SUBBAND-MULTIPULSE DIGITAL AUDIO BROADCASTING FOR MOBILE RECEIVERS

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Abstract

The audio quality, robustness and implementational complexity of a novel mobile digital audio broadcast scheme are addressed. The audio codec proposed is based on an efficient combination of subband coding (SBC) and multipulse excited linear predictive coding (MPLPC). The bit allocation is dynamically adapted according to both the signal power in different subbands and a perceptual hearing model. Typically a segmental signal to noise ratio (SEGSNR) in excess of 30 dB associated with high fidelity (HI-FI) subjective quality was achieved for 2.67 bits/sample transmissions at a bit rate of 86 kbits/s. Perceptually unimpaired audio quality was achieved for a bit error rate (BER) of about 10^{-4}, when injecting random errors, which was degraded for increased BERs. In order to provide robust error protection, the audio codec was also subjected to a rigorous bit sensitivity analysis.

Four different forward error correction (FEC) schemes were investigated in order to explore the complexity, bit rate and robustness trade-offs. The powerful but complex half-rate Reed-Solomon RS(380,190,95) code over GF(512) was studied in contrast to the similar rate low-complexity binary BCH(63,30,6) codec, and the performance of the lower complexity BCH codec was found superior. Furthermore, the ramifications of using the even lower complexity, lower bit rate, approximately 2/3-rate BCH(63,45,3) codec in contrast to a more complex, similar coding rate source sensitivity-matched twin-class FEC scheme were studied using the codecs BCH(63,51,2) and BCH(63,36,5) for the lower and higher sensitivity audio bits, respectively.

In order to maintain non-dispersive channel conditions and hence remove the need for power-hungry and complex channel equalisers, while maintaining high bandwidth efficiency, 4 bit/symbol 16level starconstellation quadrature amplitude modulation (16-StQAM) was deployed and the overall signalling rate became approximately 30 kBaud for the BCH(63,45,3) and the twin-class source-matched schemes. When transmitting via a twin-channel time division multiple access (TDMA) scheme the bandwidth requirement was less than 200 kHz, allowing four digital HI-FI channels to be accommodated in a conventional analogue FM channel's bandwidth. Our diversity assisted broadcast scheme required a channel signal to noise ratio (SNR) of about 25 dB for unimpaired audio quality via the worst-case Rayleigh fading mobile channel, when the mobile speed was 30 mph and the propagation frequency was 1.5 GHz. In case of the stationary Gaussian scenario an SNR of about 20 dB was required.

1 Introduction

Analogue frequency modulated (FM) radio broadcasting originates from 1949 and it was designed mainly for directional antennae with about 12 dB antenna gain. The audio quality of portable stereo FM radios has been under renewed criticism against the background of the proliferation of high fidelity (HI-FI) portable compact disc (CD), digital audio tape (DAT) and digital compact cassette players, leading to a growing demand for a terrestrial or satellite-based Hi-Fi digital audio broadcast (DAB) system.

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In 1986 the European Community (EC) launched the Eureka EU 147 project [1] with the aim of proposing a robust, mobile HI-FI DAB system, additionally capable of decoding concomitant traffic and control data, while listening to audio programs and synthesizing the corresponding voice messages on a demand basis at the required time. Furthermore, the system will be capable of producing five-channel surround sound with ambient-dependent dynamic control, catering for example for reduced dynamic range in a noisy vehicle, where pianissimo music cannot be heard. Originally two transform codecs (TC) and two subband (SB) codecs were proposed, and the best candidate codec [2], [3] reduces the 1.411 Mbit/s stereo CD bit rate to 192 kBit/s using the so-called 'masking pattern adapted universal subband integrated coding and multiplexing' (MUSICAM) technique.

The DAB system's transmission scheme is based on the coded orthogonal frequency division multiplex (COFDM) principle, whose origin goes back to the 1960s [4]-[6] and was considered recently for mobile radio telephony [7]. Its principle is that the total transmission bandwidth is divided in a high number of narrow-band sub-channels and each sub-channel is assigned a low-rate modem. Fortunately, the sub-channel modems do not have to be implemented separately, because it can be shown that the bank of sub-channel modems can be substituted by a pair of inverse fast Fourier transform (IFFT) and FFT processing, if the number of subchannels is an integer power of two. If the transmitted signal is corrupted by bursty channel errors, after FFT-demodulation at the receiver the errors will be distributed over the whole block, minimising the probability of erroneous decisions. The remaining randomly distributed errors can be more easily combatted by error correction coding. The interested reader is referred for further details on the DAB system to Reference [9].

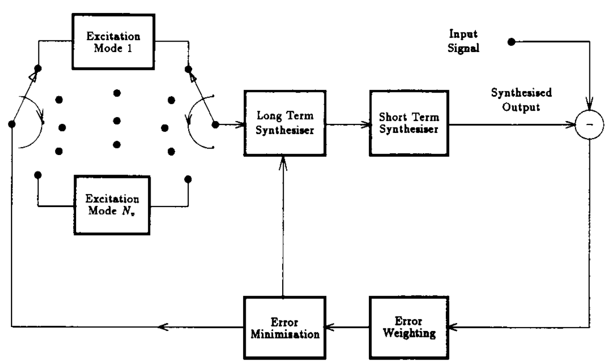

Figure 1: MMPLPC Codec Schematic

In our contribution we propose an alternative lower complexity DAB scheme for mobile channels, which is based on a specific subband split modified multipulse excited linear predictive (SB-MMPLPC) codec studied in Section 2. The audio bits are protected by a variety of block codes and transmitted using 4 bits/symbol 16-level quadrature amplitude modulation (16-QAM) described in Section 3. Issues of for-

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ward error correction coding (FEC) are addressed in Section 4, while the proposed transceiver scheme is discussed in Section 5. Finally the performance of the suggested broadcast scheme is characterised in Section 6, which is followed by our conclusions.

2 The Audio Codec

In recent years subband coding [11, 12] and transform coding [13, 14], multi-pulse LPC (MPLPC) [15] and wavelet transform [16], have been successfully used for HI-FI audio coding. In this paper, we propose a novel modified multi-pulse excited linear predictive codec (MMPLPC) structure combined with a subband splitting technique in order to further improve the coding efficiency of audio signals. The MMPLPC codec is developed from the amplitude re-optimization method of conventional multi-pulse LPC [17] schemes by choosing a number of different excitation modes which result in an improved subjective audio quality. By using split band coding, perceptually motivated dynamically adapted bit allocation procedures can be deployed in our subband MMPLPC (SB-MMPLPC) codec [18].

2.1 The MMPLPC Codec

The MMPLPC codec's schematic diagram is shown in Figure 1, which is similar to that of a conventional MPLPC arrangement, except that it incorporates N_v number of different excitation modes. The audio input signal $s(n)$ is divided into frames of N samples for LPC analysis and the LPC filter parameters are determined by minimising the mean squared prediction error E_w over this interval. Each frame is further divided into contiguous subframes of N_s samples, for which the long term predictor (LTP) parameters are initially determined under the assumption of no excitation, since at this stage the excitation is unknown.

