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THE DESIGN AND DEVELOPMENT OF A SCPC DIGITAL RADIO PROGRAM RECEIVI SOURCE CODING AND MODULATION TECHNIQUES DOCUMENTATION

MCS MILLER COMMUNICATIONS SYSTEMS LTD.

THE DESIGN AND DEVELOPMENT OF A SCPC DIGITAL RADIO PROGRAM RECEIVER SOURCE CODING AND MODULATION TECHNIQUES DOCUMENTATION

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TABLE OF CONTENTS

1.0 INTRODUCTION

This report consists of source coding, multiplexing, modulation and channel coding documentation intended to lead up to a preliminary hardware specification. Wherever relevant, several potential techniques were analysed and their relative merits compared. Both qualitative and quantitative results are presented to substantiate the comparison, and to the extent possible, justification is provided for the techniques finally selected. However, more work needs to be done to optimize certain parameters which may be important from the implementation point of view.

The source coding techniques including the implementation aspects are discussed in detail in Section 2. The multiplexing techniques are discussed in Section 3. In Section 4, the modulation techniques are discussed. Quantitative simulation results which predict the modem performance are also presented. In Section 5, the performance and implementation aspects of a number of potential channel coding techniques are discussed.

A number of PCM derived source coding techniques which employ instantaneous companding (ICOM) as well as near-instantaneous companding (NICOM) were compared. The ICOM techniques appear to have an edge over the NICOM techniques. The PCM technique employing 14 to 11 bit instantaneous A-law companding is a CCITT Standard. The codec implementation is relatively simple and straightforward. In the existing NICOM implementation, different error correction techniques are suggested for bits of different importance. Therefore the use of highly powerful FEC codes may be difficult. In ICOM, except for the 1 bit parity protection, the error correction (FEC) is done independently of the source codec. This enables one to use commercially available FEC codecs to achieve a large coding gain. Larger coding gains are important as the transponder is typically power limited and not bandwidth limited. For this reason, the slightly higher bit rate compression achievable in NICOM as compared to ICOM has less practical significance. While the multiplexed NICOM channels can easily interface with the (European) 2.048 Mbps hierarchy, ICOM channels easily interface with the (North American) 1.544 Mbps hierarchy. Further digital conversion of the 14 - 11 bit A-law compressed (A=43.8) data stream to a 15 - 11 bit μ-law compressed (μ=255) data stream (some U.S. networks already use this technique) and the reverse process can be hardware implemented relatively easily. These advantages coupled with a North American market target potential favour heavily on 14 - 11 ICOM-PCM technique. The detailed implementation and performance specifications of this technique are discussed in this report.

In the case of data multiplexing, the choice of continuous multiplexing as opposed to packet multiplexing is rather obvious. The advantages of continuous multiplexing are:

- requires only a small proportion of overhead bits for framing and/or signalling.

- operational flexibility

- for a rigid (fixed data rate) structure, the receiver complexity is minimized.

The channel multiplexer and demultiplexer will have design features suitable to accommodate different program (15 kHz or 7.5 kHz) and data channels. Additionally, T1 link mux and T1 link demux can be used in conjunction with the channel mux-demux for providing a terrestrial link between the studio center and the earth station.

The choice of modem technique is relatively complex and is influenced by several important factors;

- the E_b/N_o requirement for a given BER,

- implementation complexity and cost,

- immunity to interference (especially that due to intermodulation),

- sensitivity to synchronization errors.

The use of existing transponders for the digital SCPC operation with typical receivers (with 3M or 4.5M antenna and 120°K LNA) results in power limitations. The power efficiency is a critical factor in regard to maximizing the number of channels per transponder. The transponder nonlinearity will cause

- spectral spreading
- intermodulation
- signal suppression
- modulation transfer

While the choice of constant envelope, continuous phase modulation schemes may be logical from the spectral spreading point of view, the power efficiencies of such schemes (duobinary MSK, TFM, etc.) are relatively poor when used in conjunction with simple detection techniques.

Further, in order to minimize intermodulation, the transponder may be operated with several decibels backoff and therefore spectral spreading may not be significant. Intermodulation interference and weak signal suppression can be minimized by proper allocation of frequency and level to the SCPC carriers.

In view of the outstanding requirements on the power efficiency, differential detection schemes have not been given full considereation as they generally perform inferior to the coherent detection schemes. Among the coherent schemes, the potential schemes considered were binary PSK (BPSK), quartenary PSK (QPSK), offset-QPSK (OQPSK) and minimum shift keying (MSK). Based on several factors such as (1) the probability of error performance in the linear and nonlinear channels, (2) the implementation complexity and cost, (3) the immunity to interference including intermodulation, and (4) the sensitivity to synchronization errors, QPSK appears to be a favourable choice.

In order to reduce the transmission bit errors a number of forward error correction (FEC) coding schemes were studied. This includes a number of block and convolutional coding techniques. Several hard and soft decision decoding schemes including threshold, Viterbi, sequential, and Berlekamp-Massey-Chien (BMC) algorithms were studied. The objective code rate can be 1/2, 3/4 or 7/8. A larger coding gain is particularly important as this gain can be used to increase the number of carriers in a transponder and/or reduce the receiver antenna size. In general a higher coding gain target can be met at the cost of higher bandwidth redundancy and a higher implementation complexity and cost. It is necessary that the coding gain be >2.5 dB. A coding gain in the range of 4 to 5 dB is highly desirable. Commercial codecs which can achieve this coding gain can be purchased at reasonable cost.

2.0 TASK 3: SOURCE CODING

The selection of a source coding scheme for the digital encoding of high quality stereo audio is a principal consideration in the design of a satellite program distribution system. In this report, the various factors which impact the choice of a digital audio encoding scheme are documented.

The intent of this phase of the study is to select a PCM-based encoding scheme, having taken into account all relevant considerations. Although audio performance is an important factor in this choice, the overriding consideration is that the system be compatible with the CCIR/CCITT specifications for the transmission of digital audio in North America.

Section 2.1 of this report discusses the various fidelity criteria that determine the audio performance of a digital codec. Section 2.2 then considers the pre-digitization processing of the audio signal at the encoder and the corresponding analog processing following the D/A converter at the decoder. In Section 2.3, the properties and performances of linear, nonlinear and segmented quantization techniques and instantaneous and nearly-instantaneous companding schemes are discussed. Section 2.4 then considers the performance of the source codec in the presence of transmission errors and suggests methods of improving it, and Section 2.5 considers the problems associated with transcoding and cascading. Section 2.6 then discusses practical codec implementation considerations such as system compatibility, equipment availability, system configuration restrictions, and internetwork interfacing requirements. Section 2.7 concludes by summarizing the important recommendations of the report.

2.1 Fidelity Criteria

In an audio distribution network, the fidelity of the received audio signal depends on the processing of the input analog signal prior to digitization, the manner in which the signal is digitized, and the error performance achieved over the transmission medium. The implementation of the system with respect to each of these factors will be discussed in the following three sections. However, before the codec parameters can be determined, a knowledge of the required audio fidelity criteria to be met by the system is necessary.

In this section, a discussion of the major parameters which impact audio performance is first presented. Such factors as audio bandwidth, sampling rate, signal-to-noise ratio, program-modulated noise and idle channel noise are considered. This is followed by a brief discussion and summary of the additional audio specifications recommended by the CCIR/CCITT.

2.1.1 Bandlimiting and Sampling

In order to select the sampling frequency to be used in a digital system, the bandwidth required for high quality transmission of stereo audio programs must be determined.

The issue of what is an acceptable limit on the high frequency components of audio programs is largely subjective in nature. The conclusions of subjective listening tests performed by several authors [1]-[3] can be summarized as follows:

(i) The rise in audibility threshold at 15 kHz with respect to 1 kHz is only about 20 dB, but at 20 kHz it is on the order of 70 dB.

(ii) Although an isolated high-frequency tone may remain audible, it will be perceived very indistinctly when a second tone, a narrow-band noise component, or a natural sound is added (this is demonstrated in Figure 2.1).

(iii) Tests conducted using equal amplitude harmonics, critical program signals, and white noise did not show a perceptible reduction in quality due to bandlimiting the source to as low as 15 kHz.

These observations, coupled with the desire to minimize the transmission rate required for distribution, led to a choice of 15 kHz as the bandwidth for high quality representation of audio programs [4]. Accordingly, a sampling rate of 32 kHz was selected to accommodate the non-ideal nature of the analog filters necessary for the bandlimiting operation.

This transmission standard of 32 kHz compares with the production studio sampling rate of 48 kHz and the standard for digital encoding and playback (i.e. compact discs) of 44.1 kHz. Note, however, that the transmission rate was selected to be integrally related to the production studio standard. This should facilitate any direct digital transcoding that may be necessary.

2.1.2 Signal-to-Quantization Noise Ratio

The signal-to-quantizing noise ratio (SN_QR) is perhaps the most (objective) informative measure of source codec performance. Quantization, or granular, noise results from the finite step size used in the quantization process and is a measure of the quantizer's ability to resolve low-

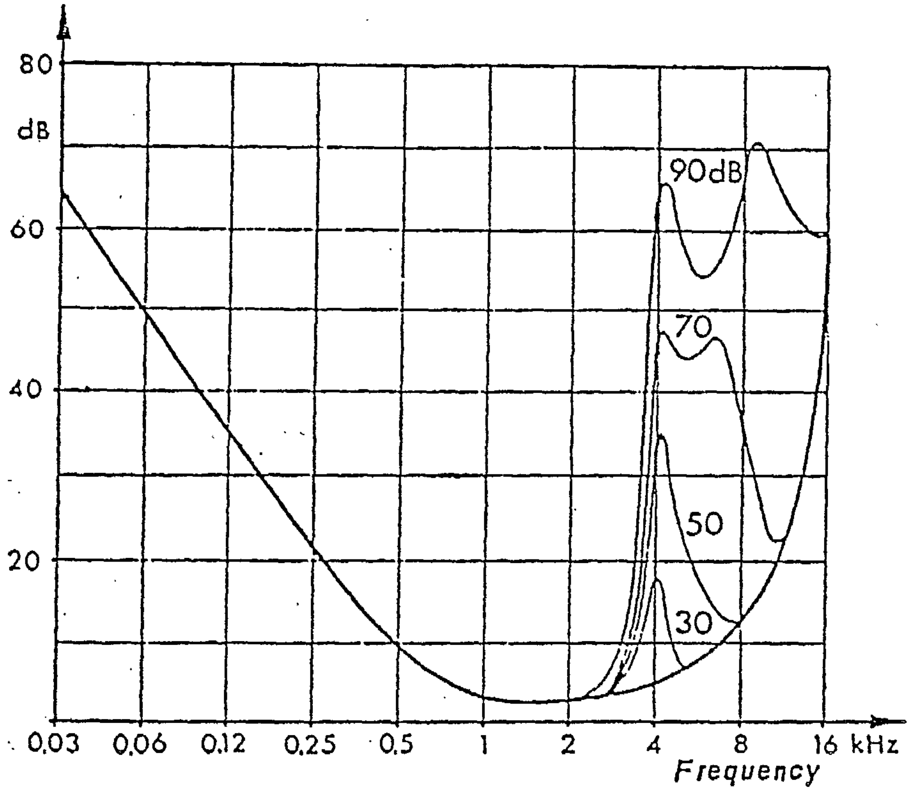


Figure 2.1 - Perceptibility threshold in the presence of an interfering 4-kHz tone of variable level (as a function of the level) [1]

amplitude samples. For linear PCM, it is generally perceived as a relatively unannoying hiss, similar to the effect of additive white Gaussian noise. In this case, SN_QR varies directly with the input signal level, since the quantization step size remains constant over the range of the quantizer (i.e. the quantization noise is effectively masked by higher level signals). It has been determined that a minimum resolution of 14 bits is required in order to reduce the quantizing noise to an acceptable level.

In nonlinear quantization (i.e. companded systems), the step size increases with the signal level*. Although the concept behind this approach is to make the SN_QR constant over a wide range of signal levels, this property leads to a phenomenon known as program-modulated noise. This effect is highly subjective since the perceived effect of the quantizing noise varies with the program material. In general, as the level of the input signal increases, the quantization noise increases and becomes predominantly high frequency in nature due to the larger quantizing errors. Hence, this type of noise is most annoying on high-amplitude, low-frequency signal components. High-frequency components tend to mask the noise somewhat.

2.1.3 Signal-to-Clipping Noise Ratio

The signal-to-clipping noise ratio (SN_CR) is a measure of the overload distortion that occurs when the input signal exceeds the limits of the quantizer range. This phenomenon, which manifests itself in the form of clipping of the input waveform, generally has a significantly more annoying effect than quantization noise. In fact, if noise due to this effect is much greater than the granular noise,

*The implementation of this type of system will be discussed in Section 2.3.

the effective SNR is virtually independent of quantizer step size and is dominated by overload distortion.

Subjectively, the codec overload level and quantization step size have been found to be the primary determinant of listeners' evaluation of the system, while bandwidth is a relatively secondary measure of quality. Hence, it is imperative that the signal levels throughout the system be defined to avoid an undue amount of overload distortion.

2.1.4 Audio Specifications

A considerable amount of effort in the area of digital encoding of audio programs has been carried out in recent years by organizations such as Scientific Atlanta and Tau-tron in the United States, the BBC in Great Britain, and Bayly Engineering in Canada. In parallel with this work, the CCIR and CCITT have been updating their performance specifications to cover the transmission of 15 kHz digital audio programs.

It is not the intent of this project to research the various parameters which affect digital audio quality from the ground up, but rather to take advantage of the work which has already been done in this area. The most important aspect of this project is that a satellite distribution system be developed in which the operation of the source codec complies with the specifications and standards set by the CCIR and CCITT in North America, such that compatibility with existing systems will be a possibility. This position is strengthened by the fact that CBC (who represent one of major potential users of the proposed system) have stated that they will be interested in a system that conforms to the North American standard.

In order to define the required performance of the source codec, the CCIR/CCITT specifications governing the basic audio parameters are provided in Table 2.1. These specifications are largely taken from CCIR Report 647-2 [5], which deals with the performance of a single codec. The corresponding CCITT recommendation J.21 [6] considers the overall performance of three such codecs connected in tandem at audio frequencies. More information concerning these specifications is provided in the AT&T Technical Advisory No. 74 [7], on which the specifications adopted for use in Canada were based.

2.2 PRE- AND POST-PROCESSING OF ANALOG SIGNALS

Before the audio signal can be converted to a digital format by the source encoder, there are several steps taken in the processing of the analog input signal. Similarly, following the D/A converter at the source decoder, some analog processing is required in order to restore the reconstructed signal to its original analog form. This section considers the various analog processing operations at both the encoder and decoder, as illustrated in Figure 2.2

2.2.1 Pre- and De-Emphasis

Pre- and de-emphasis of the high-frequency signal components is a technique commonly used in FM transmission to reduce the effects of high-frequency noise. However, research has shown that this signal processing technique is applicable to the digital encoding of audio as well [8]-[10].

