ROBUST MODEM AND CODING TECHNIQUES FOR FM HYBRID IBOC DAB

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ABSTRACT

The design of robust modem and FEC (Forward Error Correction) code techniques with application to the transmission of an FM Hybrid analog/digital In-Band On-Channel (IBOC) Digital Audio broadcast (DAB) signal is presented here. The FEC codes are derived from an original lower rate convolutional code (R=1/3). The original code is segmented into a pair of "complementary" components, which form independent codes, each with a higher rate (less redundancy) than the base code. The exploitation of Channel State Information (CSI) and special interleaving techniques are described for application to FM Hybrid IBOC DAB with its unique interference environment and selective fading due to multipath. Simulation results confirm the robustness of the design.

I. INTRODUCTION

The focus of this paper is on the FEC, interleaving, CSI estimation and adaptive symbol weighting for the FM Hybrid IBOC DAB system. A system overview along with analysis and simulation results on the mutual interaction among analog FM and DAB signals, adjacent channel interference, and time diversity of the analog FM and DAB signals were presented at the 51st Annual Broadcast Engineering Conference (NAB) in April, 1997 [1].

Forward error correction and interleaving improve the reliability of the transmitted digital information over a corrupted channel. Complementary Pair Convolution (CPC) FEC code techniques were developed for ARQ schemes where retransmissions were coded using complementary codes instead of simply retransmitting the same coded sequence [2]. Construction of CPC codes is derived from previously published puncturing techniques [3,4,5]. The CPC code technique allows the individual transmissions to be combined to form a more powerful code than the sum of the individual transmissions.

IBOC DAB is an ideal candidate for the application of CPC codes since the digital DAB transmission is accomplished over two sidebands (upper sideband and lower sideband) which are potentially impaired by nearly independent interferers with independent fading. If one sideband is completely corrupted by a strong first adjacent FM signal in the vicinity of the receiver, the opposite sideband must be independently decodable at the receiver. Therefore each sideband must be coded with an independently decodable FEC code. However, when both sidebands contain useful information that is not completely corrupted by an interferer, then CPC codes provide additional coding gain above that achieved by power combining the two sides. Furthermore, CSI and OFDM interleaving techniques have been developed to deal with the unique interference and selective fading characteristics of the FM IBOC DAB channel.

FM Hybrid IBOC DAB Parameters

A brief description of the FM IBOC DAB system is presented. Details of the FM IBOC system presented here are updated since the presentation of a previous paper [1]. The PSD of a typical FM broadcast signal is nearly triangular with a slope of about -0.35 dB/kHz from the center frequency. First adjacent FM signals, if present, would be centered at a spacing of 200 kHz.

Although FM channel spacing in some countries is 100 kHz, these first adjacents are geographically separated such that FM reception is not impaired within the coverage area. Therefore this should pose no problem to the FM IBOC system. The DAB to DAB interference at 300 kHz spacing can impair performance on one sideband, but the CPC code is designed to tolerate this condition.

In the baseline FM IBOC design, 95 OFDM subcarriers are placed on each side of the host FM signal occupying the spectrum from about 130 kHz through 199 kHz away from the host FM center frequency as shown in Figure 1.

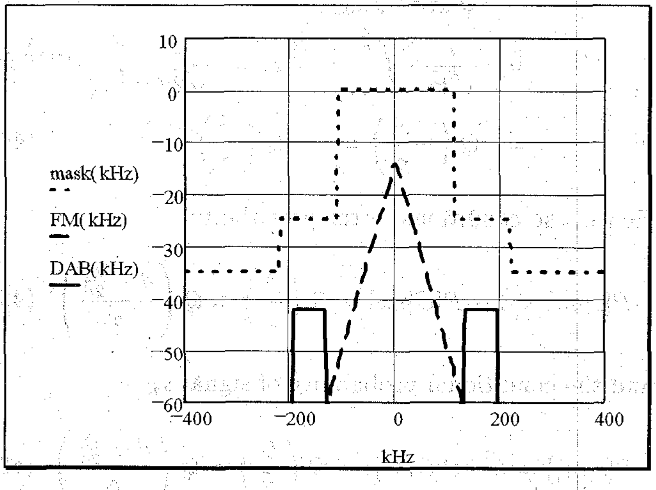

Figure 1. Power spectral densities of FM and DAB signals below FM spectral mask.