In order to find the optimum LTP delay D and gain G_1 minimising E_w for the current subsegment, the term

$$T = \frac{\left(\sum_{n=0}^{N-1} x(n) y_D(n)\right)^2}{\sum_{n=0}^{N-1} y_D^2} \quad (1)$$

has to be maximised [10] over the range of delays D, where $x(n) = s_w(n) - \hat{s}_o(n)$ is the perceptually weighted audio signal after removing the memory contribution $\hat{s}_o(n)$ of the weighted synthesis filter $1/A(z/\gamma)$ due to its input in the previous subframe and $y_D(n) = \sum_{i=0}^{n} u(i-D)h(n-i)$ is the convolution of the previous history of $u(i)$ at delay D with the impulse response $h(n)$ of $1/A(z/\gamma)$. The factor γ controls the grade to which the error signal has to be perceptually de-emphasised in the spectrally prominent frequency regions during the excitation optimisation. The corresponding LTP gain G_1 is then given by

$$G_1 = \frac{\sum_{n=0}^{N-1} x(n) y_D(n)}{\sum_{n=0}^{N-1} y_D^2(n)}. \quad (2)$$

The optimum excitation is determined by filtering the candidate innovation sequences $v(n)$ through the LTP synthesis filter $1/P(z)$ and the perceptually weighted short term prediction (STP) synthesis filter $1/A(z/\gamma)$ in order to generate the perceptually weighted synthesized audio signal $\hat{s}_w(n)$. The total mean squared weighted error can be expressed as

$$E_w = (\mathbf{X} - \mathbf{HG})^{\mathbf{t}}(\mathbf{X} - \mathbf{HG}) \quad (3)$$

where $\mathbf{t}$ denotes the transpose operation and

$$\mathbf{X}^{\mathbf{t}} = (x(0), x(1), \cdots, x(N_s - 1)), \quad (4)$$

$$\mathbf{G}^{\mathbf{t}} = (g_0, g_1, \cdots, g_{N_g-1}), \quad (5)$$

while

$$\mathbf{H} = \begin{bmatrix} h(0-m_0) & h(0-m_1) & \cdots & h(0-m_{N_g-1}) \\ h(1-m_0) & h(1-m_1) & \cdots & h(1-m_{N_g-1}) \\ \vdots & \vdots & \ddots & \vdots \\ h(N_s-1-m_0) & h(N_s-1-m_1) & \cdots & h(N_s-1-m_{N_g-1}) \end{bmatrix} \quad (6)$$

Note that g_i represents the excitation pulse amplitudes, while m_i is the position of the pulse g_i in the excitation frame. By setting $\partial E_w/\partial \mathbf{G} = 0$ we arrive at

$$E_w = \mathbf{X}^{\mathbf{t}}\mathbf{X} - \mathbf{G}^{\mathbf{t}}\mathbf{H}^{\mathbf{t}}\mathbf{HG}, \quad (7)$$

which is minimised, when $\mathbf{G}^{\mathbf{t}}\mathbf{H}^{\mathbf{t}}\mathbf{HG}$ is maximized. If $\mathbf{G}$ is quantized to $\hat{\mathbf{G}}$, minimizing the weighted error becomes equivalent to maximizing

$$Q_p = 2(\mathbf{H}\hat{\mathbf{G}})^{\mathbf{t}}\mathbf{X} - (\mathbf{H}\hat{\mathbf{G}})^{\mathbf{t}}\mathbf{H}\hat{\mathbf{G}}. \quad (8)$$

When the total number of bits used to quantize the pulse amplitudes $\{g_i, i = 0, 1, \cdots, N_g - 1\}$ and to encode the pulse positions $\{m_i, i = 0, 1, \cdots, N_g - 1\}$ is fixed, the number of excitation pulses N_g can be varied to find the best set of quantised excitation pulses $\{\hat{g}_i, i = 0, 1, \cdots, N_g(k) - 1\}$, where the excitation mode index k in the range $[1, N_v]$ represents the specific excitation mode which maximizes Q_p above. Explicitly, the best compromise in terms of finding the number of excitation pulses and the associated number of quantisation bits must be found. If the number of excitation pulses is higher, only a lower quantisation precision is possible and vice versa. We refer to the afore mentioned method as MMPLPC.

2.2 The performance of MMPLPC

Using MMPLPC, simulations were performed with a variety of music signals listed in Table 1. The signals were band limited from 0.05 to 15 kHz and sampled at 32 kHz, and each signal had a duration of 6 to 16 seconds. Our results were compared with the conventional MPLPC. The simulation results for MPLPC were similar to those given in

music	contents
1	strings, trumpet, flute
2	trumpet with orchestra
3	saxophone, keyboards, drums
4	pop music(vocals, keyboards)
5	keyboards, guitar, drums
6	Pavarotti with cello
7	saxophone
8	soprano with orchestra
9	french horn, strings, cymbal

Table 1: Music Excerpts Used in Simulations

reference [15]. For our bandwidth of about 15 kHz and sampling frequency of 32 kHz we chose the parameters $N_s = 160$ corresponding to a subsegment length of 5 ms, $\gamma = 0.95$ representing a mild weighting, and a 10th order LPC filtering. At this stage no LTP filtering was invoked. The LPC analysis frame size of $N = 320$ samples was equivalent to a frame duration of 10 ms, and a Hamming window duration of 13.75 ms was used. Figure 2 shows the SEGSNR performance of two previously published codecs, MPLPC1 and MPLPC2 [17] in contrast to that of our proposed MMPLPC scheme. In MPLPC1 22 excitation pulse amplitudes $\{g_i, i = 0, 1, \cdots, 21\}$ were quantized with 7 bits/sample, while in MPLPC2 25 pulses $\{g_i, i = 0, 1, \cdots, 24\}$ were quantized using 6 bits/sample, as summarised in the bit allocation table, Table 2.2.

For our MMPLPC arrangement we set the number of excitation modes in Figure 1 to $N_v = 2$. Mode 1 and 2 used 22 and 25 excitation pulses, quantised using seven and six bits, respectively, as seen in MPLPC1 and MPLPC2. The full quantization schemes for the parameters of MPLPC1, MPLPC2 and MMPLPC are listed in Table 2.2. All the parameters were linearly quantized except the maximum excitation magnitude which was logarithmically quantized to eight bits precision and used in the normalization of g_i. The encoding of the positions m_i of the excitation pulses used the enumerative method, outlined in reference [19]. The simulation results of Figure 2 show that the segmental signal to noise ratio (SEGSNR) of the MMPLPC codec was almost always the highest when compared with the MPLPC1 and MPLPC2 schemes. Our informal listening tests confirmed also subjective improvements by the MMPLPC over MPLPC1

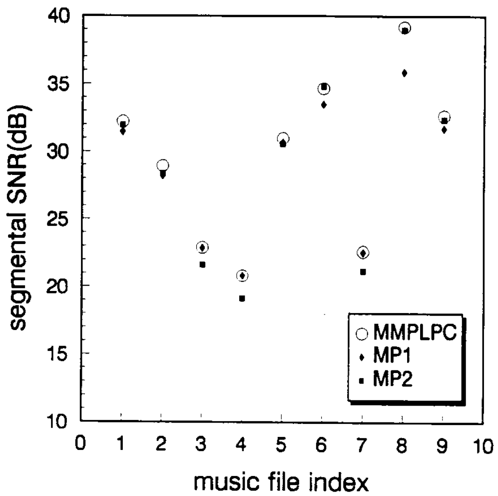

Figure 2: SEGSNR Performance of Various MPLPC Schemes

and MPLPC2. Figure 2 shows however that for example for music 6 the MMPLPC did not have the highest segmental SNR, because the minimization in the analysis-by-synthesis (ABS) loop was for the