The use of pre-emphasis at the encoder and the corresponding de-emphasis at the decoder has two principle benefits:

Audio Bandwidth	0.04 to 15 kHz
Gain vs. Frequency Response	0.04 to 0.125 kHz: +0.25 to -1.0 dB 0.125 to 10 kHz : +0.25 to -0.25 dB 10 to 14 kHz : +0.25 to -1.0 dB 14 to 15 kHz : +0.25 to -1.4 dB
Group Delay Distortion	0.04 kHz : < 18 ms 0.075 kHz: < 8 ms 14 kHz : < 3 ms 15 kHz : < 4 ms (Between these points the specification varies linearly in a linear-delay vs. logarithmic-frequency diagram)
Full Load Level	+21 dBm
Total Harmonic Distortion	0.04 to 0.125 kHz: < 0.4% 0.125 to 15 kHz : < 0.3% for -15 to +18 dBm sinusoid input
Intermodulation Distortion	< 0.3% for simultaneous sinusoid inputs of 500 Hz and 2 kHz at +12 dBm
Signal to Quantizing Noise Ratio	> 52 dB for 1004 Hz sinusoid input at level in range -6 to +18 dBm
Idle Circuit Noise	< -76 dB referenced to full load level (i.e. < -55 dBm)
Maximum Gain Difference Between Channels of a Stereo Pair	0.04 to 0.125 kHz: < 0.4 dB 0.125 to 10 kHz : < 0.3 dB 10 to 15 kHz : < 0.8 dB
Maximum Phase Difference Between Channels of a Stereo Pair	0.04 to 0.125 kHz: < 6 degrees 0.125 to 10 kHz : < 3 degrees 10 to 15 kHz : < 6 degrees
Crosstalk Attenuation	> 75 dB over 40 Hz to 15 kHz

Table 2.1 - Digital Audio Specifications

(i) it improves the program-modulated SNR, and

(ii) it reduces the effect of "clicks" caused by bit errors.

As was mentioned in Section 2.1.2, the program-modulated noise increases as the input signal increases and becomes predominantly high-frequency in nature. If pre-emphasis is incorporated at the encoder, then the de-emphasis at the decoder tends to reduce these high-frequency noise components. Hence, the program-modulated noise is reduced because signal excursions to the levels where quantization noise is high will occur only for above-average levels of higher-frequency components (these will be infrequent and will tend to mask the noise). Subjective tests carried out in [8] indicated that for all program material tested the use of pre- and de-emphasis reduced the audibility of the program-modulated noise to the extent that, on average, one less bit would be required than in the case with no pre- and de-emphasis.

The perceived effect of transmission errors in a PCM system is similar to that caused by random noise with a relatively intense high-frequency component [9]. However, as the bit error rate increases, the sound due to bit errors gradually changes from a continuous noise to sporadic clicks. The use of high-frequency de-emphasis at the decoder tends to diminish the severity of these clicks. The overall effect of this operation is to effectively shift the significance of the bit in which an error was caused downwards by two positions.

In general, the motivation for using pre- and de-emphasis on digital audio circuits from a technical standpoint is well established. A recent CBC report [10] on the

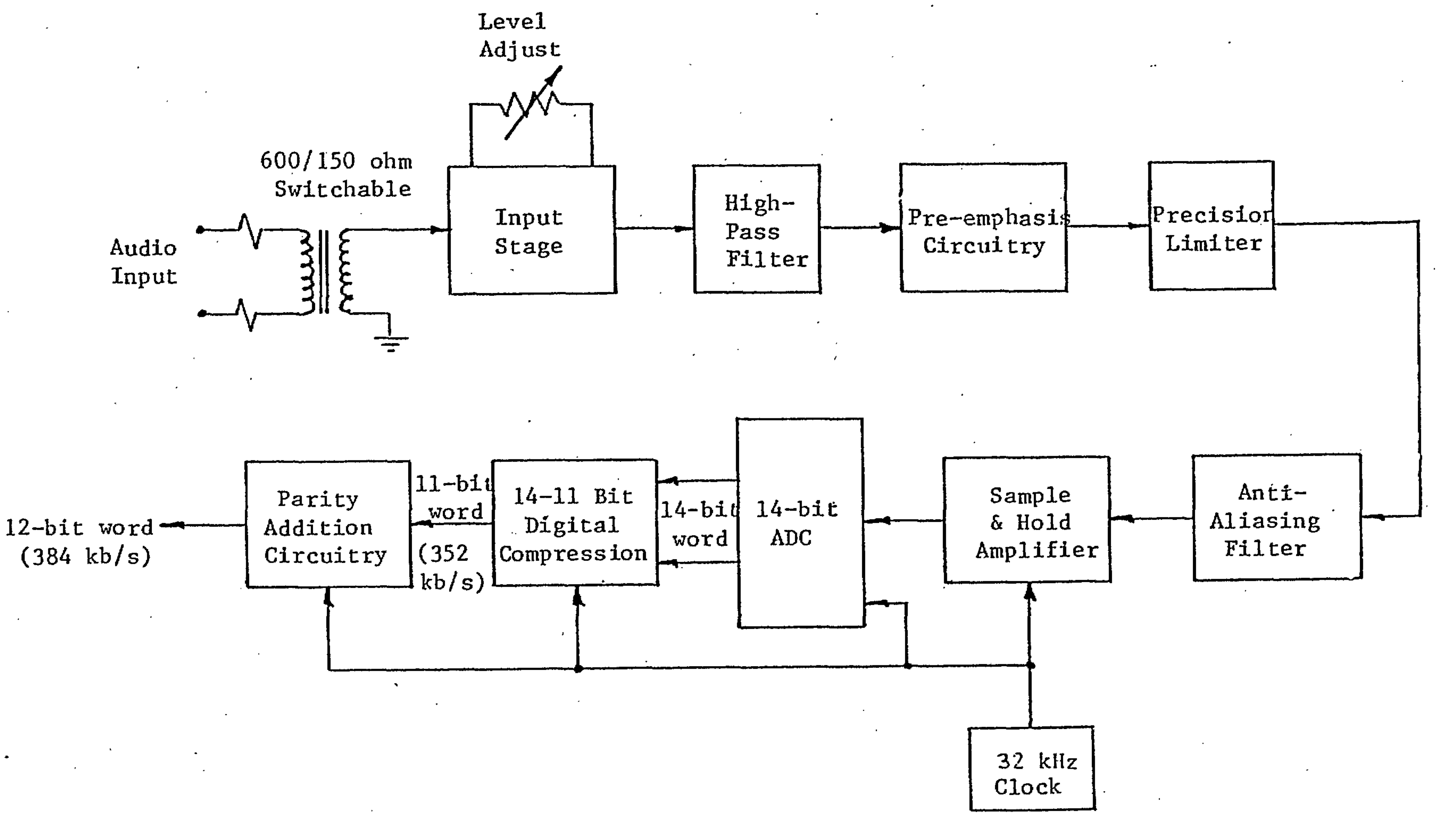


Figure 2.2 (a) - Source Encoder Block Diagram

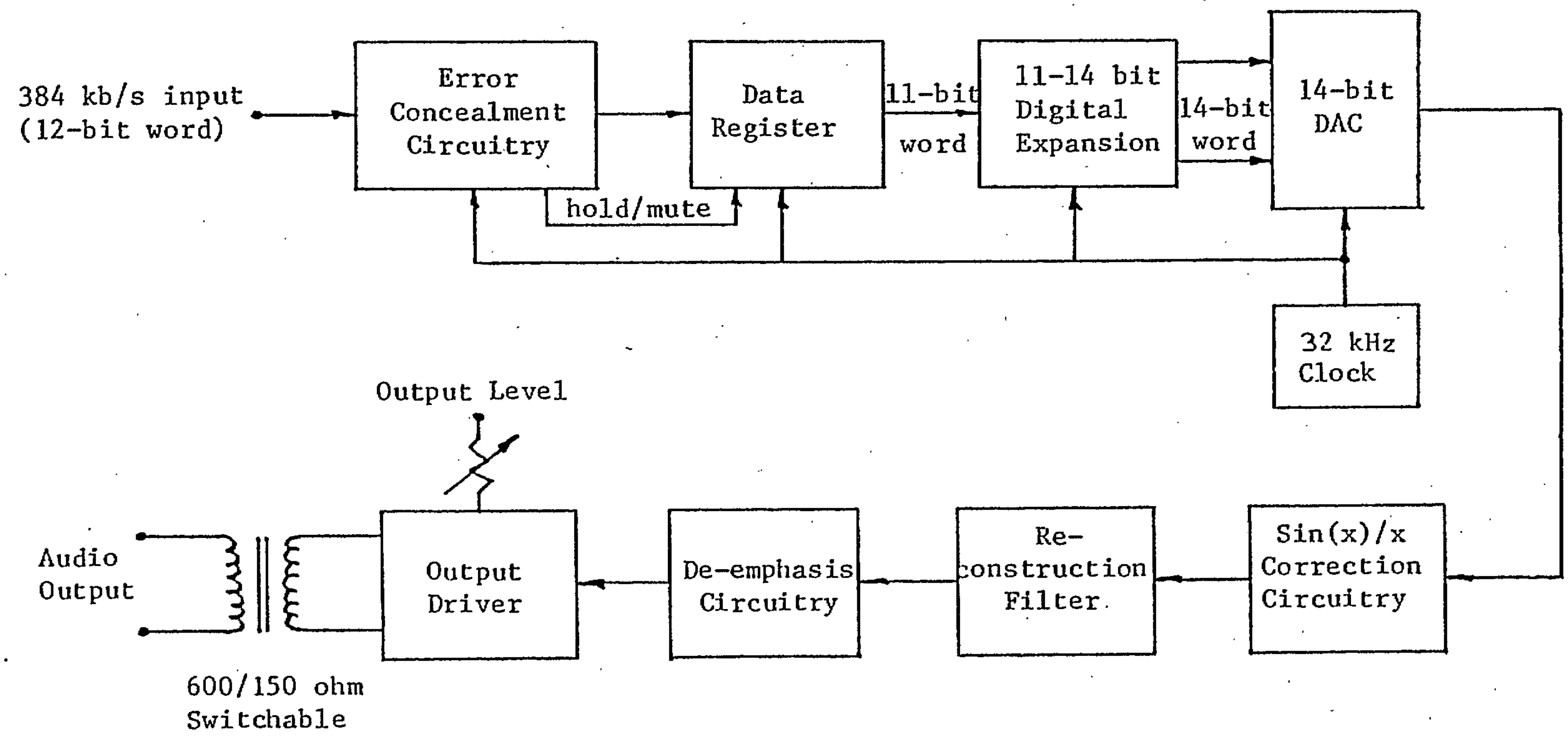


Figure 2.2 (b) - Source Decoder Block Diagram

subjective evaluation of a digital audio codec concluded that the use of pre- and de-emphasis provided consistently improved subjective performance, particularly in the case of multiple codecs operating in tandem. Despite this evidence, the CCITT does not currently recommend the use of pre- and de-emphasis for digital audio (primarily for compatibility reasons). However, it is anticipated that it will be recommended for use on a national basis within Canada, with a concerted effort being carried out to persuade the U.S. networks to incorporate this feature as well [11].

If pre-emphasis is used, the circuit will conform to that described in CCITT Recommendation J.17 [12]. The normalized pre-emphasis curve (such that the high-frequency gain is zero) is given by:

$$\text{Attenuation} = 10 \log \frac{75 + (\omega/3000)^2}{1 + (\omega/3000)^2} \text{ dB}$$

where $\omega = 2\pi f$ is the angular frequency in rad/sec.

In practice, the overall gain of the circuit should be adjusted so that the tradeoff between the probability of overload distortion and the low-level signal to quantizing noise ratio is optimized. In CCITT Recommendation J.41, the circuit gain is defined such that 6.5 dB of attenuation is provided at 800 Hz. This pre-emphasis curve, which has a zero-gain point at 2.12 kHz and 6.6 dB gain at high frequencies, is shown in Figure 2.3. The corresponding de-emphasis curve will have the inverse characteristic.

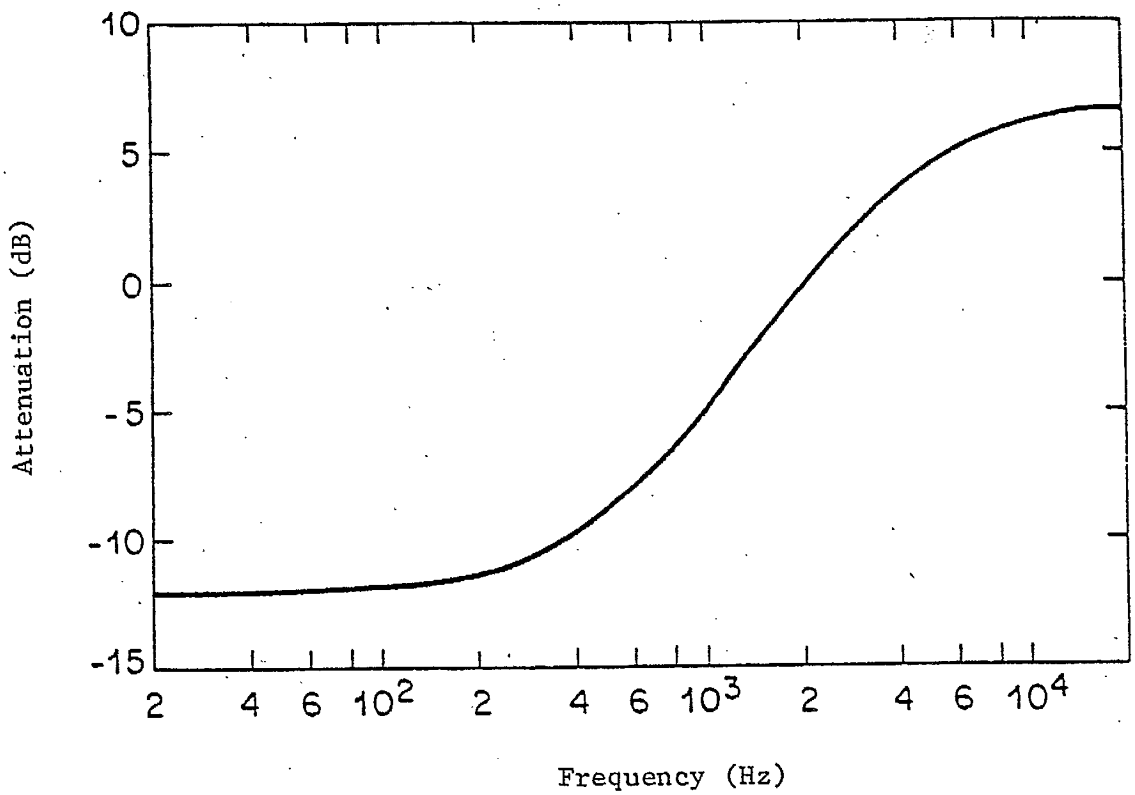


Figure 2.3 - CCITT Pre-Emphasis Characteristic [12]

2.2.2 Analog Limiting

In order to prevent the A/D converter from being driven into saturation when the input signal is overdriven, it is recommended that a precision analog limiter be used following the pre-emphasis network. This limiter should operate very infrequently due to the fact that the pre-emphasis network actually reduces the mean signal level, thus decreasing the probability of overload in general. Any harmonic distortion products introduced by the process of limiting the analog signal (which will be primarily low-order in nature) must be within the recommended distortion specifications.