The total DAB power in each sideband is set to about -25 dB relative to its host FM power. The individual OFDM subcarriers are QPSK modulated at 689.0625 Hz (44100/64) and are orthogonally spaced at about 726.7456055 Hz (44100*135/8192) after pulse shaping is applied (root raised cosine time pulse with 7/128 excess time functions as guard time). The potential subcarrier locations are indexed from zero at the FM center frequency to plus or minus 275 at the edges of the 400 kHz bandwidth. The outside assigned subcarriers are at plus or minus 274 with a center frequency of plus or minus 199128 Hz. The inside information bearing subcarriers of the baseline system are located at plus or minus 179 with center frequencies of plus or minus 130087 Hz. The pilot subcarriers are located at plus or minus 178 with center frequencies of plus or minus 129361 Hz. The subcarriers are differentially coded across frequency using the inside pilot subcarriers as the reference for the first differentially detected symbol. These reference (pilot) subcarriers are modulated with an alternating sequence to permit assistance in frequency and symbol timing acquisition and tracking. Recent evolution of the design shows that adequate acquisition and tracking performance is achievable without the pilots.

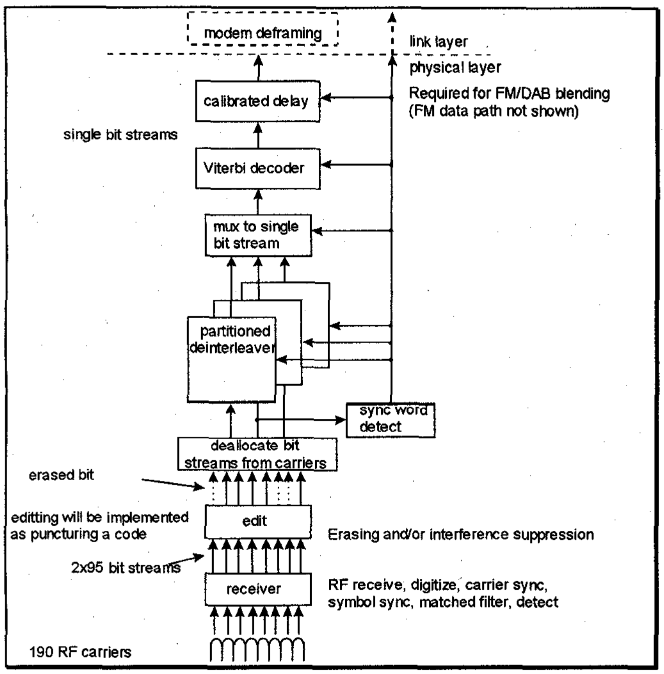

Figure 2. Functional diagram showing the mapping and processing of bits through the receiver, deinterleaver, and FEC decoder.

In the presence of adjacent channel interference, the outer OFDM subcarriers are most vulnerable to corruption, and the interference on the upper and lower sidebands is independent. Since the PSD of an FM broadcast signal is nearly triangular, then the interference increases as the OFDM subcarriers approach the frequency of a first adjacent signal. The coding and interleaving are specially tailored to deal with this nonuniform interference such that the communication of information is robust.

The IBOC DAB system will transmit all the digital audio information on each DAB sideband (upper or lower) of the FM carrier. Although additional subcarriers beyond the baseline system can be activated to enable the transmission of all the code bits of the rate 1/3 FEC code, the baseline system employs a code rate of 2/5. Each sideband can be detected and decoded independently with an FEC coding gain achieved by a rate 4/5 (optionally rate 2/3) convolutional code. This redundancy permits operation on one sideband while the other is corrupted. However, usually both sides are combined to provide additional signal power and coding gain commensurate with a rate 2/5 (optionally rate 1/3) code. Furthermore special techniques can be employed to demodulate and separate strong first adjacent interferers such that a "recovered" DAB sideband can supplement the opposite sideband to improve coding gain and signal power over any one sideband.

A simplified functional block diagram of the flow of the demodulated bits in an FM IBOC receiver is shown in Figure 2.