	MPLPC1	MPLPC2	MMPLPC
mode(bit)			1
g_i(bit)	22 × 7	25 × 6	22 × 7, 25 × 6
m_i(bit)	65	69	65 , 69
normalization of excitation pulse (bit)	8	8	8 , 8
LPC in LAR_i(bit)	50	50	50
total bit rate (kbits/s)	95.8	95.8	96.2

Table 2: Bit Allocation Schemes for Codecs MPLPC1, MPLPC2 and MMPLPC

perceptually weighted error signal rather than for the original audio signal. Although this did result in a lower SEGSNR, the MMPLPC codec maintained a higher perceptual quality. Figure 2 demonstrates that for MPLPC1 and MPLPC2, some excerpts of music needed more excitation pulses per frame with less precise quantization, while some other sections needed more accurate quantization with fewer excitation pulses per frame. This property was the fact that led us to the concept of the MMPLPC.

2.3 Subband MMPLPC Structure

The audio codec's efficiency can be further improved, if the human ear's frequency and energy sensitive properties are exploited by dividing the audio bandwidth into subbands corresponding to the critical bands found in hearing [11, 20]. However, after band splitting, the correlation between adjacent time domain samples is reduced, and the more the band is split, the more this correlation is decreased. The MMPLPC codec utilizes linear prediction requiring high correlation between adjacent samples. In order to compromise, we chose four-band splitting.

The subband MMPLPC scheme is shown in Figure 3. The input audio signal $s_a(n)$ is split into four subbands: 0-4 kHz, 4-8 kHz, 8-12 kHz, 12-16 kHz, by a *Quadrature Mirror Filter* (QMF) bank, using two cascaded 64th order QMF filters [21]. The four subband signals $\{s_k(n), k = 1,2,3,4\}$ are each encoded by an MMPLPC codec. If we were to deploy pure waveform coding for the subband signals in the form of pulse code modulation (PCM) without taking account of perceptual hearing properties, the bit allocation would have to be

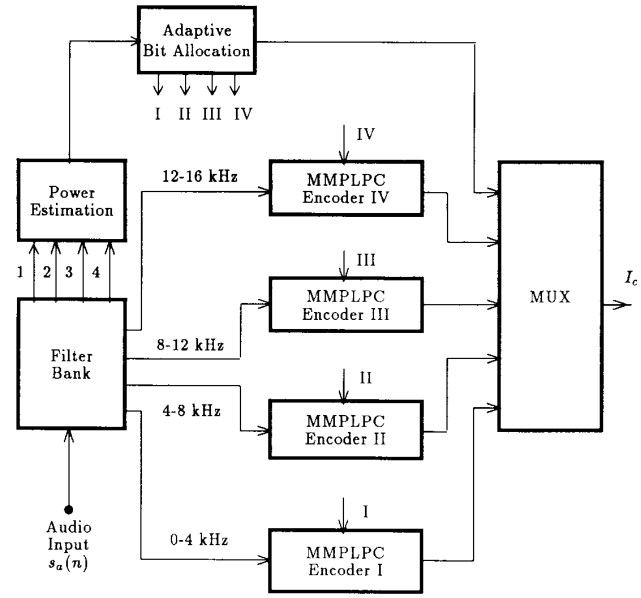

Figure 3: Subband Coded Multipulse Excited Schematic

adjusted according to the signal power σ_k^2 in each band k=1,2,3,4 using the formula [22]:

$$R_k = R + \frac{1}{2} log_2 \frac{\sigma_k^2}{[\prod_{i=0}^{N_L-1} \sigma_i^2]^{1/N_L}}, \tag{9}$$

where R is the total average number of bits/sample, and R_k is the number of bits/sample in band k. However, hearing sensitivity is different for the different subbands. For the same sound pressure, human hearing within 0.5-8 kHz is more sensitive than for frequencies higher than 8 kHz, and especially, for those higher than 12 kHz. Consequently, for the same subband input signal power $\sigma_k^2, k = 1,2,3,4$ more bits must be allocated to the more sensitive frequency bands $k = 1,2$ than to the less sensitive high frequency bands $k = 3,4$. Furthermore, since in the subband we propose to use perceptually motivated MMPLPC, Eq. 9 can only be used as an initial guide in our experimentally determined bit allocation scheme.

2.4 Subband-MMPLPC parameters

Accordingly, the short time energy σ_k^2 in each subband was estimated, then the proportion of bits allocated to band k was initially determined using Equation 9 and every subband k was assigned to one of sixteen empirically designed different bit allocation classes $C_j, j = 1\ldots16$, as demonstrated by Tables 3- 5. These tables summarise for both excitation modes the number of excitation pulses N_g, their quantisation accuracies in terms of the number of bits/pulse as well as the number of bits needed for the encoding of their positions, when using the enumerative method [19]. Observe from these tables that for the same bit allocation class $C_j(k)$ the lower frequency subbands $k = 1,2$ were typically assigned a higher number of excitation pulses and higher number of pulse amplitude quantisation bits, whose values were previously determined from a series of subjective experiments. The LPC analysis frame size of 20 ms was found to be suitable for every subband. As expected, the LPC prediction gain increased in each subband, when the LPC order was increased. To achieve high fidelity audio, much higher excitation densities were needed than for encoding speech.

The number of LPC filter coefficients was 6,4,4,4 for subbands 1,2,3,4 respectively. The LPC filter parameters were quantized by linear quantization of log-area ratios [10] $LAR_i(k)$. After band spliting to a bandwidth of 4 kHz, the sampling frequency was reduced to 8 kHz, yielding a subsegment length of 5 ms, equivalent to 40 samples. Accordingly, a subsegment excitation frame size of 5 ms or 40 samples was used for our SB-MMPLPC codec. Again, the excitation pulse positions were encoded using the enumerative method [19]. A long term predictor (LTP) was also invoked, as it provided a noticeable increment in subjective and objective quality when the excitation pulse density was low, and even for high excitation densities, it retained the same performance as without the LTP both in terms of bit rate and SEGSNR. For each subband, 4 bits were needed to linearly quantize the LTP filter gain, while 7 bits were used to encode the LTP delay.