2.2.3 Filtering/Equalization

As is indicated in Figure 2.2, there are several stages of filtering and equalization of the analog signals throughout the source encoder and decoder.

2.2.3.1 High-Pass Filter

The first analog processing stage of the encoder consists of a high-pass filter intended to limit infrasonic signals such as those associated with turntable rumble, tone arm resonance, and record warps. The filter should be designed such that the low-frequency components of the audio signal (i.e. down to 40 Hz) are passed unaffected, but that more than 25 dB of attenuation is provided at 4 Hz.

2.2.3.2 Anti-Aliasing Filter

In the digital encoding of analog signals, it is extremely important that the input signal be bandlimited to one-half the sampling frequency (32 kHz in this case). This is to ensure that any higher frequency components, which are aliased back into the audio spectrum, are attenuated such that they are below the A/D converter noise floor. Hence, the requirements of the anti-aliasing filter are fairly stringent, since it must be designed to pass signals up to 15 kHz undistorted while providing a minimum attenuation of 65 dB for components above 16 kHz.

2.2.3.3 Sin(x)/x Equalizer

At the decoder, it is necessary to compensate for the sin(x)/x frequency response factor associated with the sampled nature of the analog output from the D/A converter. The equalization should be such that the equalizer output has a flat frequency response over the audio spectrum.

2.2.3.4 Reconstruction Filter

In order to reconstruct the analog signal from the discrete-level output of the D/A converter, a 16 kHz low-pass filter is necessary. This reconstruction filter must be identical to the anti-aliasing filter used in the source encoder.

2.3 Audio Digitization

This section considers the digitization of the audio signal to a bit stream suitable for transmission purposes. First, the general concepts behind such techniques as uniform (linear), nonuniform (nonlinear) and segmented quantization are presented. Included in this description is a discussion of instantaneous and near-instantaneous companding. The performance of these various approaches to audio digitization are then assessed on both a subjective and objective basis.

2.3.1 Quantization and Companding Techniques

2.3.1.1 Uniform Quantization

Uniform, or linear, quantization is perhaps the most direct method of quantizing a sampled analog signal. With this approach, the quantization levels are uniformly spaced between the signal limits. Since the quantization noise is constant over the entire range of the quantizer, optimum performance is achieved only when the signal occupies the full quantizer range (i.e. the SN_QR varies linearly with input signal level). This implies that, in order to achieve acceptable performance for low signal levels, a large number of quantization levels are necessary. Typically, 14 bits are required for the encoding of high quality audio signals. When this word size is combined with the sampling rate of 32 kHz and the addition of one parity bit per word for error concealment purposes, the result is a transmission rate of 480 kb/s.

2.3.1.2 Nonuniform Quantization

The probability distribution of audio signal amplitudes is not uniform in nature, since signal peaks several dB above the mean program level occur on a relatively infrequent basis. Consequently, providing increasingly fine quantization to the more probable amplitudes and increasingly coarse quantization to the less probable amplitudes will help to improve the SNR without increasing the transmission bandwidth. The result of implementing this type of quantization is to effectively compress the signal at the encoder and then expand the signal at the decoder (a technique referred to as "companding"). In this application, companding is viewed as a method of reducing the transmission data rate from that required for 14-bit linear PCM. Choosing an appropriate companding law such that a specified subjective quality is achieved under the constraint of limited transmission bandwidth is an important design concept.

Although there are several practical approaches for implementing a nonlinear quantization scheme, this report will concentrate on the digital companding of the output from a linear quantizer. This technique simply involves a table look-up process; an approach which is made attractive by recent advances in integrated circuits. The two commonly-used companding laws and the concept of segmented quantizaion will be discussed in the following sections.

2.3.1.2.1 μ-Law Companding

The variation of quantizer step size with signal amplitude is defined by the chosen companding law. Perhaps the most widely used (in speech applications) of these laws is the μ-law, which may be written for positive signal excursions as [13]:

$$V_o = \frac{\ln(1 + \mu V_i)}{\ln(1 + \mu)} ,$$

where μ is the compression factor and V_i and V_o are the input and output signal voltages, respectively. A value of $\mu = 255$ has been found to be nearly optimum for speech and is used in the transmission of digital telephony.

2.3.1.2.2 A-Law Companding

A somewhat different companding law is known as A-law. This characteristic was formulated to be ideally linear at small amplitudes, but logarithmic at large amplitudes, and is defined by [13]:

$$V_o = \begin{cases} \dfrac{AV_i}{1 + \ln A} & 0 < V_i < 1/A \\[2ex] \dfrac{1 + \ln(AV_i)}{1 + \ln A} & 1/A < V_i < 1 \end{cases}$$

A value of A = 87.6 has been selected as optimum for telephony and yields a compression curve virtually identical to that corresponding to the μ-law with μ = 255, except at low signal levels.

2.3.1.3 Segmented Quantization

In order to implement either of the nonlinear companding schemes described above by the digital transformation of the output from a linear quantizer, a piece-wise linear approximation to the compression curve is necessary. This is accomplished by segmenting the smooth companding curves into a series of chords, each of which represents a region of uniform quantization (a fixed step size). The A-law is ideally suited to segmentation, and it is in this form that it is most commonly used.

Figure 2.4 illustrates a 13-segment approximation to the A = 87.6 compression curve. The segment (or chord) number refers to the total number of segments required to cover the entire quantization range (only the positive half is shown). In this case, the chords are arranged such that

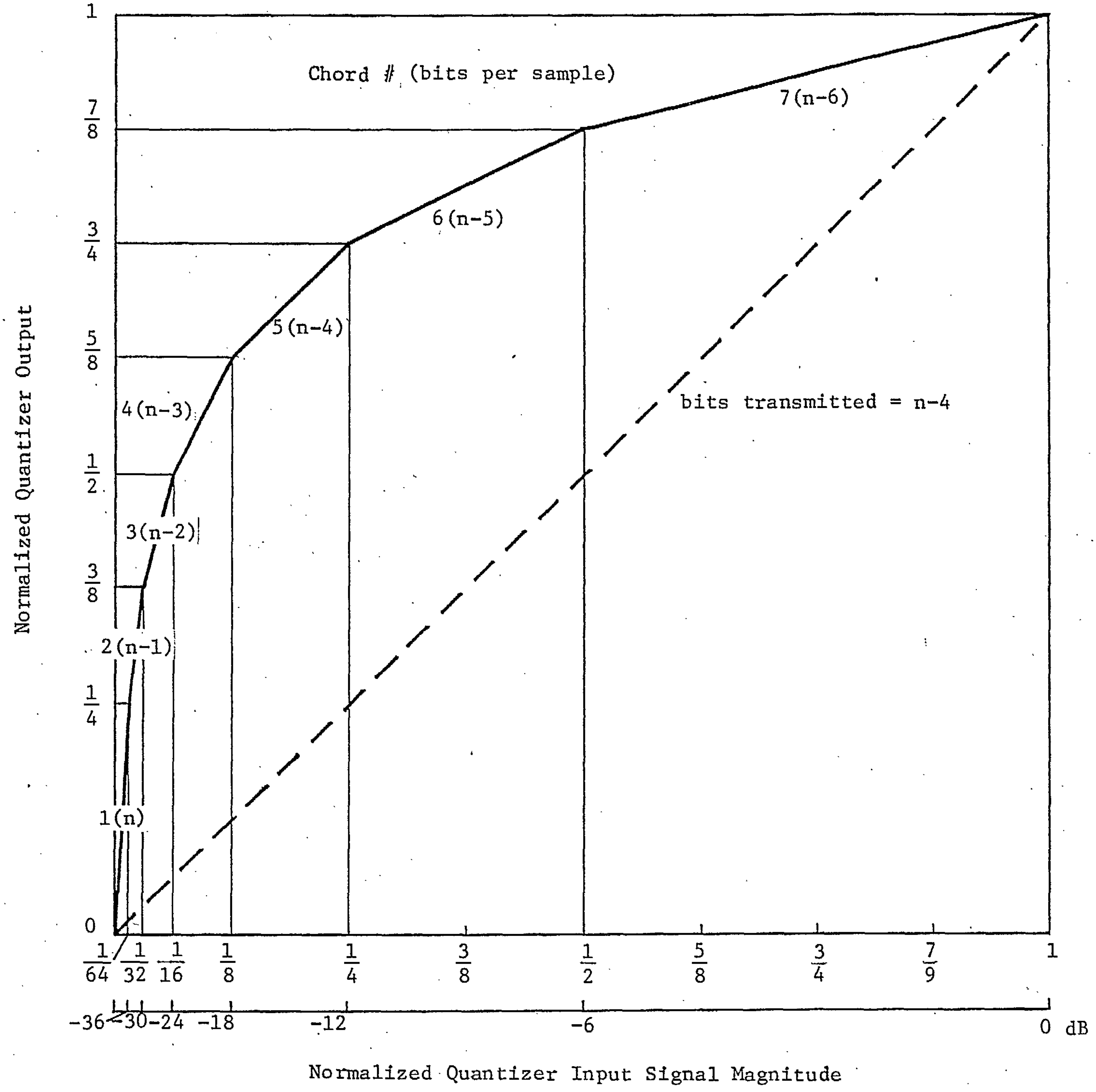


Figure 2.4 - 13-Segment Approximation to the A-Law Compression Curve (A=87.6)

the quantizing slope varies by a factor of two from one chord to the next. Note also that the lower chord is twice as large as the other segments, and thus contains twice as many quantization levels.

The effective resolution (i.e. number of bits quantization) associated with each chord is shown in Figure 2.4, assuming that 14-bit linear quantization is initially used. In this scenario the 13-segment approach corresponds to an effective compression from 14 bits to 10 bits, with the lower chord representing 14-bit resolution and the upper chord representing 8-bit resolution. The bit-mapping procedure corresponding to this companding method is illustrated in Table 2.2.

2.3.1.4 Nearly-Instantaneous Companding

The companding schemes described so far have been instantaneous in the sense that each digital word from the A/D converter is instantaneously transformed to another word containing fewer bits. However, other source coding schemes exist in which the digital companding is performed on the basis of blocks of samples, as opposed to individual samples.

In the nearly-instantaneous companding (NIC) system developed by the BBC [3], blocks of 32 samples are examined to find the maximum sample value in the block. The coding slope is then selected (from a finite set) to accommodate the amplitude of the maximum sample and all the samples in the block are coded to an accuracy determined by that largest sample value. For each block it is necessary to derive a scale factor word which is sent to the decoder and used to control the expander.

Normalized Analog Input	Compressed Analog Code	No. of Bits Effective Resolution	Compressor Output (10 bits)			Normalized Analog Output (lower & upper limits of chord)
			Sign Bit*	Chord Indicator	Significance of Bits 5-10 transmitted**	
			1	2 3 4	5 6 7 8 9 10	
4096 - 8191	448 - 512	8	S	1 1 1	3 4 5 6 7 8	4128 - 8160
2048 - 4095	384 - 447	9	S	1 1 0	4 5 6 7 8 9	2064 - 4080
1024 - 2047	320 - 383	10	S	1 0 1	5 6 7 8 9 10	1032 - 2040
512 - 1023	256 - 319	11	S	1 0 0	6 7 8 9 10 11	516 - 1020
256 - 511	192 - 255	12	S	0 1 1	7 8 9 10 11 12	258 - 510
128 - 255	128 - 191	13	S	0 1 0	8 9 10 11 12 13	129 - 255
0 - 127	0 - 127	14	S	0 0	8 9 10 11 12 13 14	0.5 - 127.5

* Compression characteristic is symmetric about zero. S=0 for positive half and S=1 for negative half

** The bit designations are based on 14-bit linear PCM, where bit 1 is the most significant and bit 14 is the least significant.

Table 2.2 - 13-Segment, 14-bit to 10-bit Instantaneous Companding A-Law (A=87.6)

The distinctive features of NIC can be summarized as follows:

- NIC is an adaptive PCM scheme in which the quantization is uniform and the step size is varied in accordance with the short-term signal level (i.e. on the duration of the block size).

- NIC is a form of companding which lies between instantaneous and syllabic companding.

- NIC extracts some common characteristic of an entire block, and transmits it only once for the block.

- It seems unnecessary to finely quantize a small sample when a neighbouring sample is quite large and hence more coarsely quantized. NIC quantizes each with similar accuracy, assuming they are in the same block.

Figure 2.5 shows the number of bits of resolution corresponding to each of the 5 quantizing slopes used in the BBC's NICAM3 system. The other salient features of this system will be discussed in more detail in Section 2.6.

2.3.2 Objective Performance Evaluation

The achievable signal to quantizing noise ratio is perhaps the most objective measure of an encoder's performance. This factor depends on the nonlinear quantization law, the number of quantization levels and the compression ratio. In order to compare the different companding laws, the following notation will be used:

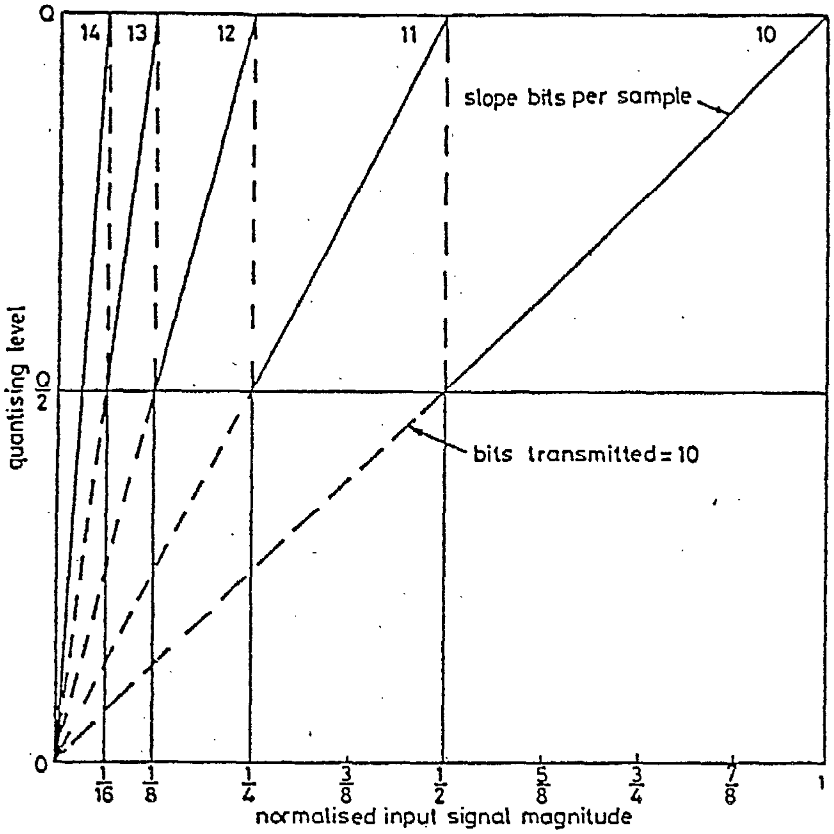


Figure 2.5 - Quantization Slopes for the NICAM3 System [3]

X(a - b - c)

where: X is the type of companding (I for instantaneous and NI for nearly-instantaneous),

a is the number of segments,

b is the original number of bits per sample of the linear quantizer, and

c is the compressed number of bits per sample.