II. CSI AND ADAPTIVE WEIGHTS

Soft-decision Viterbi decoding with (near) optimum soft-decision weighting for maximum ratio combining (MRC) for differentially detected QPSK subcarrier symbols is employed to minimize losses over the channel. Since the interference and signal levels vary over the subcarriers (frequency) and time due to selective fading, timely CSI is needed to adaptively adjust the weighting for the soft-symbols. The CSI estimation technique should be designed to accommodate a fading bandwidth of up to 13 Hz for typical vehicle speeds in the FM band around 100 MHz.

An expression for the weighting factor can be derived assuming gaussian noise into a differential QPSK detector resulting in non-gaussian statistics at the output. The fading factor can be computed as a function of the statistics of the output of the differential detector where we define the soft decision of the form

$$S = (a + n_1) \cdot (a \cdot e^{j \cdot \phi} + n_2) \tag{1}$$

where ϕ denotes the phase information imposed between a pair of adjacent symbols in the differential encoding, and n are the independent noise samples. The fading factor a of the adjacent symbols is assumed to be approximately equal. The signal to noise ratio after differential detection is easily computed to be

$$SNR = \frac{a^4}{2 \cdot a^2 \cdot \sigma^2 + \sigma^4} \tag{2}$$

The ideal weighting factor for the post-differentially detected symbols is therefore

$$w = \frac{a^2}{2 \cdot a^2 \cdot \sigma^2 + \sigma^4} \tag{3}$$

The first differential approach described here uses statistical estimates of the second and fourth moments of the differentially detected symbol magnitudes to form the weighting factor. These second and fourth are described by the following relationships.

$$E\{|S|^2\} = (a^2 + \sigma^2)^2$$
$$E\{|S|^4\} = (a^4 + 4 \cdot a^2 \cdot \sigma^2 + 2 \cdot \sigma^4)^2 \quad (4)$$

Then the fading factor can be estimated as

$$\hat{a}_k = \sqrt[4]{2 \cdot E\{|S_k|^2\} - \sqrt{E\{|S_k|^4\}}} \quad (5)$$

and the noise can be estimated as

$$\hat{\sigma}^2{}_k = \sqrt{E\{|S_k|^2\}} - \hat{a}_k^2 \quad (6)$$

The estimates of equations (5) and (6) are inserted into equation (3) to obtain the weight.

Simulations were performed using adaptive weighting as described in equations (3), (5), and (6). Although long-term estimates without fading yielded good results, a compromise must be reached between long filter time constants for accurate estimation versus short filter time constants needed to track varying statistics due to fading.

In our Digital Audio Broadcasting simulation, the OFDM symbol rate of 689.0625 Hz was chosen with a fading bandwidth of 13 Hz. Then the reciprocal of the fading bandwidth is about 53 symbols in this case. A filter time constant of 16 symbols was chosen since this time constant must be small compared to the fading time. Unfortunately, the statistical estimation errors over this short filter time yielded poor performance results for the adaptive weighting compared to what would be possible with perfect statistical estimation. Even reducing the fading bandwidth down to 3 Hz and increasing the filter time constant to 64 samples left a significant loss.

Equations (5) and (6) reveal that, in effect, quantities raised to the fourth power are subtracted to yield smaller numbers. This situation is most pronounced when the signal and noise powers are approximately equal, resulting in large estimation errors. Simulation results support this observation. Therefore another estimator is sought that does not rely upon subtraction of fourth order statistics.

<u>Pre-differential detection statistics</u>

The optimum soft-symbol weight to be applied before differential detection of QPSK can be described as a function of time (k index) and OFDM subcarrier (n index). Similar to equation (3), this weight is

$$w_{k,n} = \frac{a_{k,n}}{\sqrt{2a_{k,n}^2\sigma_{k,n}^2 + \sigma_{k,n}^4}} \quad (7)$$

where $a_{k,n}$ is the fading coefficient of the kth symbol for the nth subcarrier, and $\sigma_{k,n}$ is the corresponding standard deviation of the noise or interference, both prior to differential detection. Notice that the weight of equation (7) is the square root of equation (3). This is a result of the reasonable assumption that the weight changes slowly over the symbol-pair time used in the differential detection. In effect, the differential detection squares the predetection weight of equation (7), which would result in equation (3). A method for improving the statistical estimates of equation (7) is sought.