When quantising the excitation pulse amplitudes g_i, we found that if the number of excitation pulses N_g was less than six, 3, 4, 5, or 6 bit quantization achieved almost the same quality with the segmental SNR differing by only 0.2 dB, while the number of excitation pulse quantisation bits doubled. If N_g was from six to ten, 4, 5, or 6 bit quantization had a similar effect, whereas if N_g was from eleven to sixteen, 5 or 6 bit quantization also got to within 0.2 dB and so on. So when we constructed the Tables 3-5, we used less precision to quantize the lower excitation density pulses, while higher precision was employed to quantize the higher excitation density pulses. The excitation pulse amplitudes were normalised by their maximum value within the subsegment and this maximum value was logarithmically quantised using eight bits for each subsegment and each subband before quantisation. The MMPLPC codec structure was identical for all four subbands. The total number of bits per 20 ms became 1707, which yielded a mono bit rate of about 86 kbits/s at a coding rate of 1707 bits/640 samples $\approx$ 2.67 bits/sample.

	mode 1			mode 2			
	N_g	bit/pulse	position	N_g	bit/pulse	position	total bit
1	2	2	10	2	2	10	14
2	6	3	22	7	2	25	41
3	9	3	29	12	2	29	58
4	13	4	34	17	3	37	89
5	18	4	37	24	3	37	110
6	17	5	37	22	4	37	126
7	20	5	38	26	4	35	140
8	22	5	37	28	4	33	148
9	27	5	37	34	4	34	170
10	23	6	37	28	5	33	176
11	26	6	35	33	5	25	192
12	27	7	34	34	6	22	227
13	33	7	25	33	7	25	256
14	37	7	14	37	7	14	273
15	39	7	6	39	7	6	279
16	40	7	0	40	7	0	284

Table 3: Bit Allocation Scheme for bands B1 and B2 of the SB-MMPLPC Codec

	mode 1			mode 2			
	N_g	bit/pulse	position	N_g	bit/pulse	position	total bit
1	2	2	10	2	2	10	14
2	4	3	17	5	2	20	31
3	6	3	22	7	2	25	41
4	7	4	25	8	3	27	54
5	7	4	25	9	3	29	57
6	11	4	32	14	3	35	78
7	14	4	35	18	3	37	92
8	17	4	37	23	3	37	107
9	17	5	37	21	4	37	123
10	19	5	37	24	4	36	133
11	22	5	37	29	4	32	149
12	24	5	36	34	4	22	159
13	24	6	36	30	5	30	181
14	26	6	35	34	5	22	192
15	29	6	32	40	5	0	207
16	28	7	33	35	6	20	231

Table 5: Bit Allocation Scheme for band B4 of the SB-MMPLPC Codec

	mode 1			mode 2			
	N_g	bit/pulse	position	N_g	bit/pulse	position	total bit
1	2	2	10	2	2	10	14
2	4	3	17	5	2	20	31
3	7	4	25	8	3	27	54
4	9	4	29	11	3	32	66
5	14	4	35	18	3	37	92
6	18	4	37	24	3	36	110
7	18	5	37	22	4	37	128
8	22	5	37	28	4	33	148
9	23	5	37	30	4	30	153
10	27	5	34	34	4	22	170
11	23	6	37	28	5	33	176
12	26	6	35	33	5	25	192
13	27	7	34	34	6	22	227
14	33	7	25	33	7	25	256
15	37	7	14	37	7	14	273
16	39	7	6	39	7	6	279

Table 4: Bit Allocation Scheme for band B3 of the SB-MMPLPC Codec

2.5 Subband-MMPLPC performance

In a further experiment the objective and subjective audio quality of our 86 kbits/s SB-MMPLPC codec was compared to that of the previously proposed 96 kbits/s full-band MMPLPC codec and a 128 kbits/s wide-band subband split adaptive differential pulse code modulated (SB-ADPCM) codec. Our SB-ADPCM benchmark codec was based on the CCITT G.722 Recommendation [24], but due to the doubled sampling rate of 32 kHz its bit rate was also doubled.

Our informal listening tests demonstrated that the SB-MMPLPC codec operating at an overall bit rate of 86 kbits/s outperformed both of the higher-rate benchmarkers in terms of subjective audio quality,

Parameter	Total of SB1-SB4	SB1	SB2	SB3	SB4
Lar_i Bit-position	103	35 1-35	24 36-59	24 60-83	20 84-103
$C_j(k)$ Bit-position	13	4 104-107	4 108-111	4 112-115	1 116
$Mode_i$ Bit-position	4	1 117	1 118	1 119	1 120
LTP delay Bit-position	28	7 121-127	7 128-134	7 135-141	7 142-148
LTP gain Bit-position	16	4 149-152	4 153-156	4 157-160	4 161-164
$\max(g_i)$ Pulse-position of g_i g_i Redundant	A B C D				

Table 6: SB-MMPLPC bit allocation for the first subframe

although its objective SEGSNR performance was slightly lower. This fact is attributable to the error weighting filter, which de-emphasized the error signal in the perceptually less audible frequency regions.

2.6 Bit-Sensitivity Analysis

In order to ensure robust source-matched error protection for our favoured 2.67 bits/sample SB-MMPLPC audio codec it was subjected to bit sensitivity investigations by systematically corrupting all of its bits in a 1707 bit frame and evaluating the SEGSNR penalty inflicted. When for example the sensitivity of bit 1 was investigated, this bit was consistently corrupted in every frame, while keeping all other bits intact. The 1707-bit frame is constituted by 103 bits for the LPC parameters and 401 bits for every 5 ms subframe. The detailed bit allocation within a frame is shown in Table 6, where A, B, C, and D represent quantities having a variable number of quantisation bits in the subbands SB1-SB4 that add up to a fixed value of $A+B+C+D = 340$ bits.

In order to show the objective importance of the different subbands, as an example in Figure 4 we evaluated their SEGSNR using music excerpt 3. Observe that subband 1 has an average SEGSNR of nearly 25 dB, subband 2 an average of about 15 dB, while subbands 3 and 4 have fairly low SEGSNR, yet they improve the subjective quality. The formal subjective investigation of the bit error sensitivities would be desirable, but for 1707 bits constitutes a time consuming exercize, hence we had to satisfy ourselves with less reliable objective assessments. Accordingly, the subband energy and the associated SEGSNR values predictably give a lower weight to high-band coding bits than to low-band bits.

An overview of the SEGSNR degradation inflicted by systematic bit corruption is given in Figure 5 (a)-(f) for music excerpt 3 for the first 103 LAR bits and the subsequent 401 bits representing the first 5 ms sub-frame. The results for the remaining sub-frames are identical

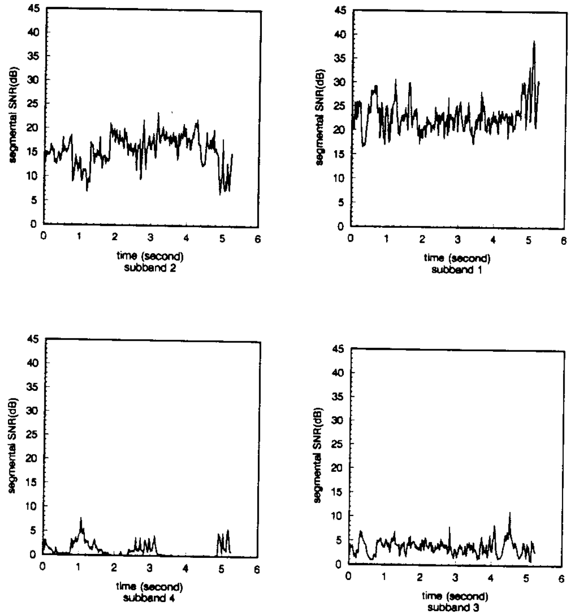

Figure 4: SEGSNR versus time for bands B1-B4

to those of the first one. Observe in the global Figure 5(a) that the degradation caused by the first 250 bits, and in particular by the first 50 low-band LAR coefficients bits are the most dramatic, which can be more clearly seen in the expanded Figure 5(b). This figure also reflects that the LAR sensitivity in subbands 3 and 4 is not dramatic.