For example, a 13-segment instantaneous companding scheme in which a 14-bit sample is compressed to 10 bits for transmission is denoted I(13-14-10).

Figure 2.6 shows the variation of SN_QR with input signal level for the I(13-14-10), I(11-14-11) and NI(5-14-10) companding schemes. For signal levels more than 35 dB below the quantizer saturation level, all three companding schemes yield the same SN_QR. However, for higher signal levels, the nearly instantaneous companding scheme provides an SN_QR which is approximately 9 dB better than that for the I(13-14-10) scheme. By reducing the number of segments used in the instantaneous companding scheme from 13 to 11, which results in an increase in the number of bits per compressed code word from 10 to 11, the high-level SN_QR is improved by 6 dB. However, this is still 3 dB worse than the NI(5-14-10) performance.

Figure 2.6 also illustrates that the value of SN_QR remains nearly constant at higher levels*. The resultant

*The "ripples" in the high level SN_QR curves are due to the uniform quantization levels used over finite intervals of the input signal range. In all cases, these intervals are 6 dB (factors of 2 in signal amplitude).

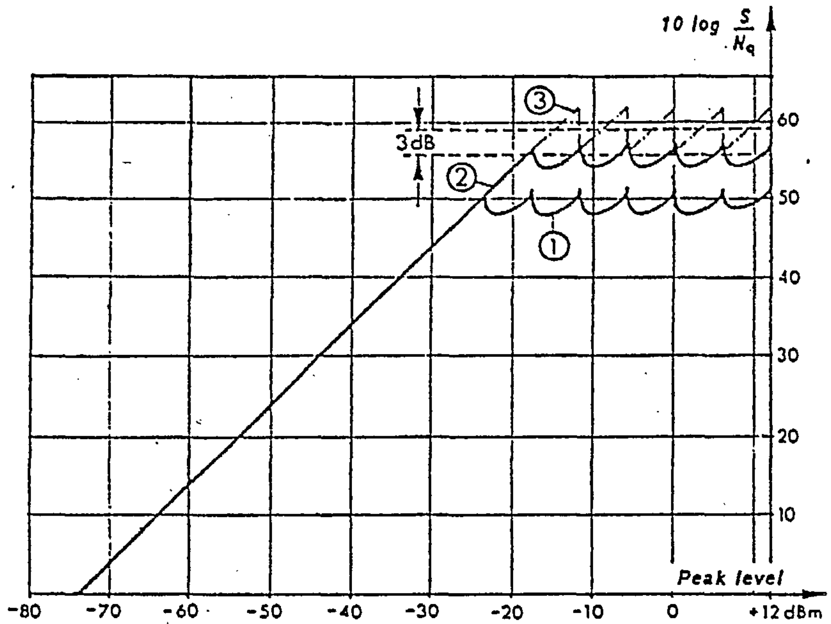


Fig. 2.6 Variation of the signal-to-quantising noise level as a function of the level of a 1-kHz tone, for different types of companding :

instantaneous with 13 segments (curve 1)
instantaneous with 11 segments (curve 2)
near-instantaneous with 5 scales (curve 3).

subjective impression is that of the afore-mentioned program-modulated noise, in which the noise level follows the signal level. The curves, however, give no information concerning the annoyance caused by this type of noise. Listening tests alone will serve to ascertain the quality associated with the different companding laws. CCIR Recommendation 562-1 [14] describes the principles of these tests.

2.3.3 Subjective Performance Evaluation

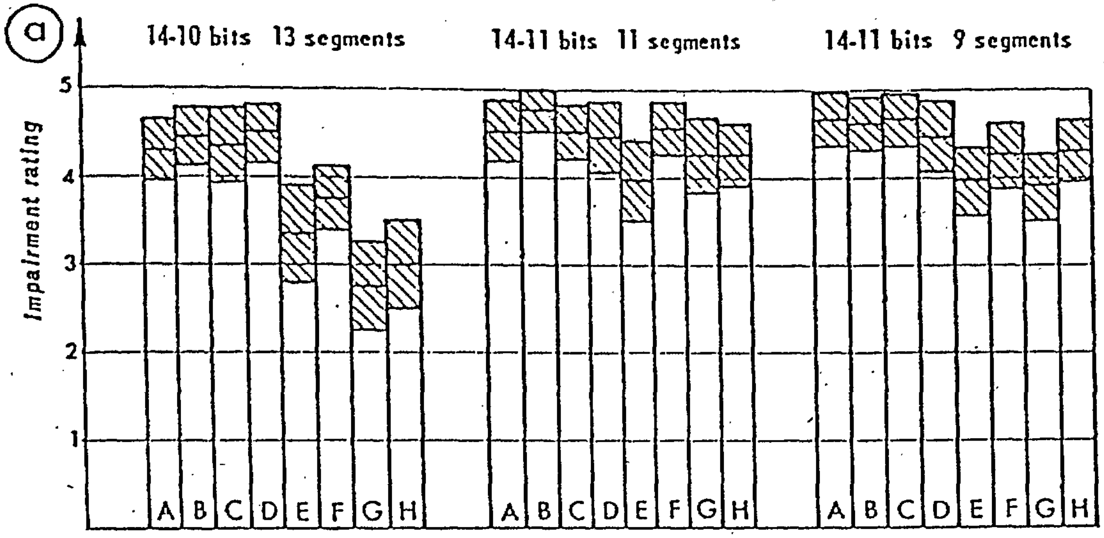
Figure 2.7 shows the results of subjective tests conducted using different instantaneous companding schemes [1]. The tests clearly showed that a significant improvement was realized in going from a 13-segment, 10-bit scheme to a 11-segment, 11-bit scheme, particularly in the case when four identical codecs were connected in tandem. Any further improvement associated with a 4-segment scheme was marginal, at best (for some material, the performance was actually worse).

A different series of subjective tests comparing A-law, several versions of the nearly-instantaneous companding NICAM system, and a coding technique known as TDF (Telediffusion de France) were carried out by the BBC [3]. The A-law scheme is I(13-14-10), the TDF scheme is a nearly-instantaneous companding scheme which requires only 9 bits for transmission, and the NICAM schemes are as defined below:

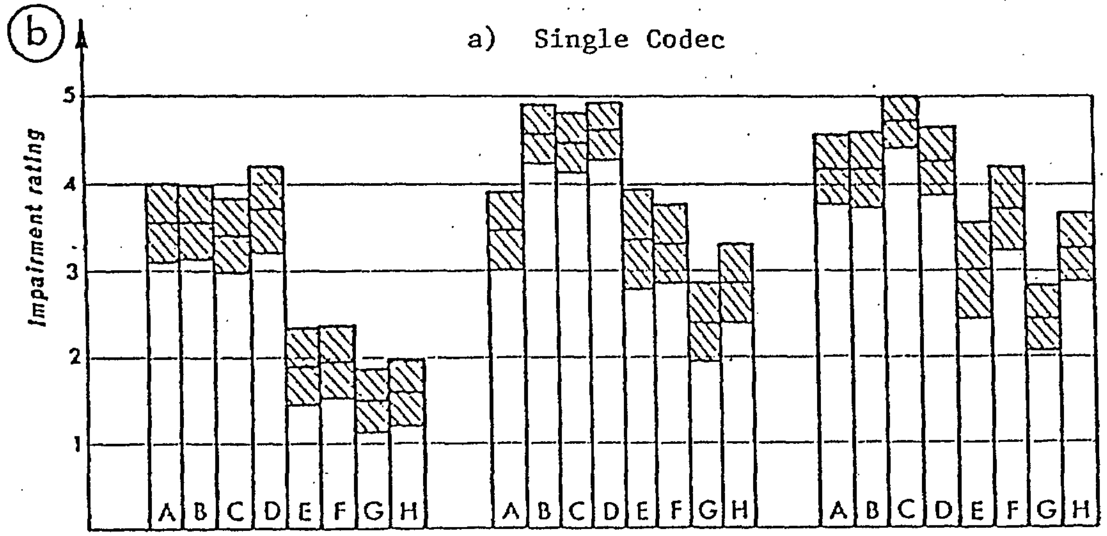
NICAM1 = NI(4-13-10),

NICAM2 = NI(4-14-11), and

NICAM3 = NI(5-14-10).



a) Single Codec



b) Four Codecs in Cascade

A : piano 1 (Mozart)
B : piano 2 (Schubert)
C : violin
D : trumpet
E : glockenspiel
F : « Frère Jacques »
G : 100-Hz tone
H : 500-Hz tone

Sequences F, G and H were synthesised electronically.

<u>CCIR 5-Grade Impairment Scale</u>

5 – Imperceptible
4 – Perceptible, but not annoying
3 – Slightly annoying
2 – Annoying
1 – Very annoying

Figure 2.7 – Subjective Test Results Using the CCIR 5-Grade Impairment Scale [1]

The results of these subjective tests (based on the 7-point CCIR comparison scale) are shown in Table 2.3, with part (a) representing the performance of a single codec and part (b) representing the performance of 4 codecs operating in tandem. In general, the NICAM systems perform much better than either the A-law or TDF systems. Although NICAM2 achieves the best performance, it was considered to be wasteful of transmission bandwidth. Hence, NICAM3 was selected as the companding scheme to be used in Great Britain for reasons of transmission compatibility (as will be discussed in the following section).

Unfortunately, this series of tests only considered the I(13-14-10) A-law companding scheme and not the I(11-14-11) scheme, which was shown previously to provide a considerable performance improvement. In general, comparisons via computer simulation between NICAM3 and I(11-14-11) have shown that there is practically no subjective difference between the noise at high levels [1]. However, it has been observed that there is a just-discernible difference favouring NICAM3 for certain synthetic program material. Hence, the choice between the two companding schemes will be influenced by other factors such as compatibility and interfacing requirements, which will be considered in the following sections.

2.4 Error Performance

In any digital communications system, the occurrence of bit errors is an ultimate determinant of the overall system performance. However, the severity of the degradation due to these transmission errors is dependent on the exact nature of the program material. In this section, such factors as quantization numbering schemes, error correction

Test item: Systems compared A	B	Electronic 'Frere Jacques' test signal	Piano music (Schubert)	Glockenspiel arpeggio	Mean result for the 3 items
A-law	TDF	−1·24 (0·18)	−0·35 (0·26)	−0·88 (0·24)	−0·82
NICAM 1	TDF	+1·29 (0·32)	−0·15 (0·27)	+1·24 (0·25)	+0·72
NICAM 2	NICAM 1	+0·32 (0·16)	−0·09 (0·17)	+0·47 (0·23)	+0·23
13-bit linear p.c.m.	NICAM 1	+0·71 (0·21)	+0·12 (0·21)	— —	+0·42
13-bit linear p.c.m.	NICAM 2	−0·32 (0·11)	−0·03 (0·21)	— —	−0·18
NICAM 3	NICAM 1	0·00 (0·16)	−0·22 (0·26)	+0·03 (0·18)	−0·06

Note: The standard error of the mean grade is given in brackets.

a) Mean Grade Awarded by Listeners in Subjective Tests Comparing Single Codecs

Test item: Systems compared A	B	Electronic 'Frere Jacques' test signal	Piano music (Schubert)	Glockenspiel arpeggio	Mean result for the 3 items
A-law	TDF	−1·76 (0·14)	−0·12 (0·26)	−0·26 (0·33)	−0·71
NICAM 1	TDF	+2·24 (0·26)	−0·41 (0·24)	+1·88 (0·32)	+1·24
NICAM 2	NICAM 1	+0·88 (0·36)	+1·24 (0·25)	+0·65 (0·26)	+0·92

Note: The standard error of the mean grade is given in brackets.

b) Mean Grade Awarded to Listeners in Subjective Tests Comparing 4 Codecs in Tandem

CCIR 7-Point Comparison Scale

+3 – A much better than B
+2 – A better than B
+1 – A slightly better than B
0 – A same as B
−1 – A slightly worse than B
−2 – A worse than B
−3 – A much worse than B

Table 2.3 – Subjective Test Results Using the CCIR 7-Point Comparison Scale [3]

coding and error detection/concealment schemes are considered with respect to the transmission of high-quality stereo digital audio programs.

2.4.1 Quantization Numbering Schemes

In a PCM system (whether it is linear or nonlinear), an important consideration is the exact manner in which the quantization levels are represented as digital words. This concept affects the subjective performance of the source codec because the severity of the impairment due to bit errors depends on which bit within the digital code word is affected. In effect, the noise power associated with digit errors is weighted by the significance of the affected bit.

In reference [15], the performances of several different PCM codes were evaluated. It was shown that the commonly-used binary-folded PCM (i.e. binary plus sign bit) is less sensitive to bit errors than either natural binary or gray-coded binary for A-law companding. However, it was also demonstrated that a PCM code exists which is even less sensitive overall to digital noise than binary-folded PCM. The code developed in [15] is known as the minimum distance code (MDC) and is based on equalizing the error magnitude over all possible error sequences. Hence, an overall performance improvement is achieved by trading off the relatively infrequent large erroneous spikes (to which binary-folded PCM is susceptible) for smaller noise spikes which occur more frequently.

In a practical system, the use of de-emphasis at the decoder has the effect of reducing the severity of the clicks due to bit errors in the most significant positions when binary-folded PCM is used (as was discussed in Section 2.2.1). Also, when error concealment techniques (which would be used regardless of the type of PCM code) are considered, it is important that the noise power associated with errors in the unprotected lower significant bits be kept to a minimum. These considerations, along with the inherent simplicity of the coding technique, make binary-folded PCM the recommended choice for digital audio codecs.

2.4.2 Error Correction Coding

In a practical transmission system, error protection coding is generally used in order that the BER at the receiver meet some specified performance criterion. This is particularly true for transmission via satellite, where the use of error correction coding can reduce the requirement for (valuable) transponder capacity.

Two types of transmission errors generally occur in practice; random and burst errors. Error correction coding is particularly adept at correcting random errors and, if used in conjunction with an appropriate interleaving scheme, can handle burst errors of relatively short duration. Long bursts of errors due to such factors as frame loss cannot be corrected in this manner.

For transmission of digital audio, a BER of 10^{-7} is desirable. However, it has been shown that if an error concealment scheme is used, BER's as high as 10^{-5} may be tolerated before any perceptible degradation is noticed [9]. Figure 2.8 shows the degradation in subjective performance that occurs for bit error rates above 10^{-5}.