<u>Practical CSI and Pre-Detection Weighting</u>

Practical methods for estimating CSI and weights using pre-differentially detected soft-symbols and weight also applied to the soft decision symbol prior to differential detection are explored here.

For moderate to high SNR, the weight of equation (7) could be conveniently approximated by

$$\lim_{SNR \to \infty} w_k = \frac{1}{2\sigma_k^2}, \quad (8)$$

where simple statistical measurements were used to estimate σ^2. However, simulation confirmed that this weight estimate performed poorly during times when the SNR was very low due to fading interference. For example, the optimum weight would have suppressed the noisy samples more than the high SNR approximation to the weight. Therefore, another approximation was sought which would estimate CSI statistics over a large SNR range. Furthermore the estimate should not be sensitive to a gaussian noise or interference assumption, and should be estimated with sufficient accuracy in a time (filter time constant) significantly less than the reciprocal of the fading bandwidth.

A simple and robust estimation technique evolved after simulation and some experimentation. This estimation technique approximates the previously-defined weight expressions, but uses lower-order statistical approximations and satisfies the compromise between statistical accuracy and agility to track the fading signal. This technique is described in the following 4 steps:

1. Create a sequence $v_{k,n}$ for each QPSK subcarrier consisting of the magnitudes of the complex soft decision outputs $s_{k,n}$ from the matched filter for the nth subcarrier.

$$\boxed{v_{k,n} = |s_{k,n}|} \quad (9)$$

2. Create a sequence $d_{k,n}$ consisting of the differences of successive time samples of $v_{k,n}$.

$$\boxed{d_{k,n} = v_{k,n} - v_{k-1,n}} \quad (10)$$

3. Filter the sequences $v_{k,n}$ and $d_{k,n}$ using second-order digital IIR filters, then compensate for any differences in effective group delay to yield sequences $filtv_{k,n}$ and $filtd_{k,n}$. The time constant for the $filtv_{k,n}$ filter should be somewhat smaller than the reciprocal of the fading bandwidth, while the time constant for the $filtd_{k,n}$ filter can be somewhat larger. These sequences are representative (approximately proportional) of the local mean and standard deviation of the sequence $v_{k,n}$.

4. The sequence of weights for the soft decisions for each subcarrier to be applied prior to differential detection is defined as

$$\boxed{w_{k,n} = \frac{1}{filtd_{k,n} \cdot \left(1 + \left(\dfrac{filtd_{k,n}}{filtv_{k,n} - filtd_{k,n}}\right)^4\right)}} \qquad (11)$$

To prevent numerical overflow, check to ensure that $filtv_{k,n} > 1.5 \cdot filtd_{k,n}$ in equation (11); otherwise, set the weight to zero. Simulation results verified that this weight yields good performance under a variety of channel impairments with fading and interference.

<u>Smoothing filters for statistical estimates</u>

The values of $filtd_{k,n}$ and $filtv_{k,n}$ are estimated using filtering techniques described next. Filtering is performed first for each subcarrier at the kth symbol instant in time. Then the rows of $filtd_{k,n}$ and $filtv_{k,n}$ are simply updated across the N subcarriers. Equation (12) filters the sequences $v_{k,n}$ with a time delay of approximately 16 symbols, and equation (13) filters the sequences $d_{k,n}$ with a time delay of approximately 64 symbols. Both filters have a zero frequency gain of nearly unity.

$$\boxed{subv_{k,n} = \frac{960 \cdot subv_{k-1,n} - 451 \cdot subv_{k-2,n} + 3 \cdot v_{k,n}}{512}} \qquad (12)$$

$$\boxed{subd_{k,n} = \frac{16128 \cdot subd_{k-1,n} - 7939 \cdot subd_{k-2,n} + 3 \cdot d_{k,n}}{8192}} \qquad (13)$$

Additional filtering is performed across the N subcarriers. Smoothing the estimates across the N subcarriers requires 3 passes of a simple IIR filter. The first pass sets the appropriate initial condition of the filter, but does not update the estimates. The direction of the second pass is reversed from the first, while the third pass is reversed again. This results in an approximately symmetric (linear phase) filter characteristic which is desirable for providing the estimates on the center carrier. Although it is impossible to provide this symmetric filtering for the subcarriers at each end of the band, the impulse response "tails" are folded back into the active subcarriers.