The results of Figure 5(c) are also interesting to analyse. For example, according to Table 6 bit positions 106-110 that encode the high-energy low-band bit allocation classifiers $C_j(k)$ are vulnerable, but those of the lower band are less sensitive. These bits are followed by the important high-energy band (SB1, SB2) excitation mode bits 117-118 and the less vital low-energy mode bits 119-120. On the same note, the high-energy LTP delay bits 121-127 are followed by more robust lower energy band LTP bits approaching bit position 148. Again, the more vital low-band LTP gain bits cause a deep SEGSNR curve cut above this position, but the curve improves for the higher bands. The sub-frame maxima $\max(g_i)$ follow from position 165 onwards, with the low-band ones yielding a deep SEGSNR valley, which after a temporary recovery for the high-bands dips again to about 10 dB, indicating the location of high-band excitation pulses. After this last sensitive region the SEGSNR curve exhibits substantial robustness for the remaining bits. The overall shape of these bit sensitivity curves suggests that basically there are two sensitivity classes, the sensitive C1 and the more robust C2 categories, associated with more than 15 dB and less than 10 dB SEGSNR degradations, respectively.

The robustness of the proposed codec evaluated in terms of segmental SNR degradation was also evaluated injecting random errors assuming a given fixed BER, as would be experienced over a Gaussian channel. On the basis of the previously discussed bit sensitivities initially we divided the bits in six sensitivity classes, subjected each class to random bit errors and evaluated the SEGSNR degradation as a function of the BER for all six classes, although these results are

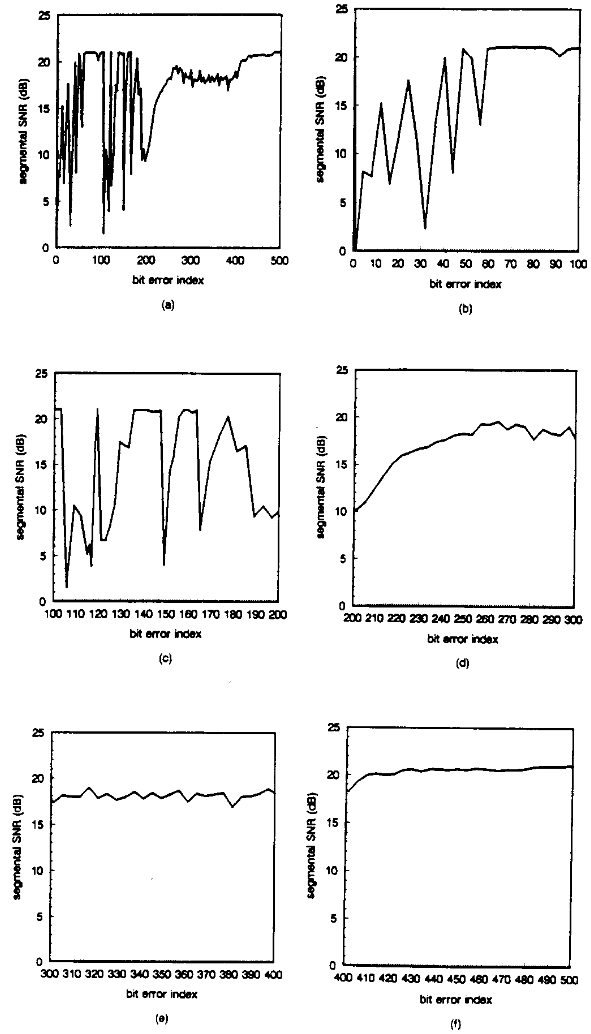

Figure 5: Bit Error Sensitivity of the 2.67 bits/sample Audio Codec in Terms of SEGSNR

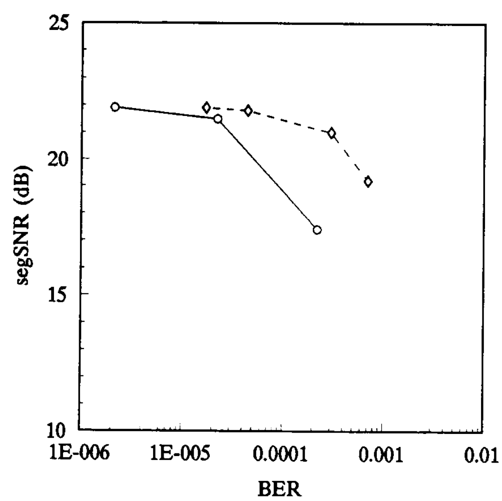

Figure 6: SEGSNR Degradation versus BER for two sensitivity classes

not shown in this treatise due to lack of space. As expected on the basis of the bit sensitivity curves shown for the systematic corruption of bits in Figure 5, fundamentally only two different sensitivity classes exist. The SEGSNR degradation of these two classes due to random, rather than consistently periodic bit errors introduced with various fixed error probabilities is shown in Figure 6, where it becomes clear that the BER of the more sensitive C1 bits must be below about 10^{-4}, while that of the more robust C2 bits below about 10^{-3}, in order to ensure acceptable audio quality, although even lower BERs are preferable.

Having designed the audio codec, we now embark upon considering the pertinent audio transmission issues via fading mobile channels.

3 Non-Coherent 16-QAM

The well-known square shaped 16-QAM phasor constellation is optimum for transmissions over Gaussian channels due to its maximal noise protection distance of its constellation points [25]. With Rayleigh fading channels its performance is impaired by occasional false locking of its carrier phase.

For fading channels the differentially coded Star QAM (StQAM) scheme proposed in reference [26] has a better performance than square QAM in spite of the error-doubling property of its differential coding. Indeed, when using non-coherent demodulation, the penalty in terms of channel SNR is lower than the expected 3 dB. Therefore we opted for this 16-StQAM scheme, where the first bit of a four-bit symbol is differentially encoded onto the twin-rings of the phasor constellation, while the remaining three bits are differentially Gray-coded onto the eight phase positions. The two rings are at normalised radii of 1 and 3 units. At the receiver the ratio of the previous and present phasor magnitudes is determined and if less than two units and greater than half a unit, a logical zero is generated, otherwise a logical one is inferred for the first bit of a four-bit symbol. The remaining three bits are differentially Gray-decoded from the phase change of the previous symbol.

Instead of utilising tolerance-sensitive linear-phase Nyquist filters, the low-complexity concept of non-linear filtering (NLF) joining time-domain signal transitions with a quarter of a period duration raised-cosine (RC) segment can be invoked [33]. In our scheme the time-domain RC-segment fitting was followed by a low-pass filter ensuring a 50% excess bandwidth above the signalling rate.

In our previous discourse we have assumed that the mobile channel exhibits flat fading, having a considerably higher coherence bandwidth than the signalling rate. If this condition no longer applies in spite of the fact that the 16-QAM signalling rate is a quarter of the transmission bit rate, our channel equaliser proposed in reference [27] must be invoked. This considerably increases the DAB transceiver's complexity. Alternatively, the similarly complex COFDM scheme [9] can be used in order to combat dispersion, although our simulation model of the 16-StQAM based COFDM arrangement had slightly inferiror performance over narrowband channels, when compared to our conventional 16-StQAM system.