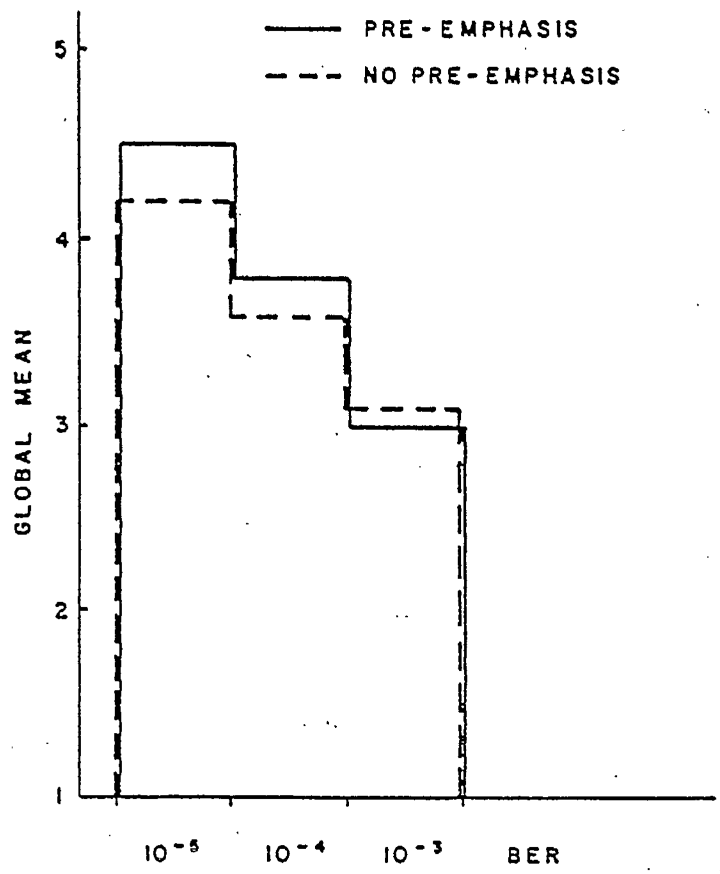


Figure 2.8 - Impairment versus Bit Rate Using the CCIR 5-Grade Impairment Scale [10]

The selection of an error correction code and interleaving technique will be the subject of a subsequent report in this project. For now, it will be assumed that a BER of 10^{-5} may be realized and that the bit errors are distributed in a random manner.

2.4.3 Error Concealment Techniques

Given that the digital audio system is to operate in an environment where random bit errors persist, a significant improvement in subjective performance can be achieved by using error concealment techniques. In general, these techniques consist of detecting when a sample is in error and replacing the corrupted sample with a better estimate of its true value. If multiple word errors occur in succession, the output is eventually muted.

The simplest method of applying error detection coding to the digital sample words is to add a parity bit to each word. Although this only enables an odd number of bit errors to be detected (i.e. an even number of errors will pass undetected), the occurrence of single bit errors will dominate the error patterns. This is demonstrated by Table 2.4, in which the probabilities of single and double bit errors are computed for a 11-bit word size and varying BER. These figures indicate that double errors will occur on such an infrequent basis that their effect should be tolerable.

There are several methods by which a corrupted digital word can be replaced. The simplest of these is zero order extrapolation, which consists of replacing the erroneous word with the last previous correct sample. This is based on the assumption that the change in the audio signal level from sample to sample will generally be small enough that

Bit Error Probability	Probability of Single Bit Error in 11-Bit Word	Probability of Double Bit Error in 11-Bit Word
10^{-5}	1.1×10^{-4}	5.5×10^{-9}
10^{-6}	1.1×10^{-5}	5.5×10^{-11}
10^{-7}	1.1×10^{-6}	5.5×10^{-13}

Table 2.4 - Word Error Probabilities Versus Bit Error Probability

such a replacement will not be objectionable. More complicated replacement methods can be used as well, such as first order interpolation. However, these are considerably more complex to implement and yield only marginal improvements. A BBC investigation into this problem demonstrated that the use of zero order extrapolation reduced the bit rate at which signal degradation becomes "just perceptible" from 10^{-7} to 10^{-5} [9]. Hence, provided that a BER $< 10^{-5}$ is achievable, zero order extrapolation is the preferred approach.

The number of bits in a digital word that should be protected against errors in this manner represents a tradeoff between the impairments due to bit errors and the impairments due to imperfections in the bit error concealment technique. This consideration arises from the fact that replacing a word which has an error in its least significant bits may result in a greater degradation than would have occurred had the erroneous word been left alone. The BBC report [9] suggested that at least the 5 most significant bits should be protected by the parity bit. CCITT Recommendation J.41 [4] states that the parity bit shall protect the 7 most significant bits of the 11-bit digital word.

2.5 Transcoding and Cascading

Satellite transmission of high quality audio from the production studio to a local network distribution centre is only one stage in the overall audio distribution system. In this light, such factors as the transcoding required between different coding schemes and the cascading of multiple codecs must be considered.

2.5.1 Transcoding

Digital techniques are being increasingly used in broadcasting studios, program production studios and the disc-recording industry for the purposes of generation, processing, storage and transmission of digitized analog signals. Accordingly, the signal quality requirements for these various processes will all be different. In order to interface the transmission system to these other systems, a transformation of one or more of the following parameters may be required:

- sampling frequency,
- quantizing resolution, and
- companding technique.

The most likely interfaces that will be necessary will be between the transmission system and either the production studio audio signal or the output from a compact disc system. The production studio uses a 48.0 kHz sampling frequency and at least 16-bit (perhaps 20-bit) linear PCM, and compact disc technology is based on a 44.1 kHz sampling frequency and 16-bit linear PCM. The most direct approach for both of these cases is to decode the digital signal to its analog form and then re-encode it in the format used for transmission. However, this scheme results in the degradations associated with each system combining in a cumulative manner. There are more elegant approaches to this problem which incur less of a degradation.

The recommended approach is to perform any necessary transformation entirely in the digital domain. Since the two audio soruces under consideration both use linear PCM, the transformation of coding schemes is greatly simplified (i.e. there is no conversion from one companding format to another). The reduction in quantization resolution is straightforward since the least significant bits of the 16-bit (or 20-bit) word can simply be discarded to obtain the required 14-bit accuracy, with its attendent quantization noise performance. Similarly, the transformation of sampling frequency to the 32 kHz transmission standard can be achieved by an existing device that is capable of changing a particular sampling frequency to any other sampling frequency [16]. This procedure is simplified in the case of a production studio by the integral relationship of the 48 kHz standard to the 32 kHz sampling rate used for transmission.

If an interface is required between two digital codecs that make use of different companding laws, matters are complicated somewhat. In general, the digital signal would be expanded and then compressed according to the desired compression curve. A specific example will be considered in the interface requirements of Section 2.1.3.

2.5.2 Cascading

There may be some instances within the audio distribution network where the analog signal must be recovered, perhaps re-processed (e.g. edited), and then re-encoded into a digital format for transmission. These situations of multiple codecs operating in tandem will have the effect of reducing the overall quality of the received audio signal. However, the exact amount of degradation in the various codec operational parameters is difficult to assess. Because of the nonlinear and random nature in which most of these degradations add, only worst-case approximations concerning their cumulative effect can be made.

The audio performance specifications of CCITT Recommendation J.21 apply to the case of three digital codecs connected in tandem at audio frequencies (i.e. back-to-back). This scenario is considered to be a worst-case situation and, in most instances, only a single digital encoding operation will be necessary between the source and destination. Accordingly, CCIR Report 647-2 outlines the audio specifications to be met by a single digital codec operating in a back-to-back configuration.

2.6 Codec Implementation Considerations

Having examined the technical details associated with the choice of a digital source coding scheme, we shall now turn to the more practical considerations in the development of a satellite distribution system for digital audio. Perhaps the most important consideration involved in this selection is compatibility with existing (and proposed) standards. This is discussed first, followed by a description of the existing equipment offered by various manufacturers. The general configuration of a basic system is then examined and the interfacing requirements necessary for compatibility with the U.S. network are considered.

2.6.1 Compatibility

Perhaps the most important consideration in the selection of a digital audio encoding scheme for program distribution via satellite is that the system be compatible with the standards defined for North America. The international community, with the exception of Britain, France and, possibly, the USSR, are very firmly moving towards standardizing on instantaneous digital companding at this year's CCITT meeting. A draft recommendation has already been submitted to the CCITT, and a similar document was submitted to the CCIR and agreed upon between Telecom Canada and AT&T as a partial standard for North America.

The companding method selected is 11-segment, 14-to-11 bit A-law with A = 43.8. Although details of the bit mapping are still to be worked out, it is likely that two standards will result - one for North America and another for Europe. This is due to the different transmission standards used in North America (1.544 Mb/s) and Europe (2.048 Mb/s).

The standard which Britain, France and, possibly, the USSR are opting for is NICAM3. This version of the nearly-instantaneous companding scheme was selected over NICAM2 (which actually performed better) for reasons of transmission compatibility. This encoding method permits 6 monaural channels (3 stereo programs) to be transmitted in the standard 2.048 Mb/s format (compared with 5 for NICAM2). These details are summarized in Table 2.5 along with the parameters for the various systems throughout the world.

In North America, the transmission standard is 1.544 Mb/s (known as a T1 carrier). By using a compression scheme that reduces the data to 11-bit samples, plus one parity bit, four 384 kb/s monaural channels (two stereo programs) can be multiplexed onto a T1 line. There is no need to use a compression scheme that outputs 10-bit samples, since a T1 carrier would still be able to support only four channels, with wasted capacity. In addition, as was shown in Section 2.3.3, 14-11 bit companding provides the performance improvement over 14-10 companding necessary to satisfy the requirements of multiple codecs operating in tandem.

	Coding methods using different companding principles								Units
	Near-instantaneous		Instantaneous						
	A1 A2	A3	B1	B2	B3	B4	B5	B6	
Nominal bandwidth	0.04-15[1][2]	0.04-15[1]	0.04-15[1]	0.04-15[1]	0.04-15[1]	0.04-15[1]	0.04-15[1]	0.05-7[3]	kHz
Pre/de-emphasis	CCITT Rec. J.17 6.5 dB/ 0.8 kHz[4]	CCITT Rec. J.17 6.5 dB/ 0.8 kHz[4]	See Note[5]	None	CCITT Rec. J.17 6.5 dB/ 0.8 kHz[4]	CCITT Rec. J.17 6.5 dB/ 0.8 kHz[4]	CCITT Rec. J.17 6.5 dB/ 0.8 kHz[4]	CCITT Rec. J.17 6.5 dB/ 0.8 kHz[4]	–
Overload point[6]	+12	+12	+12	+12	+12	+12	+15	+15	dBm0s
Sampling frequency	32	32	32	32	32	32	32	16	kHz
Companding law	5 ranges	5 ranges	13 segments	7 segments	7 segments	11 segments	11 segments	11 segments	–
Bit rate reduction	14/10.1	14/10.1	14/10	13/11	14/12	14/11	14/11	14/11	bits
Finest resolution and	14	14	14	13	14	14	14	14	bits/sample
corresponding noise	−65	−65	−66	−55	−65	−65	−62	−60	dBq0ps
Coarsest resolution at +9 dBm0s/f_0*	10	10	8	10	11	9	9	9	bits/sample
and corresponding noise	−41	−41	−30	−37	−47	−35	−32	−30	dBq0ps
Resolution at +9 dBm0s/60 Hz and	12	12	10	10	12	11	12	12	bits/sample
corresponding noise	−55	−55	−42	−37	−59	−47	−50	−48	dBq0ps
Source coding	323	323[7]	320	352	384	352[8]	352[8]	176[8]	kbit/s
Error protection	12 13	11[7]	16	32	0	32	32	16	kbit/s
Framing and signalling	3 2	4[7]	0.66	0	0	0	0	0	kbit/s
Service bit rate	338	338	336.66	384	384	384	384	192	kbit/s
Transmission bit rate	~ 340** 384	340** 384	336.66** 384	384	384	384	384	192	kbit/s
Proposed by	France	UK	Italy	Japan	USSR	Switzerland, Canada	Federal Republic of Germany, Netherlands, Denmark, Norway	Federal Republic of Germany, Netherlands,	
Doc. CMTT (1978-82)	35, 110, 93, 112, 277	240	203 (1970-74) 271, 275	35, 51, 110, 231	35, 51, 110	110, 35, 276	6, 80, 81, 82, 110, 35, 251, 258, 297, 234	81, 82, 251, 258, 234	

* f_0: zero loss frequency of pre-emphasis.

** Dedicated frame.

Table 2.5 - Summary of Digital Encoding Scheme [5]

These considerations indicate that there is not much flexibility in the choice of a digital audio encoding scheme for use in the proposed satellite distribution network. In addition to this, CBC (who represent one of the major potential users of the proposed system) have indicated that they will only be interested in a system that conforms to the North American standard agreed upon by Telecom Canada and AT&T. Hence, an 11-segment approximation to A-law companding (A = 43.8) which provides a 14 to 11 bit compression factor will be the source coding scheme used. The segmented approximation to the compression curve and the corresponding bit-mapping table are shown in Figure 2.9 and Table 2.6, respectively.

2.6.2 Existing Equipment

Two manufacturers produce digital audio codecs that are suitable for this system. Both Tau-tron in the U.S. and Bayly Engineering in Canada have codecs that comply with the CCIR/CCITT specifications. In fact, these manufacturers were sent draft versions of the recommendations and consulted with Telecom Canada in the definition of their final system parameters.

The specification sheets for the Tau-tron and Bayly codecs are provided in Appendices 2A and 2B, respectively. Both use 14-11 bit, 11-segment A-law companding and have very similar specifications overall. The exception is that the Bayly codec also has the capability of operating in the 15-11 and 14-10 bit companding mode. Both digital program units (as they are referred to) may be factory-designed to operate with either a 15 kHz or 7.5 kHz audio program and both have optional pre-/de- emphasis circuits.