The first pass across the subcarriers sets the initial values of $filtv_{N-1}$ and $filtd_{N-1}$ without replacing the time-filtered values for each subcarrier. The time index k is ignored here since it is understood that the filtering over the subcarriers is performed over each kth OFDM symbol.

$$\boxed{\begin{array}{c} filtv_{N-1} \Leftarrow (1-\beta) \cdot filtv_{N-1} + \beta \cdot subv_n\ ; \\ filtd_{N-1} \Leftarrow (1-\beta) \cdot filtd_{N-1} + \beta \cdot subd_n \\ n = 0,1,\ldots N-1 \end{array}} \qquad (14)$$

The second pass smooths the values across the filtered estimates for each subcarrier, $subv$ and $subd$.

$$\boxed{\begin{array}{c} filtv_n \Leftarrow (1-\beta) \cdot filtv_{n+1} + \beta \cdot subv_n; \\ filtd_n \Leftarrow (1-\beta) \cdot filtd_{n+1} + \beta \cdot subd_n; \\ n = N-2, N-3, \ldots 0 \end{array}} \qquad (15)$$

The third pass smooths the frequency values again to achieve a nearly symmetrical impulse response (except for the subcarriers near the endpoints).

$$\boxed{\begin{array}{c} filtv_n \Leftarrow (1-\beta) \cdot filtv_{n-1} + \beta \cdot filtv_n; \\ filtd_n \Leftarrow (1-\beta) \cdot filtd_{n-1} + \beta \cdot filtd_n; \\ n = 1,2,\ldots N-1. \end{array}} \qquad (16)$$

The resulting filtered values for $filtv$ and $filtd$ are used in equation (11) at each OFDM symbol time to yield the appropriate weight for each soft symbol prior to differential detection, but after matched filtering, in the receiver.

III. CPC CODE FOR OFDM BROADCAST SYSTEM

A simple method of constructing the CPC code for this application is to start with an industry standard rate 1/3 convolutional code. This code can be generated as shown in Figure 3.

The rate 1/3 convolutional encoder of Figure 3 can be viewed as producing 3 encoded bit streams (G1, G2 and G3), each at the same rate as the input. The combination of these 3 bit streams produces the R=1/3 coded output sequence. To create a complementary code pair, for example, a subset of the output code bits is assigned to the lower DAB sideband and a different (complementary) subset is assigned to the upper sideband. Each subset must contain at least the same rate of bits as the information input rate, plus some additional bits to provide some coding gain.

The coded bit mask of a Puncture Pattern matrix is shown in Figure 4. The Puncture Pattern matrix of represents the encoder output symbols over each set of 4 information bits. Therefore the output symbols are identified and indexed modulo 4. A logic 1 in any of the 12 locations of the mask indicate that that particular bit is used; otherwise, a logic zero indicates that the bit is not used.

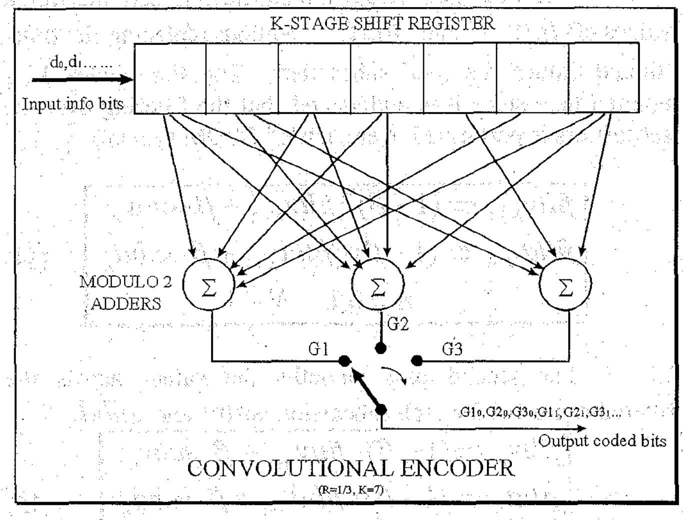

Figure 3. Example of a R=1/3, K=7 convolutional encoder.

$$\begin{bmatrix} G1_0 & G1_1 & G1_2 & G1_3 \\ G2_0 & G2_1 & G2_2 & G2_3 \\ G3_0 & G3_1 & G3_2 & G3_3 \end{bmatrix}$$

Figure 4. General Puncture Pattern Matrix.