Having considered the modem scheme used for our DAB transceiver we now focus our attention on the error correction arrangements used.

4 Error Correction Coding

Block codes are powerful against both random and bursty channel errors and have reliable error detection capability, which is a useful feature in many applications. Hence in our DAB transceiver designed for bursty fading mobile channels we favoured block codes [10], [32].

Over fading mobile channels the channel interleaver has to disperse burst errors concentrated around deep fades so that the FEC codec can have approximately equal number of errors in each block. Naturally, long Reed-Solomon (RS) block codes inherently possess this randomising property, but are complex to implement. In our deliberations we will represent this class of codes using the powerful half-rate RS(380,190,95) code over the finite Galois field GF(512), which encodes 190 nine-bit symbols into 380 symbols and can correct 95 symbol errors. This code represents the highest practically acceptable implementational complexity for our system and when encoding the 86 kbits/s audio information generated by the SB-MMPLPC encoder we have a bit rate of about 171 kbits/s. This yields a signalling rate of 42.75 kBd that requires a bandwidth of about 64 kHz, when using an excess bandwidth of 50 %.

A very good compromise in terms of implementational complexity and error correcting power is constituted by the family of binary BCH codes of 63 bits length. Often used members of this family are the BCH(63,30,6), BCH(63,36,5), BCH(63,39,4), BCH(63,45,3), BCH(63,51,2) and BCH(63, 57,1) codes, correcting 6, 5, 4, 3, 2 and 1 bits per frame, respectively. The BCH(63,30,6) code has a coding rate of $R = 30/63 \approx 0.5$ and will be used in our experiments as a low-complexity alternative to the complex but similar rate RS(380,190,95) code. The associated bit rate for our DAB transceiver becomes 179.55 kbits/s, yielding a signalling rate of 44.9 kBd and necessitating a bandwidth of 67 kHz.

A third FEC scheme that we will investigate is constituted by a twin-class embedded source-matched un-equal protection arrangement. Our proposed audio codec has a framelength of 20 ms, which is encoded using 1707 bits. Accordingly, these bits are mapped into two sensitivity classes, C1 and C2 based on our findings in Section 2.6 and matched BCH codes are assigned to them. The number of bits in the sensitivity classes C1 and C2 are 684 and 1020, respectively, yielding a total of 1704 bits, while three bits are unprotected. The FEC codes assigned are the five-error-correcting BCH5 = BCH(63,36,5) and double-error-correcting BCH2 = BCH(63,51,2) codes, respectively. In C1 there are nineteen, while in C2 twenty BCH codewords, respectively, and the total number of FEC-coded bits is $2457 + 3 = 2460$.

In the audio transmission frames bits 1-103 represent the LAR parameters, while each of the four subsegments is encoded by 401 bits, yielding a total of $103 + (4 \cdot 401) = 1707$ bits. The LAR bits 1-103 are assigned to C1, along with the most important 145 bits of each subsegment, yielding a total of $103 + (4 \cdot 145) = 684$ C1 bits. The remaining $4 \cdot 255 = 1020$ C2 bits are assigned to the weaker BCH2 codec. The overall coding rate becomes $R = 1707/2460 \approx 0.69$, while the transmission rate is 123 kbits/s, giving a Baud rate of about 30.7 kBd. The required bandwidth is $1.5 \cdot 30.7 = 46$ kHz.

The performance of this scheme will be gauged against that of a similar rate, but less complex single-class code, namely the BCH(63,-45,3) code, for which $R = 45/63 = 0.71$, the total number of bits per frame is 2394, the bit rate becomes 119.7 kbits/s and the signalling rate is ≈ 30 kBd. The bandwidth requirement of this scheme is about 45 kHz.

In order to randomise the bursty channel errors, for each FEC scheme we use rectangular interleaving over one audio frame corresponding to 20 ms, before modulation takes place. This measure curtails error propagation across frame boundaries. As an example, let us relate this interleaving memory to a vehicular speed of 30 mph or 13.3 m/s giving a travelling distance of 26.6 cm/20 ms. For propagation frequencies of 1.5 GHz, as in the proposed DAB system, the wavelength is about 20 cm, and therefore interleaving over an interval of 26.6 cm/20 ms ensures adequate randomisation. However, for slowly walking pedestrians there is a danger of idling in deep fades, which is detrimental as regards to reception quality. In this situation a switcl iversity or frequency hopping scheme are essential.

5 The Proposed Transceiver Scheme

The schematic diagram of our audio broadcast transceiver is depicted in Figure 7. The 'FEC encoder/decoder scheme' drawn in dashed lines represents the four previously mentioned coding schemes. If the RS(380,190,95) scheme is used, no mapping and interleaving is needed, since the entire audio frame is encoded by a single $9 \cdot 190 = 1710$-bit codeword, requiring only 3 padding bits. In case of the BCH(63,30,6) code $1710/30 = 57$ codewords constitute an audio frame, allowing for an interleaving depth of 57 words, but again, no source mapping is re-

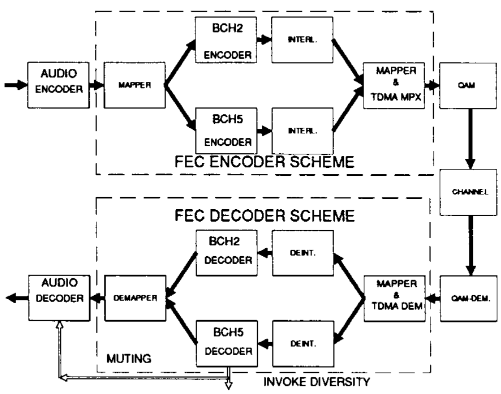

Figure 7: Broadcasting System Schematic

quired. When using the BCH(63,45,3) code, 1710/45 = 38 codewords encode an audio frame, hence the interleaving depth is 38.

Lastly, when the twin-class source-matched FEC scheme is used, the audio encoder's bits are sorted by the 'Mapper' into two sensitivity classes according to their vulnerability, as described in Section 2.6. Twin-class unequal error protection is deployed using the previously proposed 63-bit binary two- and five-error correcting BCH codes denoted by BCH2 and BCH5, respectively. The FEC coded bits are rectangularly interleaved over the current 20 ms audio frame, namely over 20 and 19 BCH5 and BCH2 codewords, respectively. The interleaved bits are mapped back in one stream by another bit 'Mapper', multiplexed by the Time Division Multiple Access (TDMA) multiplexer (MPX) with three additional mono or another stereo audio program, as it will be explained in the next paragraph, 16-StQAM modulated onto a 1.5 GHz carrier and transmitted via the Rayleigh fading broadcast channel.