As a comparison to these two units, the specificatons for the Wegener analog FM channel unit, designed for SCPC

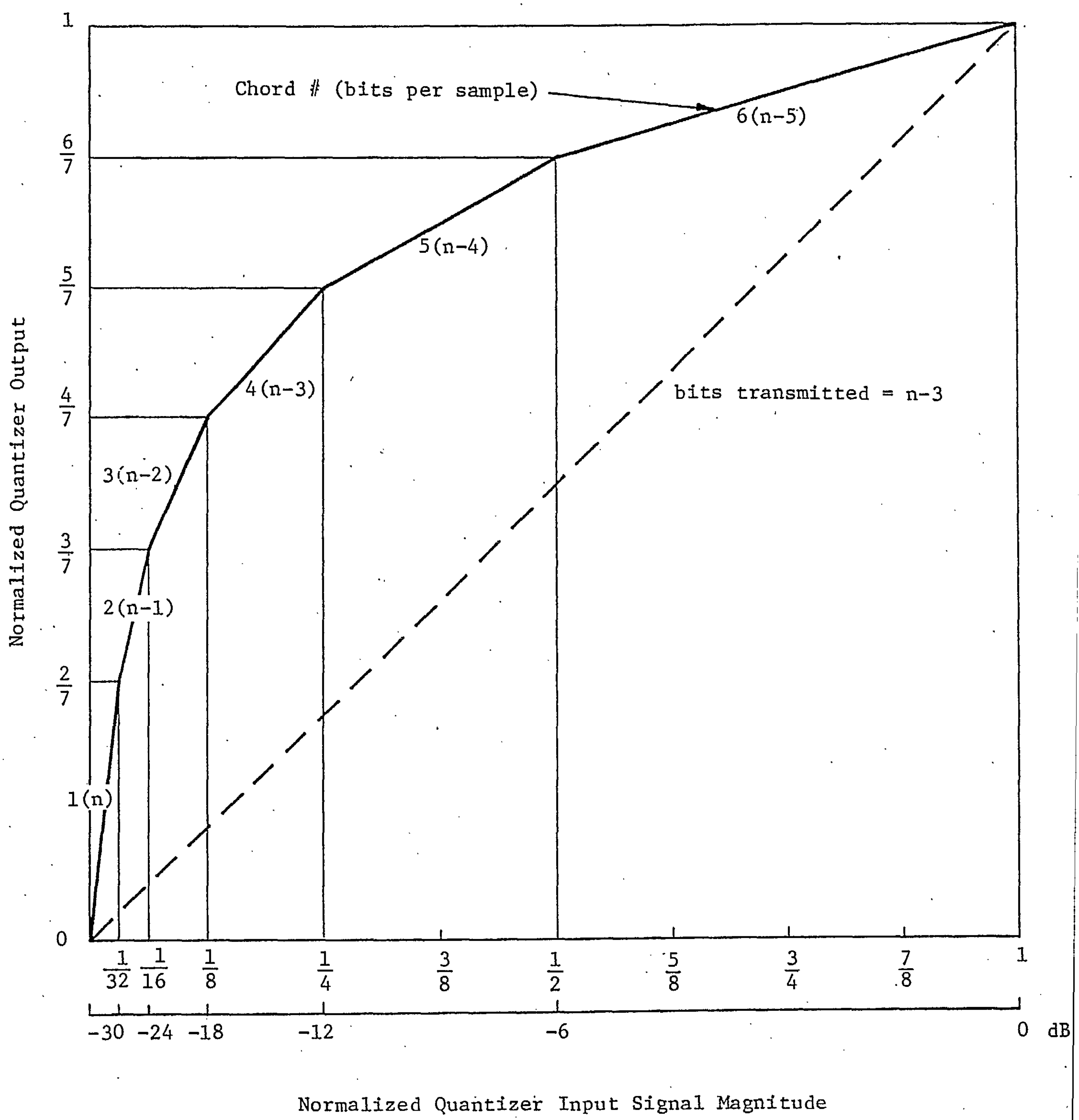


Figure 2.9 - 11-Segment Approximation to the A-Law Compression Curve (A=43.8)

Normalized Analog Input	Compressed Digital Code	No. of Bits Effective Resolution	Compressor Output (11 bits)											Normalized Analog Output (lower & upper limits of chord)
			Sign Bit*	Chord Indicator			Significance of Bits 5-11 transmitted**							
			1	2	3	4	5	6	7	8	9	10	11	
4096 - 8191	768 - 895	9	S	1	1	0	3	4	5	6	7	8	9	4112 - 8176
2048 - 4095	640 - 767	10	S	1	0	1	4	5	6	7	8	9	10	2056 - 4088
1024 - 2047	512 - 639	11	S	1	0	0	5	6	7	8	9	10	11	1028 - 2044
512 - 1023	384 - 511	12	S	0	1	1	6	7	8	9	10	11	12	514 - 1022
256 - 511	256 - 383	13	S	0	1	0	7	8	9	10	11	12	13	257 - 511
0 - 255	0 - 255	14	S	0	0	7	8	9	10	11	12	13	14	0.5 - 255.5

* Compression characteristic is symmetric about zero. S=0 for positive half and S=1 for negative half

** The bit designations are based on 14-bit linear PCM, where bit 1 is the most significant and bit 14 is the least significant

Table 2.6 - 11-Segment, 14-bit to 11-bit Instantaneous Companding A-Law (A=43.8)

operation, is provided in Appendix 2C. The specifications indicate that the total harmonic distortion is considerably worse than for the digital channel units and that the C/N_o requirement is somewhat greater. The big disadvantage of this system is that any distortions introduced in the backhaul link from the production studio to the transmit earth station will directly degrade the analog audio signal. In a digital system, the audio performance is relatively unaffected by the distortions introduced on this link, provided that a satisfactory BER is maintained. Also, the Wegener approach requires a separate modulator for each monaural channel, compared to the single modulator that can handle a multiplexed bit stream in the digital case.

As a final note, Appendix D contains the specifications for the Scientific Atlanta digital audio program unit. This unit uses a 15-11 bit, 13-segment approximation to the $\mu = 255$ compression curve (although this approximation is exactly the same as that which is used for the $A = 87.6$ curve). The Scientific Atlanta unit is being used for digital audio transmission in the U.S. by such networks as ABC, CBS and RKO.

2.6.3 General System Configuration

Figure 2.10 shows a typical application for the source coding units. A separate 15 kHz unit will be required for both the left and right channel of a stereo program. The 384 kb/s signal from each may then be combined with a 32 kb/s data input to form a composite 800 kb/s bit stream. The multiplexing operation required in this scenario, which may be complicated by the necessity of relaying the 800 kb/s data from a production studio to the satellite earth station via a T1 carrier, will be considered in a separate

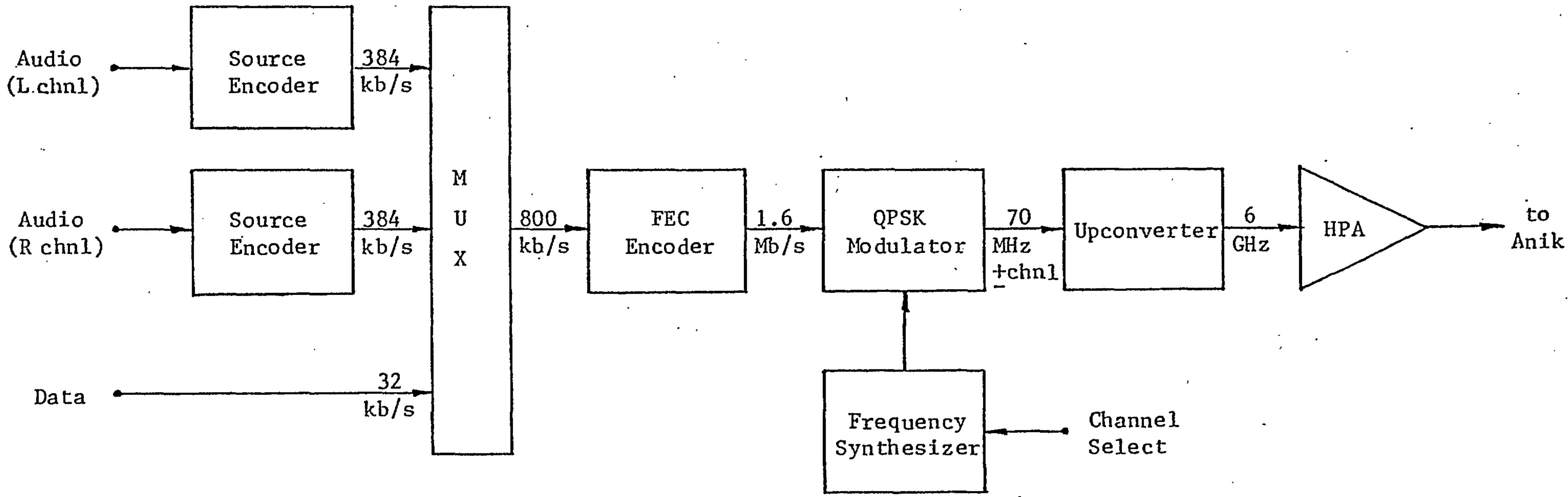


Figure 2.10 – Basic Audio Distribution System

report. Similarly, the operation of the satellite uplink equipment (e.g. modulation and channel coding types) will be considered separately.

In order to design the system for maximum flexibility, each 15 kHz channel should be able to be replaced by two 7.5 kHz channels. In principle, the digital processing of the two 7.5 kHz channels should be able to be shared with the digital processing unit of the 15 kHz encoder (i.e. by changing the analog front-end and multiplexing the two 7.5 kHz channels into a single 15 kHz bit stream). However, in the case of either the 7.5 kHz or 15 kHz program units, both the analog are digital processing operations are contained on a single card. Hence, in practice, it may be appropriate to use a separate program unit for each 15 kHz or 7.5 kHz channel, as shown in Figure 2.11.

2.6.4 Internetwork Interfacing Requirements

An important consideration in the development of an audio distribution network is that it have the capability of being compatible with other networks operating in North America. More specifically, this implies that our system should be capable of interfacing with the Scientific Atlanta system used in the U.S.

As mentioned in Section 2.6.2 (and Appendix 2D), the Scientific Atlanta program unit is based on a 15-11 bit, 13-segment approximation to the $\mu = 255$ compression curve*. The bit-mapping table used in this approach is illustrated in Table 2.7. Since this is a different compression characteristic than that used in the 14-11 bit, 11-segment A-law companding scheme proposed for our system, some form of conversion is necessary if compatibility is to be achieved.

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*As mentioned previously, this approximation is identical to the 13-segment approximation to the $A = 87.6$ compression curve shown in Figure 2.4.

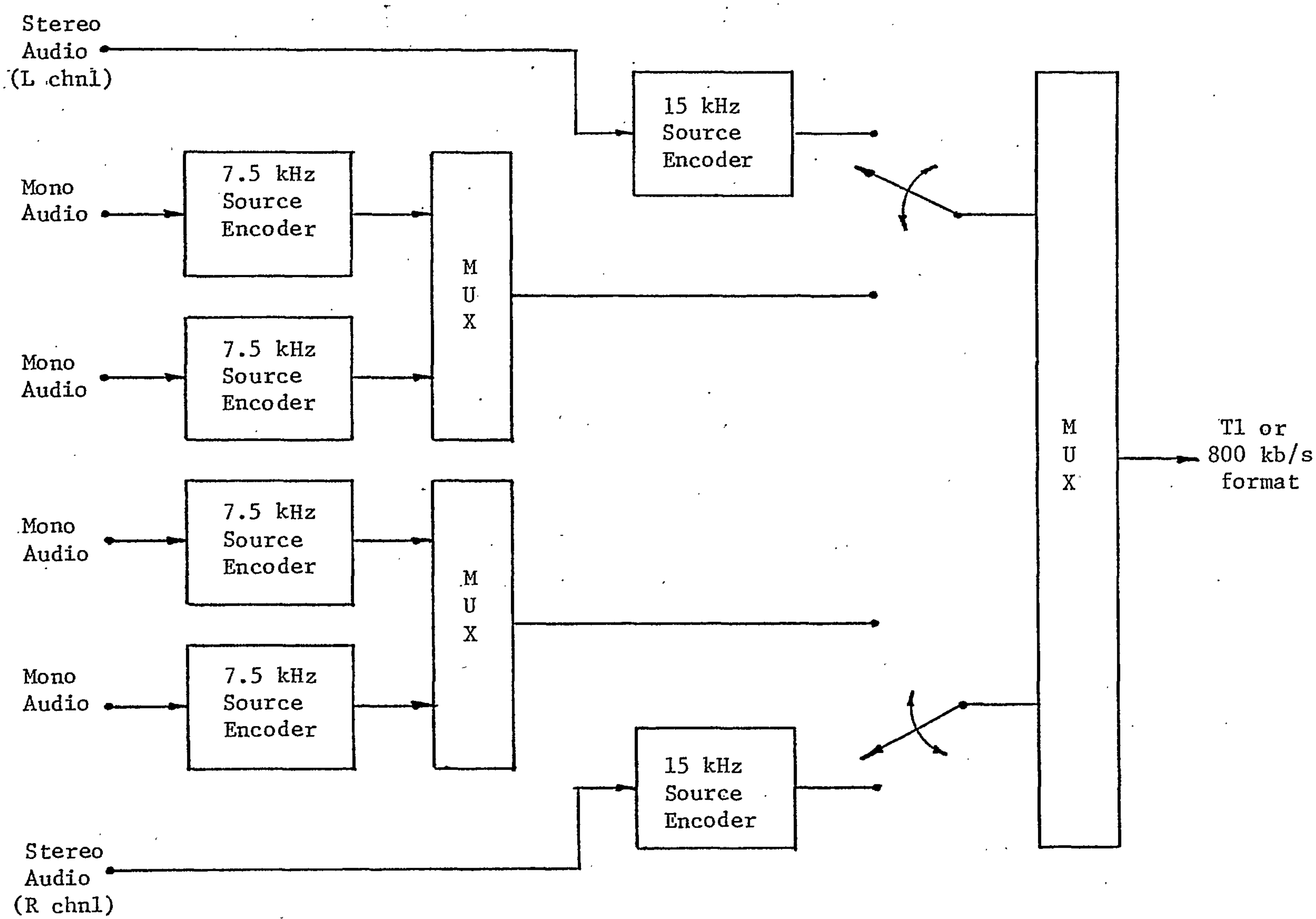


Figure 2.11- Possible Configuration for 15/7.5 kHz Programs

Normalized Analog Input	Compressed Analog Code	No. of Bits Effective Resolution	Compressor Output (11 bits)												Normalized Analog Output (lower & upper limits of chord)
			Sign Bit*	Chord Indicator			Significance of Bits 5-11 transmitted**								
			1	2	3	4	5	6	7	8	9	10	11		
8192 - 16883	896 - 1023	9	S	1	1	1	3	4	5	6	7	8	9		8224 - 16352
4096 - 8191	768 - 895	10	S	1	1	0	4	5	6	7	8	9	10		4112 - 8176
2048 - 4095	640 - 767	11	S	1	0	1	5	6	7	8	9	10	11		2056 - 4088
1024 - 2047	512 - 639	12	S	1	0	0	6	7	8	9	10	11	12		1028 - 2044
512 - 1023	384 - 511	13	S	0	1	1	7	8	9	10	11	12	13		514 - 1022
256 - 511	256 - 383	14	S	0	1	0	8	9	10	11	12	13	14		257 - 511
0 - 255	0 - 255	15	S	0	0	8	9	10	11	12	13	14	15		0.5 - 255.5

* Compression characteristic is symmetric about zero. S=0 for positive half and S=1 for negative half

** The bit designations are based on 15-bit linear PCM, where bit 1 is the most significant and bit 15 is the least significant.

Table 2.7 - 13-Segment, 15-bit to 11-bit Instantaneous Companding A-Law (Λ=87.6)

Referring back to the companding algorithm of Table 2.6, it can be shown that conversion from the 11-segment, A = 43.8 format to the 13-segment, $\mu = 255$ format can be realized by a very simple bit-mapping operation (i.e. so that the output from the program terminal under consideration can be decoded by the Scientific Atlanta terminal). This bit-mapping procedure obeys the following rules:

(i) The upper 5 chord patterns from Table 2.6 are transformed as follows (but bits 5-11 remain unaltered):

110 → 111
101 → 110
100 → 101
011 → 100
010 → 011

(ii) The lowest chord is altered in the following manner:

- if bit 7 = 0:
  - designate chord as 010 (with bits 5-11 the same),

- if bit 7 = 1:
  - drop bit 7,
  - shift bits 5-11 left by one bit, and
  - add dummy bit in bit-11 location.

This transformation allows the codec output to be decoded by the Scientific Atlanta equipment, but only provides 14-bit resolution in the finest quantization levels (the lowest chord), as compared to the 15-bit resolution associated with the Scientific Atlanta approach.