It is relatively straightforward to select a good puncture pattern for transmission of the unpunctured bits on any one sideband [3,4,5]. This bit pattern may be chosen based on a known optimal R=4/5 puncture pattern, or from a RCPC code pattern. However, after one sideband is defined in this manner, there is little flexibility on choosing the bits for the opposite sideband since they should be chosen from the punctured (complementary) bits to achieve maximum coding gain when combined to form the base code. Fortunately, analysis and computer simulation have verified that good complementary codes do exist. For example the bit patterns shown in Figure 2 yield very good performance when G1=133, G2=171, and G3=165 where the generator connections are represented by standard octal notation.

A pair of complementary puncture patterns, one for the upper DAB sideband and one for the lower DAB sideband, are shown in Figure 5. Individually, these Puncture Patterns define the pair of rate 4/5 codes.

$$PPL = \begin{pmatrix} 0 & 1 & 1 & 0 \\ 1 & 0 & 0 & 1 \\ 0 & 0 & 1 & 0 \end{pmatrix} \quad PPU = \begin{pmatrix} 1 & 0 & 0 & 1 \\ 0 & 1 & 1 & 0 \\ 1 & 0 & 0 & 0 \end{pmatrix}$$

Figure 5. Puncture patterns for rate 4/5 CPC codes.

The pair of complementary rate 4/5 codes can be combined to form the rate 2/5 base code as shown in Figure 6. Each of the rate 4/5 codes has a free distance of d_f=4 with information error weight c_d=10. The combined rate 2/5 code yields d_f=11 with c_d=8. Notice that only half the G3 bits are used in this CPC code.

$$PPL + PPU = \begin{pmatrix} 1 & 1 & 1 & 1 \\ 1 & 1 & 1 & 1 \\ 1 & 0 & 1 & 0 \end{pmatrix}$$

Figure 6. Puncture pattern for original rate 2/5 code.

Optionally the punctured bits of the rate 2/5 code can be transmitted to yield a pair of rate 2/3 CPC codes with d_f=6 as shown in Figure 7. Of course, the base code is the unpunctured rate 1/3 code with d_f=14.

$$PPLop := \begin{pmatrix} 0 & 1 & 1 & 0 \\ 1 & 0 & 0 & 1 \\ 0 & 1 & 1 & 0 \end{pmatrix} \quad PPUop := \begin{pmatrix} 1 & 0 & 0 & 1 \\ 0 & 1 & 1 & 0 \\ 1 & 0 & 0 & 1 \end{pmatrix}$$

Figure 7. Puncture patterns for optional rate 2/3 CPC codes.

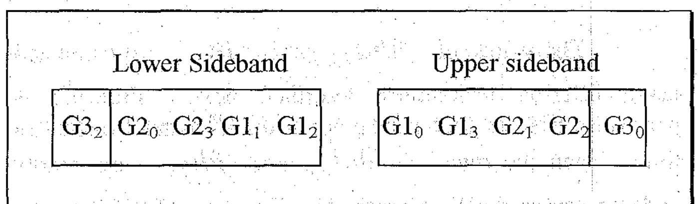

Figure 8. Example subcarrier assignment of CPC coded bits to lower and upper sidebands.

A rate 4/5 code on each sideband requires 25% additional bits. One method of allotting bits to the sidebands is presented in Figure 8. The Figure shows the relative spectral locations of the coded bits. These spectral locations are maintained after interleaving by channelizing the interleaver into distinct partitions which are mapped to the appropriate subcarriers on each sideband. The most expendable code bits are placed on the outer OFDM subcarriers. The expendable bits contribute least to the free distance or coding gain of the combined code. The optional G3 bits can be placed on the inner carriers closest to the host FM spectrum. Analysis and simulation have demonstrated that this partitioned interleaver outperforms random interleaving under typical interference scenarios.

IV. INTERLEAVER DESIGN FOR OFDM FM IBOC

Interleaving is performed both over the subcarriers (frequency) and time. Wesel and Cioffi addressed

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