The half-rate FEC-coded schemes have a stereo bandwidth requirement of about $2 \cdot 67 = 134$ khz. Single channel per carrier (SCPC) transmissions with a stereo bandwidth of about 134 kHz are convenient in terms of low transmission bitrate, which is well below the coherence bandwidth of the typical mobile broadcast channel, hence ensures that the channel is essentially a narrowband flat fading transmission medium. Hence no 'power-hungry' channel equaliser is required, which is crucial in order to maintain low receiver complexity, low battery drain and hence ultimately light-weight construction. However, at low signalling rates the time between two adjacent signalling symbols is long and hence the channel's fading envelope changes dramatically between two adjacent transmitted samples and due to this

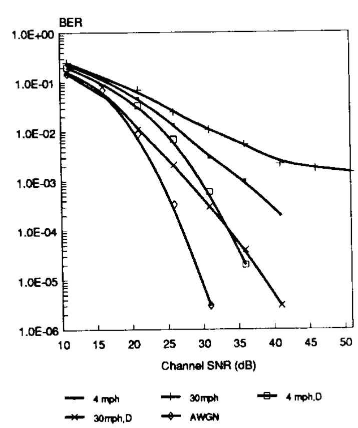

Figure 8: BER versus channel SNR performance of various 16-StQAM modems

the 16-StQAM modem's BER performance suffers from its lack of ability to efficiently trace the received signal's amplitude and phase trajectory. This is particularly true, if the mobile receiver's speed is high.

Although the proposed second order switched diversity receiver efficiently mitigates this problem, it is advantageous in terms of improving the 16-StQAM BER performance to transmit several channels per carrier in a TDMA structure. This increases the transmission signalling rate and hence improves the receiver's BER performance for slowly walking pedestrians. However, the transmission bit rate must not exceed the typical propagation channel's coherence bandwidth in order to avoid using equalisers. A TDMA structure based on two stereo channels constitues a good compromise in this respect.

A further advantage of the TDMA structure is that the receiver can monitor the reception quality of other transmitters during the unused time slot and allow seamless switching between two transmitters broadcasting the same program. of the approximately $R = 2/3$-rate schemes becomes about $4 \cdot 30$ kBd=120 kBd, which requires a bandwidth of $1.5 \cdot 120 = 180$ kHz. This rate allows us to fit two stereo digital audio channels into one conventional analogue frequency modulated (FM) channel while ensuring that the transmission bandwidth is narrow with respect to the fading channel's coherence bandwidth. The $R \approx 0.5$-rate schemes have an approximate signalling rate of 45 kBd, yielding a four-channel TDMA rate of about 180 kBd, which can accommodate a twin-channel stereo scheme slotted into a 270 kHz channel.

The receiver carries out the inverse functions of the transmitter. Observe that the error detection capability of the stronger C1 FEC decoder is exploited to invoke muting, if the FEC decoder happens to be frequently overloaded. Furthermore, this error detection capability can also be utilised to monitor the channel's BER statistics and control automatic hand-over to another transmitter broadcasting the same program but providing a higher signal-to-noise ratio.

6 Results and Discussion

The mobile satellite channel has been characterised in a number of excellent treatises [28]-[31] and for our 120 kBd signalling rate at a wave propagation frequency of 1.5 GHz it can be considered non-dispersive. The fading is typically Rician, and the best and worst case Rician channels are the Gaussian and Rayleigh channels, respectively. Therefore we limit the variety of propagation scenarios to these two cases, noting that for other Rician channels the system performance will be between these extreme cases.

Linear amplification was assumed, no AGC and carrier recovery were invoked, and our previously described non-linear raised-cosine pulse-shaping with the concatenated 50 % excess-bandwidth LPF was utilised to contain the spectrum within a band of 1.5 times the Nyquist frequency. The BER versus channel SNR performance of our 16-StQAM modem is depicted with and without diversity for mobile receiver speeds of 30 mph and 4 mph in Figure 8. Observe that at 30 mph, which is a typical urban vehicular speed, the modem's differential codec was not able to adequately trace the fading envelope and phase trajectory in the steepest fading sections, which resulted in a residual BER of about 10^{-3}.

The 4 mph BER vs. SNR performance curve, characteristic of a typical pedestrian scenario, is seen to be considerably more favourable in terms of BER than the 30 mph performance. The lower BER is attributed to the differential coding deployed, which effectively traces the fading envelope at this slow speed. When second order switched diversity (D) was invoked, the BER performance further improved both at 30 mph and 4 mph. However, in the low SNR region at 4 mph the diversity scheme was not able to improve the performance as significantly, as at 30 mph. This is because at low speeds, i.e., low fading rates both diversity channels received more correlated signals, and hence both received signals were simultaneously of high or low received signal power. Nonetheless, at both speeds near-Gaussian BER performance was achieved, when diversity was deployed.

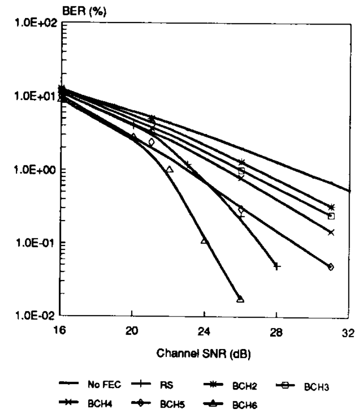

Figure 9: BER versus channel SNR performance of the proposed 16-StQAM/TDMA transceiver at 30 mph without diversity using various FEC codes

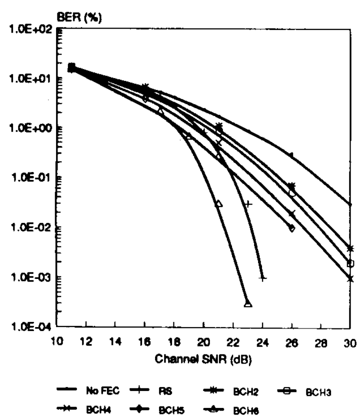

Figure 10: BER versus channel SNR performance of the proposed 16-StQAM/TDMA transceiver at 30 mph with diversity using various FEC codes

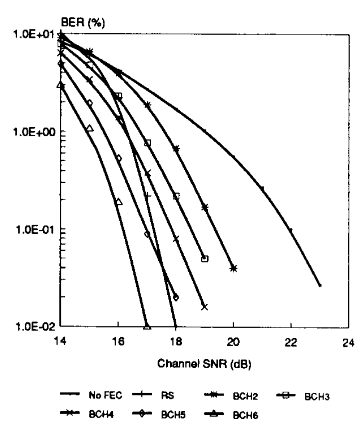

Figure 11: BER versus channel SNR performance of the proposed 16-StQAM/TDMA transceiver via AWGN channels using various FEC codes

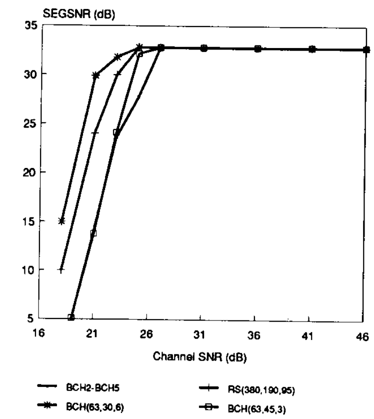

Figure 12: SEGSNR versus channel SNR performance of the proposed 16-StQAM/TDMA transceiver at 30 mph with diversity using various FEC codes