If the codec is to be used to receive transmissions initiated from a Scientific Atlanta terminal, a similar data-mapping procedure can be incorporated into the (digital) front-end of the decoder. The required transformation is as follows:

(i) The upper 6 chord patterns from Table 2.7 are transformed as follows (but bits 5-11 remain unaltered):

111 → 110
110 → 101
101 → 100
100 → 011
011 → 010
010 → 001

(ii) The last chord is altered in the following manner:

- it is designated as chord 000,
- bits 5-11 are shifted one location to the right, and
- bit 11 is discarded.

Note that chord patterns 000 and 001 both have the same 14-bit resolution as the bottom chord (which has twice as many quantizing levels as the other chords) of the 11-segment, A = 43.8 algorithm.

Also of concern is the manner in which the parity encoding is applied. The Scientific Atlanta system also uses one parity bit for every 11-bit sample, but it is not specified which bits are protected. This must be determined prior to the codec implementation phase.

Hence, it appears that compatibility between the two systems can be achieved in a relatively simple manner, by having optional circuitry which implements the required code conversion operation.

2.7 Summary

As a brief summary to this chapter, the major operational features of the digital audio source codec are listed below:

(i) the source codec will use a 14-bit to 11-bit, 11-segment approximation to the A-law companding scheme (with A = 43.8),

(ii) a single parity bit will be used to protect the 7 most significant bits of each digital word for error concealment purposes,

(iii) pre- and de-emphasis should be optionally available in the encoder and decoder units, respectively, and

(iv) the system should be capable of interfacing with the Scientific Atlanta equipment by virtue of some relatively simple code conversion circuitry.

All other aspects of the audio performance of the source codec shall conform to the specifications outlined in CCIR Report 647-2 and CCITT Recommendations J.17, J.21 and J.41.

3.0 TASK 4: MULTIPLEXING

3.1 Introduction

The multiplexing technique selected for the radio program distribution system will be based on the assumption that the traffic requirement comprises one or more of the following:

(i) one stereophonic high-quality audio channel (i.e. 15 kHz left and right channels),

(ii) two monophonic 15 kHz audio channels,

(iii) four 7.5 kHz audio channels, and

(iv) one or more low-rate data channels.

The combination of a stereophonic audio channel plus a low-rate data channel (typically 32 kb/s) is a immediate requirement that complies with the SOW. However, the other options may be provided with a minimal increase in cost or complexity as compared to the overall system.

In general, there will be two scenarios in which the multiplexing scheme will have to operate:

(i) direct transmission of audio signal and data over an SCPC satellite channel (i.e. the production studio and satellite transmit site are collocated), and

(ii) the audio signal and data are relayed via a T1 carrier to the satellite transmit site and then transmitted in an SCPC mode.

Although the first option represents the basic requirement of the multiplexing scheme, the intent is to design a multiplexer that is capable of handling both of these scenarios (operating in conjunction with a T1 interface in the case of the second option). Hence, the requirements for interfacing with a T1 carrier will have to be considered as part of the overall design.

In Section 3.2, the general properties and requirements of a multiplexing scheme are first considered. Then, Section 3.3 deals with the basic multiplexing requirements for SCPC transmission and Section 3.4 considers the integration of possible system options.

3.2 General Multiplexing Considerations

In this section, the general properties and tradeoffs associated with the selection of a multiplexing scheme are considered.

3.2.1 Continuous Versus Packet Multiplexing

The two basic multiplexing methods that can be used are:

(i) continuous, and

(ii) packet.

The different aspects of these two methods in the context of radio program distribution were discussed in a previous MCS report [17]. This study emphasizes the fact that a continuous multiplexing system is well-suited for the distribution of program sound channels, since reassignments of these types of channels are not likely to occur due to the continuous nature of the sound. In this section we shall describe the salient features of the continuous and packet multiplexing techniques.

In continuous multiplexing, the basic structure is the frame, in which fixed groups of bits play prescribed roles according to their position relative to the frame boundaries. Thus, particular bits are dedicated to the transmission of information relating to one input signal. A predetermined framing pattern enables the receiver to synchronize to the frame structure and extract from the incoming bit stream the particular bits which correspond to each of the multiplexed signals. In continuous multiplexing, the bit rate of each signal may be a precise sub-multiple of the final serial bit rate. This is known as a synchronous multiplexing scheme.

Continuous multiplexing has the following advantages:

- it requires only a small proportion of overhead bits for signalling (e.g. framing bits),

- with the low overhead required for secure synchronization, a small amount of channel capacity can more easily be justified for any signalling required to indicate a change in the of use of the channel, thus providing flexibility,

- the capacity of any channel not used for an audio program can be assigned to additional data channels without any increase in overhead, and

- for a fixed data rate structure, the receiver complexity is minimized.

In packet multiplexing, the final bit stream is composed of successive blocks called packets, each with two parts: heading and data. The heading, which is specific to the transmitter, serves to synchronize the receiver and to identify the source of the data inserted in the packet. Thus, the heading contains a sync pattern followed by a prefix. The data portion of the packet contains the useful information and is composed of data from only one input signal. This method is flexible in the sense that it does not impose a predetermined content on the final bit stream. Also, it is possible to operate either synchronously or asynchronously. In the latter case, the bit rates of the various input signals do not need to be directly related to the final serial bit rate. The system can thus easily accommodate asynchronous signals.

From these considerations, it is apparent that a continuous multiplexing scheme is more appropriate to an audio program distribution network. In this application, both the source and the format of the audio signals will be fixed, thus providing no requirement for the flexibility inherent in a packet multiplexing scheme. Also, it is important that the amount of overhead information required be kept to a minimum (an advantage of continuous multiplexing) for the sake of minimizing the use of (expensive) satellite transponder capacity.

3.2.2 Word Versus Bit Interleaving

In a PCM digital audio system, a particular audio signal is represented by a sequence of fixed length code words (12 bits in duration, including the parity bit). For a continuous multiplexing scheme, these words can be organized in one of two basic formats. If the audio signals (from either a single source or from multiple origins) are multiplexed together such that the digital words are transmitted in their original format, the multiplexed output signal is termed word interleaved. Alternatively, each channel code word may be scanned one bit at a time to produce a bit-interleaved mutliplexed signal. These procedures are illustrated for a 3-channel example in Figure 3.1.

Word interleaving has the following characteristics:

- it can operate with a 'shared' codec approach in which code words are produced sequentially from the samples of the incoming analog signals,

- depending on the input signal level and the code word length, a long sequence of 0's or 1's may occur (this may pose synchronization problems and may require 'zero' pattern elimination circuitry),

- in the presence of relatively short error bursts (e.g. burst length $<$ PCM word length) at most two successive code words will be affected (however, the errors will likely not be able to be corrected by the forward error correction coding employed).

Bit interleaving on the other hand has the following characteristics:

- it will require either an individual codec for each incoming channel or data buffers (in either case leading to increased complexity),

- the probability of occurrence of long sequences of 0's or 1's is less, and

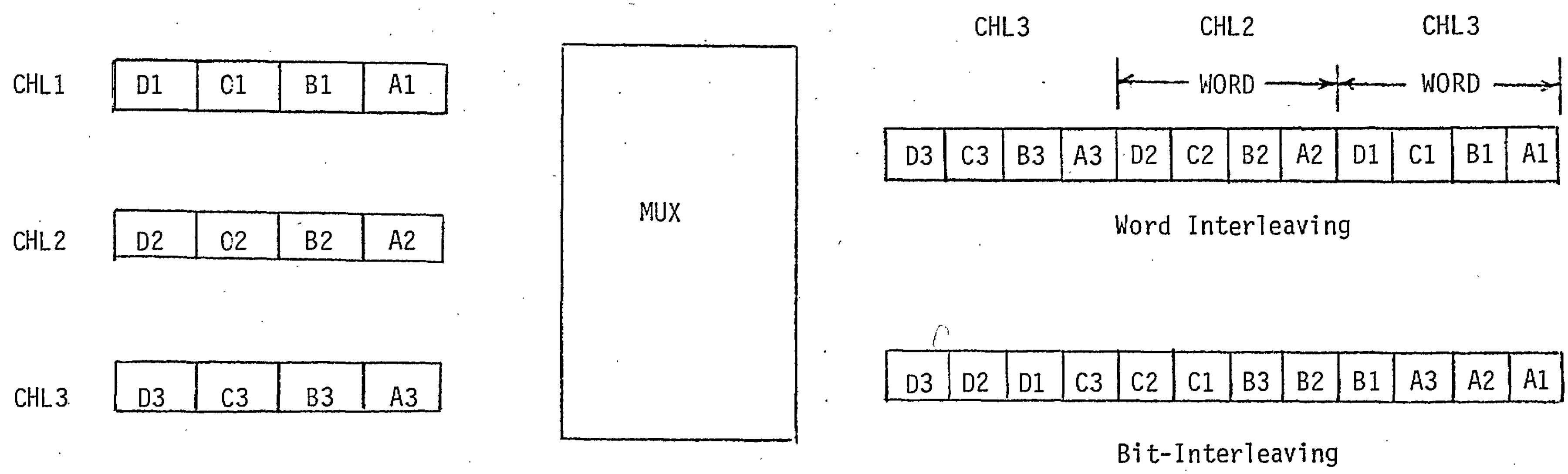


Figure 3.1 Bit and Word Interleaving Concepts

- in the presence of an error burst, several code words will be affected (however, the errors may be distributed such that the forward error correction coding is capable of correcting the resulting errors).

The choice of a particular interleaving approach will be significantly influenced by the approach adopted for the codec implementation.

3.2.3 Frame Synchronization

For the transmission of a multiplexed audio signal, the receiver must be able to synchronize to the frame format of the signal and extract the digital code words in the correct manner. In this respect, the frame structure should be formed so as to optimize the synchronization parameters. This involves the choice of synchronization codes to minimize the false synchronization probability and the choice of frame length so that the frame acquisition time is minimized. The number of control bits for this purpose should be made as small as possible (i.e. high frame efficiency).

In general, frame synchronization words in a multiplexed frame can be organized in two ways:

- if all the bits of the unique frame synchronization word(s) are provided at one location (normally at the beginning of the frame), this approach is called bunched framing, or

- if the sync words are provided at more than one location by interleaving amongst the channel words, this is called distributed framing.

Generally, for a given frame length, the sync acquisition time decreases with the increase of sync word length. Besides the actual sync word, there may be one or more sync word 'imitations' in the same frame. A serial search synchronizer recognizes the first sync word in a frame and looks for the same word in the same location in each subsequent frame. Frame synchronization is established after a finite number of consecutive detections. If the synchronizer first recognizes an imitation word, it is unlikely that the next coincidence attempt will be successful. A failure to confirm sync places the synchronizer back into the search mode where the detector slides bit by bit searching for repeated sync words. Systems based on the serial search concept operate at high bit rates and are cheap to implement. If the sync word length is small compared to the frame length, the search time could stretch over several frames before locking to the authentic sync word.

A reduction in the overall mean search time can be obtained by using the more complex parallel search approach. Parallel searching reduces the effect of imitations by marking the location of all sequences that match the sync word, and immediately ignoring those that do not repeat at one-frame intervals. Within a few frames, all imitations will be ignored and sync lock will be established with acceptable certainty. However, the implementation of a parallel search synchronizer is relatively complex. This technique may be particularly useful for situations where the sync word is small and distributed framing structures are used. In cases where imitations rarely occur there is very little which can be gained using the parallel technique.

Once frame synchronization is acquired, it must be maintained by the receiver demultiplexer so that correct decoding of the audio program channel can be realized. Due to channel noise and other system impairments, there is a small but finite probability that frame synchronization will be lost due to errors in the framing pattern. In order to keep the frame length within practical limits for fast acquisition purposes, it is usually required that frame violation be detected several times in succession before a decision is made that frame sync was lost.

Overall, the frame structure (including the synchronization word) must be selected with the objective of optimizing the following parameters: sync acquisition time, false sync probability, and probability of sync loss. However, these factors represent conflicting objectives, particularly when the constraint of maximizing the frame efficiency is considered. In an audio distribution system, where the signal is transmitted in a continuous mode, the most important parameter is the probability of losing synchronization. A loss of synchronization implies that several successive code words will be lost, the subjective effect of which will be similar regardless of how quickly sync can be re-established. Thus, although the time to re-acquire sync must be reasonable, it is extremely important to minimize the probability of a loss of sync occurring at all.

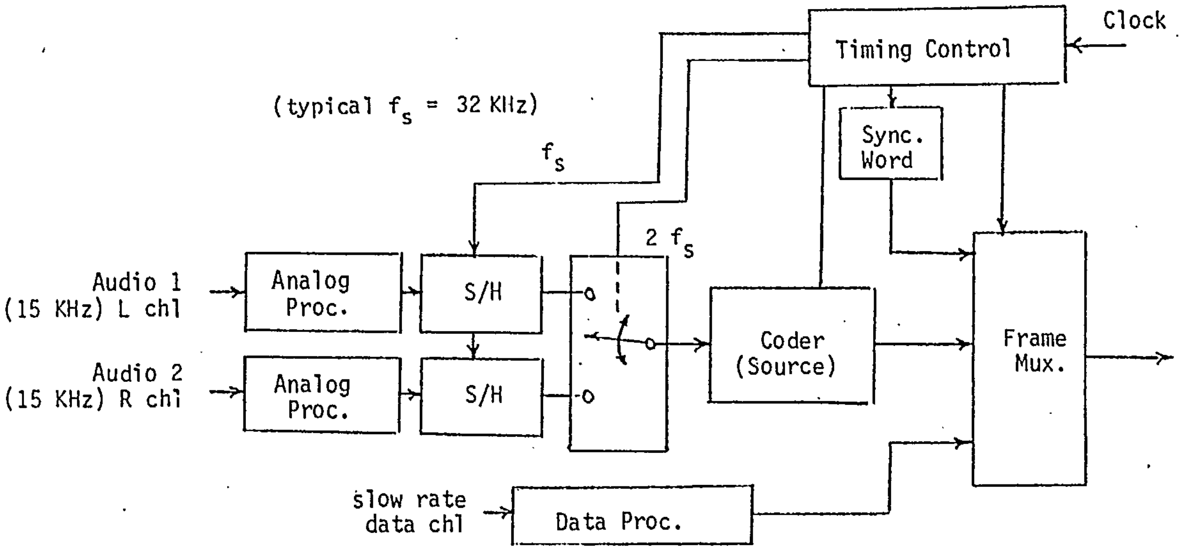
3.2.4 Shared Versus Dedicated Codecs

For the transmission of multiple audio channels (the left and right channels of a stereo program being the simplest case), the source codecs may be arranged in one of two configurations: shared or dedicated.

