of partial band interference. The SNR in the 10 jammed 16-QAM symbols (5 RS symbols) is -10 dB relative to the E_b/N_0.

offers at most 0.5 dB at the expense of much higher undetected error rates; hence, the added complexity of SED may not be worthwhile for digital audio applications in the AM band.

C. Fixed-Erasure and Errors Decoding

For fixed partial-band interference situations as depicted by Fig. 2, we also consider fixed-erasure decoding (FED), a decoding algorithm similar to SED except that a specific set of symbols are *always* erased, and at most one candidate codeword is generated. For uniform interference channels, SED is generally more effective because it has more flexibility in terms of the symbols it erases. For nonuniform interference, particularly for partial-band interference, channels, as arise, e.g., in the HIBOC AM system at some locations with second-adjacent interference [2], FED may be more effective than HDD and SED because, given side information about which symbols contain higher levels of interference, FED can erase those symbols that are known to be less reliable, rather than the general procedure performed by SED. Then errors and erasure decoding is performed for the Reed–Solomon code.

An adaptive system employing HDD and FED might operate as follows. A pilot tone is used to indicate the presence of overlapping interference in the highest and/or lowest tones, and this measurement provides the decoding algorithm with the number of tones that are subjected to interference. The pilot could be a fixed pilot tone for the lowest and highest frequency or a “roaming pilot tone” which is on only part of the time for a certain frequency and which jumps around among the tones. With no partial band interference, HDD can be employed. When the pilot indicates that partial-band interference is present, the decoder switches over to FED with the symbols transmitted on the interfered tones erased. We note that the complexity increase for this two mode operation is marginal, but requires modification to the transmitter for the insertion of pilot tones. Once a mode has been selected, the decoder remains in that mode until the channel measurements from pilot data suggest a mode change. When the pilot indicators are close to their decision threshold, the decoder can, in principle, decode using both HDD and FED and select the candidate codeword, if either, that decodes successfully.

Fig. 10 shows the flag rate P_f and undetected error rate P_u for HDD and FED of the $R = 4/5$, $(64, 51)$ Reed–Solomon code over GF(2^8) in the presence of partial band (5 RS symbols) interference (-10 dB lower SNR relative to the E_b/N_0 number). As we have seen, the HIBOC AM systems proposed in [2] use OFDM modulation. In this case 5 out of 64 RS symbols, or 10 16-QAM symbols, are exposed to severe interference. This will typically be the highest or lowest frequencies in the HIBOC AM spectrum. The performance of HDD and FED are shown, with the understanding that the performance of SED would be, at best, that of FED. This example set of simulations suggests that FED can be very beneficial in HIBOC systems in the AM band, when the receiver can obtain enough knowledge of the channel conditions for FED to be appropriate.

Fig. 11 shows the same case as in Fig. 10, but without partial band interference. The better of the two upper error probability curves in Fig. 11 shows the performance with HDD. The worse of the two curves shows the performance with FED in this case, where no partial band interference is present. Clearly, there is no need for FED in this case, but Fig. 11 shows the penalty in performance when the incorrect mode is chosen. (Also note that the lower of the two upper curves in Fig. 10 is identical to the upper curve in Fig. 11.) Fig. 11 also shows the undetected error rate P_u for the FED case without interference.

IV. Discussion and Conclusions

Designing error control systems for digital audio applications depends on many parameters and channel conditions. Among these parameters are the audio coding rate, type of error mitigation and screening for undetected errors, channel coding rate, channel decoding algorithm, system complexity,

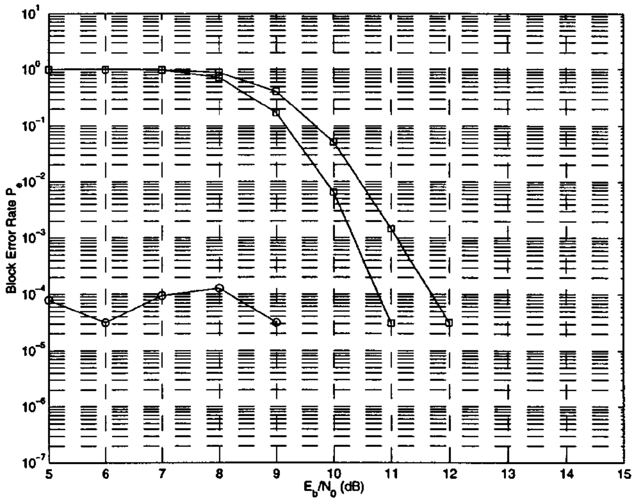

Fig. 11. Flag error rate P_f (squares) and undetected error rate P_u (circles) vs. E_b/N_0 in dB for the $R = 4/5$, $(64, 51)$ Reed–Solomon code with hard-decision decoding (bottom curve, squares) and fixed-erasure decoding (top curve, squares and curve with circles) in the presence of background noise without partial band interference.

etc. We have considered a number of these issues with respect to Reed–Solomon codes for HIBOC systems in the AM band, either in isolation with uncoded 16-QAM modulation or, in principle, in concatenation with $R = 4/5$ trellis coded 32-QAM.

In particular, we have examined several Reed–Solomon decoding algorithms that use different levels of soft symbol information and modem erasures. In contrast to the error-detecting CRC codes optimized for HIBOC systems in the FM band, longer error-detecting and correcting Reed–Solomon codes are preferred for HIBOC systems in the AM band. When no inner trellis coded modulation scheme is used, some error correction is preferred in the Reed–Solomon decoding. At a given rate, the longest Reed–Solomon code appears to give the best performance, since the flag probability is much lower than for a codeword length better matched to the audio decoder average frame length. Such a choice renders the error concealment less efficient on the rare occasions it is called for, because multiple audio frames can be erased as a result of an uncorrectable error in a long channel code block. The probability of undetected errors also decreases with code word length at a given rate, which is another benefit. Successive-erasure decoding appears to offer little benefit for the additional complexity, at the high rates and long blocklengths considered for the HIBOC AM systems in this paper. However, for nonuniform interference channels, as may arise in HIBOC AM systems with significant second-adjacent analog interference, successive-erasure and especially adaptive fixed-erasure decoding methods offer considerable improvement over hard-decision decoding.

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