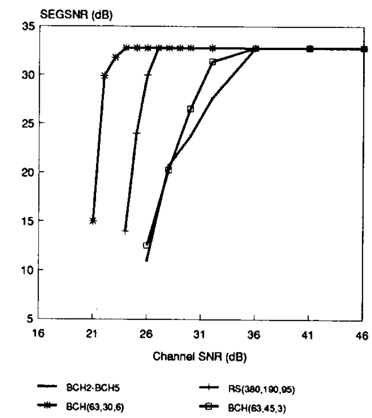

Figure 13: SEGSNR versus channel SNR performance of the proposed 16-StQAM/TDMA transceiver at 30 mph without diversity using various FEC codes

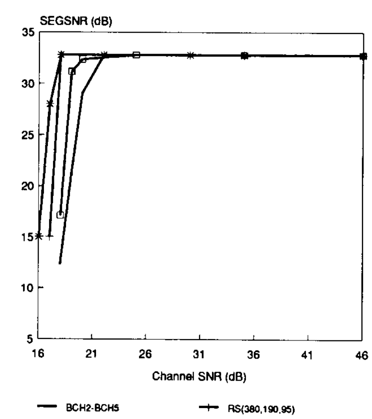

Figure 14: SEGSNR versus channel SNR performance of the proposed 16-StQAM/TDMA transceiver via AWGN channels using various FEC codes

In Figures 9 and 10 we portray the BER performance of the proposed 16-StQAM modem for a signalling rate of 180 kBd assuming four-channel mono TDMA transmissions both without and with diversity (D) at a mobile speed of 30 mph, and propagation frequency of 1.5 GHz using the set of BCH2-BCH6 codes and the RS(380,190,95) code. Observe that these performance curves are also characteristic of an SCPC system at four times lower vehicular speed, i.e at about 7.5 mph. Note the gradual BER improvements, when using stronger codes. It is also interesting to observe that the most complex RS codec is outperformed by the lower complexity BCH5 and BCH6 codecs, if the SNR is above about 22-24 dB both with and without diversity. This is due to the fact that the 9 bits/symbol RS codec's symbol error rate becomes very high, since a single bit error is sufficient to corrupt a symbol, a fact extremely inconvenient for the RS code operating on a symbol bases. Similar tendencies are also true for the AWGN channel, depicted in Figure 11, which is characteristic of a stationary scenario.

The overall objective SEGSNR versus channel SNR performance of our proposed broadcast scheme is portrayed for music excerpt 6 with and without diversity at a mobile speed of 30 mph, as well as for the stationary AWGN scenario in Figures 12, 13 and 14, respectively. As anticipated from the BER versus channel SNR curves seen in Figures 9-11, best SEGSNR performance is guaranteed by the single-class protected half-rate type systems, namely the BCH6 and the RS coded arrangements. Again, the modest-complexity BCH(63,30,6)-coded scheme outperforms the more complex RS(380,190,95) arrangement, although its SNR advantage becomes less dramatic via the preferred diversity-assisted scenario. Both of these diversity-assisted higher bit rate schemes require an SNR of about 24 dB at 30 mph for unimpaired audio quality, while this 'corner' SNR value is about 17 dB for the stationary AWGN channel. For SNRs below these values the audio quality rapidly degrades and the output signal has to be muted gradually to zero, unless the same audio program is broadcast in the other time-slot, which was previously monitored but not decoded.

In case of the less robust and lower bit rate schemes using the BCH3 codec or the twin-class BCH2-BCH5 arrangement the receiver faced substantial difficulties in removing the errors, when no diversity was used. However, in the diversity-supported scenario only about one dB excess channel SNR was required in order to ensure unimpaired audio reception over the 30 mph Rayleigh-fading channel, when compared to the half-rate type systems. For transmissions over the more benign stationary AWGN channel the minimum required SNR was about 20 dB, some 2 dB higher than for the half-rate coded arrangements. The source-matched twin-class BCH2-BCH5 coding scheme failed to outperform the less complex single-class BCH3 codec due to the inherent error sensitivity of the SB-MMPLPC codec, because in case of bad channel conditions the weaker BCH2 code was more frequently overloaded than the BCH3 code. This resulted in a SEGSNR penalty, in spite of the more robust protection of the C1 bits.

These tendencies were also confirmed by our informal subjective listening tests, although the source-matched BCH2-BCH5 twin-class scheme typically had a slightly better perceived quality than the BCH3-protected arrangement. This was due to the more robust protection of the most sensitive C1 bits. In subjective terms the audio quality was perceived as unimpaired, as long as the SEGSNR was in excess of about 30 dB for our test music excerpts, which was ensured for different channel SNRs for the various FEC schemes and channel models with and without diversity. Based on these experiences in our final proposed DAB scheme we favour the lowest complexity BCH3 codec.

7 Summary and Conclusions

A modified multipulse LPC audio codec was introduced and incorporated in an SB-MMPLPC codec for the encoding of wideband audio signals. At a coding rate of about 2.67 bit/sample, high fidelity audio reproduction was achieved for a mono bit rate of 86 kbits/s. In order to maintain a low bandwidth occupancy and low complexity, non-coherent differential 16-QAM signalling was favoured over the Rayleigh-fading mobile channel. In the majority of propagation scenarios the performance difference between the most complex RS FEC code and the least complex BCH(63,45,3) code is relatively low, hence preference is granted to the latter.

Our proposed transceiver scheme facilitates unimpaired digital HI-FI audio transmissions over fading mobile satellite broadcast channels in a mono bandwidth of about 45 kHz or equivalent stereo bandwidth of 90 kHz for channel SNR values in excess of about 25 dB and for both pedestrian as well as typical vehicular speeds. The system performance can be further improved at the cost of higher implementational complexity and battery consumption, when using a more sophisticated pilot symbol assisted, block-coded coherent square 16-QAM modem [33]. Future work will be targeted at improving the audio quality, complexity, bit rate, bandwidth occupancy and error resilience trade-off achieved.

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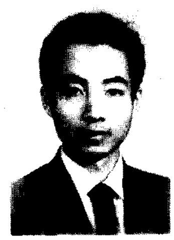

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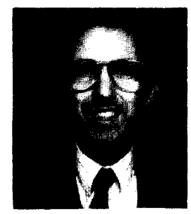

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Raymond Steele is the author of the book Delta Modulation Systems (Halstead, New York; 1975), editor of the book Mobile Radio Communications (Pentech Press, London, 1992) and the editor of a series of books on Digital Mobile Communications. Raymond Steele and his co-authors were awarded the Marconi Premium in 1979 and in 1989, and the Bell System Technical Journal's Best Mathematics, Communications, Techniques, Computing and Software and Social Sciences Paper in 1981. He has been a conference and session organiser of numerous international conferences and a keynote speaker at many international meetings. He is also a senior technical editor of the IEEE Communications Magazine. Raymond Steele is a Fellow of The Royal Academy of Engineering and a Fellow of the IEE in the UK, as well as, a senior member of the IEEE.

William Webb has gained a Ph.D. in the field of QAM techniques from the University of Southampton, UK, while working for Multiple Access Communications Ltd., UK, where he was promoted to the position of technical director. He has published over twenty papers on QAM for which he has been awarded the IERE Premium, and holds three patents on QAM. He currently works as a mobile radio consultant for Smith System Engineering Ltd., UK.