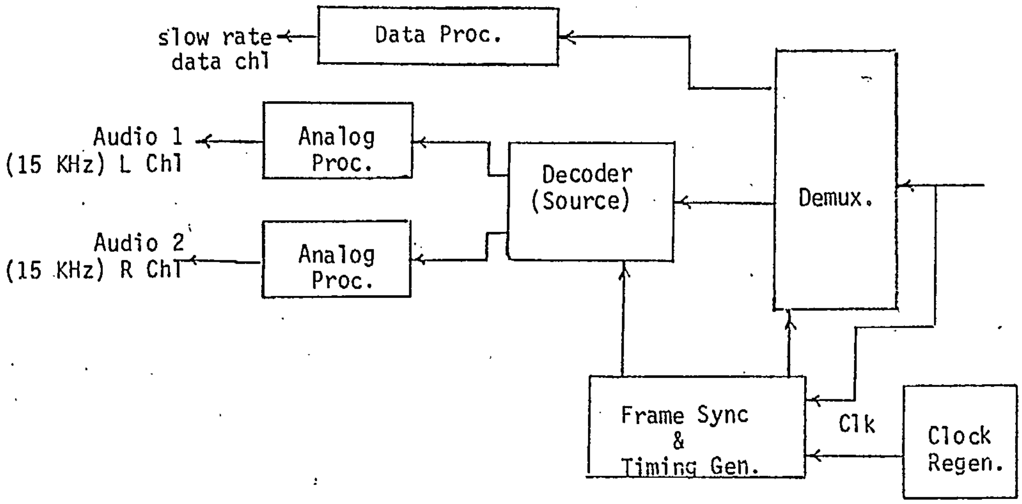
Figure 3.2 illustrates the possibility of sharing the digital portion of the encoder between the left and right channels of a stereo pair. Although the analog processing and sampling processes are implemented individually for each channel, the resulting samples from both channels are digitized sequentially by a single encoder and the corresponding code words are output alternately. The incoming low-rate data channel is then interleaved with this signal. The timing control circuitry controls these operations and also generates sync words to head the frame. At the decoder, the complementary operations are performed, with the necessary timing control derived from the frame synchronization. This technique is ideally suited for word-interleaving multiplexing, although with the addition of buffers the bit interleaving approach can also be accommodated.

Figure 3.3 illustrates the dedicated codec approach to multiplexing/demultiplexing. This approach is more straightforward and simple to implement in the sense that a separate codec is provided for each audio channel. The overall frame format can be generated in either a bit-interleaved or word-interleaved basis with the use of additional buffers.

An important practical consideration in the selection of a codec implementation scheme is that the source codecs currently manufactured contain both the analog and digital processing on a single card. This factor provides a strong incentive to make use of the dedicated codec approach, since separation of the analog and digital operations would require extensive modifications to the equipment. It is likely that purchasing separate complete codecs for each audio channel would be less expensive than the cost of either modifying existing codecs or developing the necessary processing ourselves.

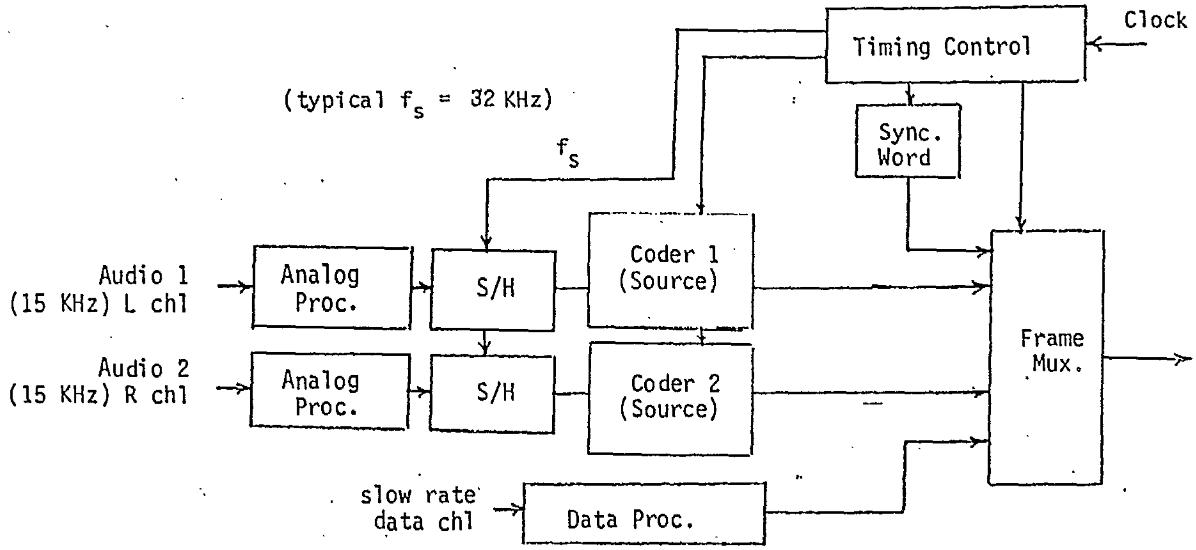


(a) 'Shared' codec type multiplexer

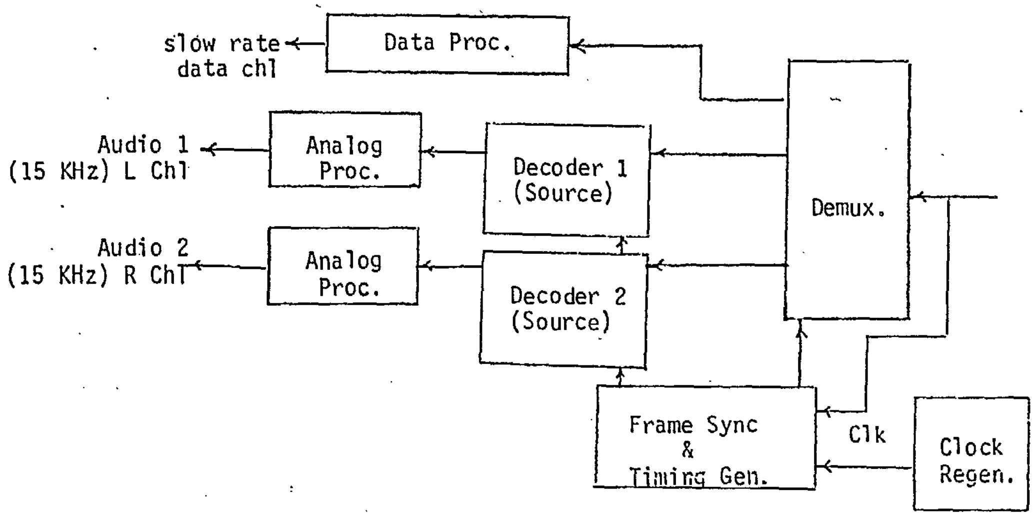


(b) 'Shared' codec type demultiplexer

Figure 3.2 'Shared' Codec Type Mux/Demux Implementation



(a) 'Dedicated' codec type multiplexer



(b) 'Dedicated' codec type demultiplexer

Figure 3.3- 'Dedicated' Codec Type Mux/Demux Implementation

3.3 Basic Audio Multiplexing Requirements for SCPC Transmission

3.3.1 General Considerations

The basic requirements for the audio distribution multiplexer are that it be able to accommodate the following traffic:

(i) one 15 kHz stereophonic audio program (comprising a 384 kb/s signal for each of the left and right channels), and

(ii) a low-rate (32 kb/s) data channel.

Figure 3.4 illustrates the general multiplexer configuration when the audio source is collocated with the transmit earth station. The two audio channels which are digitally processed by separate source encoders, are fed into the multiplexer along with the low-rate data channel and any synchronization information. The multiplexer formats these signals and the resulting 800 kb/s signal (plus any sync overhead) is then passed to the satellite transmit equipment (which includes the forward error correction coding). At the receive earth station, the de-multiplexer performs the inverse operation.

In the event that the transmit earth station is not collocated with the production studio (a scenario which is quite likely), a backhaul link between the two locations will be required. Figure 3.5 shows the case when a T1 link is used to interconnect the production studio and the transmit site. For this scenario, a T1 interface card is required at each location.

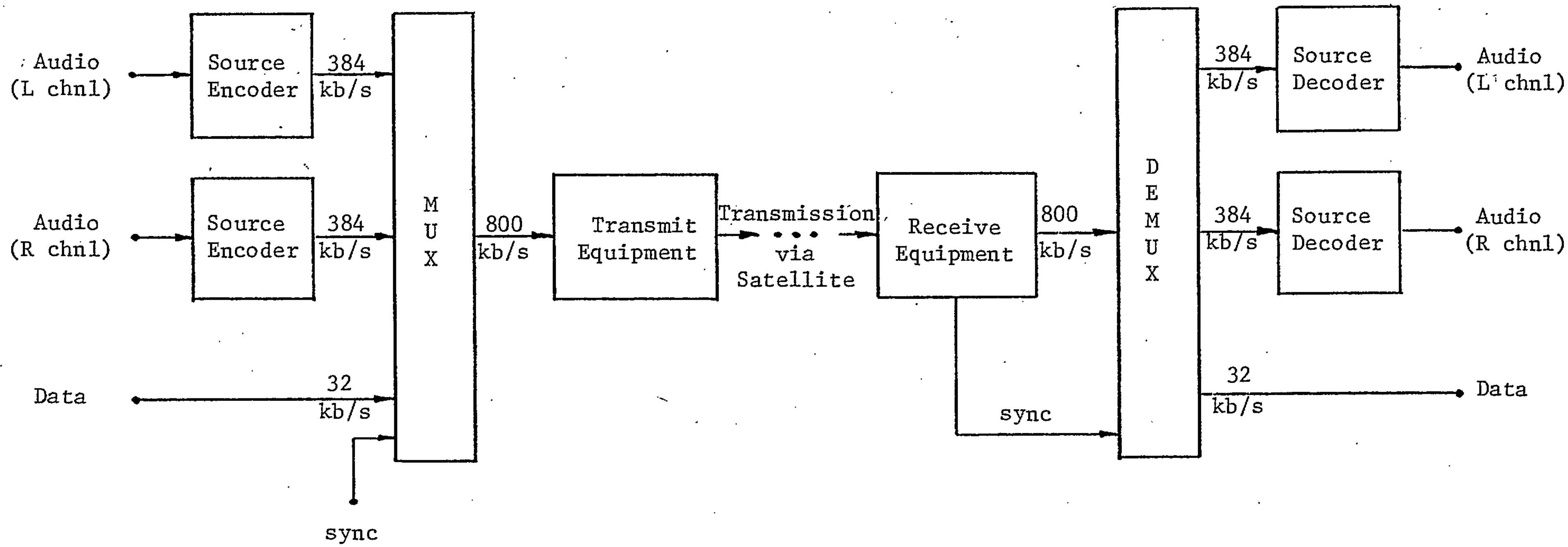


Figure 3.4 - General Multiplexer Configuration

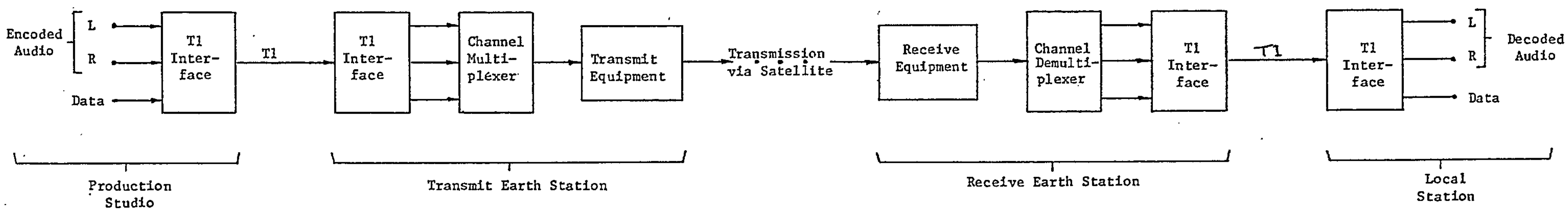


Figure 3.5 - System Configuration with T1 Transmission

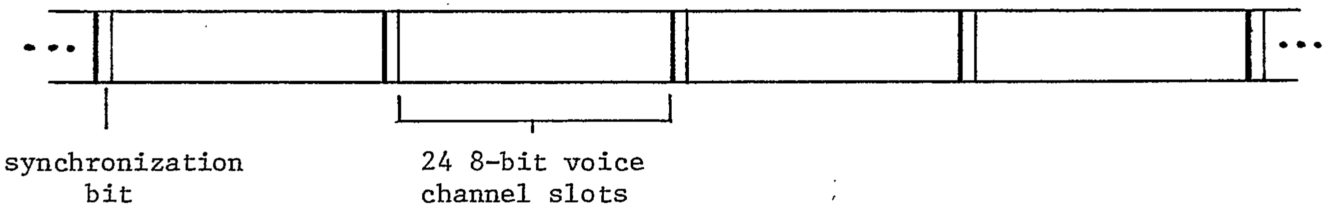
The intent, with respect to this system, is to design a multiplexer unit that is capable of operating in either of these two scenarios. This implies that either the T1 receive interface must be such that it presents an input to the multiplexer identical to that provided by the production studio, or that some relatively simple modification be made to the basic multiplexer that permits it to take advantage of some characteristic of the signals from the T1 interface.

This will be discussed further in Section 3.3.3 when the transmission frame structure is considered. However, before this can be done, the format of the T1 channel should be examined.

3.3.2 T1 Transmission

The format of the T1 transmission channel operating at 1.544 Mb/s (as specified in CCITT Recommendation G.733 [18]) is depicted in Figure 3.6(a). The basic frame size of 193 bits consists of 24 8-bit slots, each accommodating one sample of a separate 64 kb/s PCM voice channel, and 1 bit for frame synchronization. Thus, the 24 multiplexed voice channels in conjunction with the frame repetition rate of 8 kHz yields the total bit rate of 1.544 Mb/s.

In order to transmit a 384 kb/s audio channel on a T1 carrier, the AT&T Technical Advisory No. 74 [7] recommends that 4 successive 12-bit audio samples be grouped together to occupy 6 adjacent voice channel slots. This transmission scheme is demonstrated in part (b) of Figure 3.6. For example, to transmit a stereo program, one channel would be transmitted in the first 6 channel slots of each frame and the other channel would occupy the next 6 channel slots of each frame. Both the Tau-tron and Bayly audio codecs are designed to interface to a T1 carrier in



a) T1 Frame Format

L= 4 12-bit samples of left channel of stereo audio program

R= 4 12-bit samples of right channel of stereo audio program

b) Audio Transmission Format

Figure 3.6 - T1 Transmission of Audio Programs

this fashion. In fact, Tau-tron specifies that the time slots for the codec output may be manually selected as either 1-6, 7-12, 13-18 or 19-24. Any data transmission requirements would utilize the vacant channel slots.

Up until this point in time, Telecom Canada have not offered a T1 service directly to the customer (i.e. it has only been an internal network transmission standard). However, a proposed service known as Megaroute [19] is intended to provide a T1 capacity to the customer and is planned for introduction in 13 major cities across Canada later this year (for intra-city applications only). With this service, the customer must lease the entire T1 line. However, the cost estimates provided indicated that the service will be cheaper than leasing multiple 56 kb/s data channels if the data rate requirement is greater than 350 kb/s. Also, the additional multiplexing complexities associated with the multiple data channel scenario will be eliminated.

3.3.3 Frame Format

In order that the information transmitted via satellite can be properly recovered at the receiver, the audio and data samples must be multiplexed into a frame format. There are three possible approaches to the organization of the transmit frame:

(i) a frame consisting of one digital audio word for each channel, a data component and a unique word,

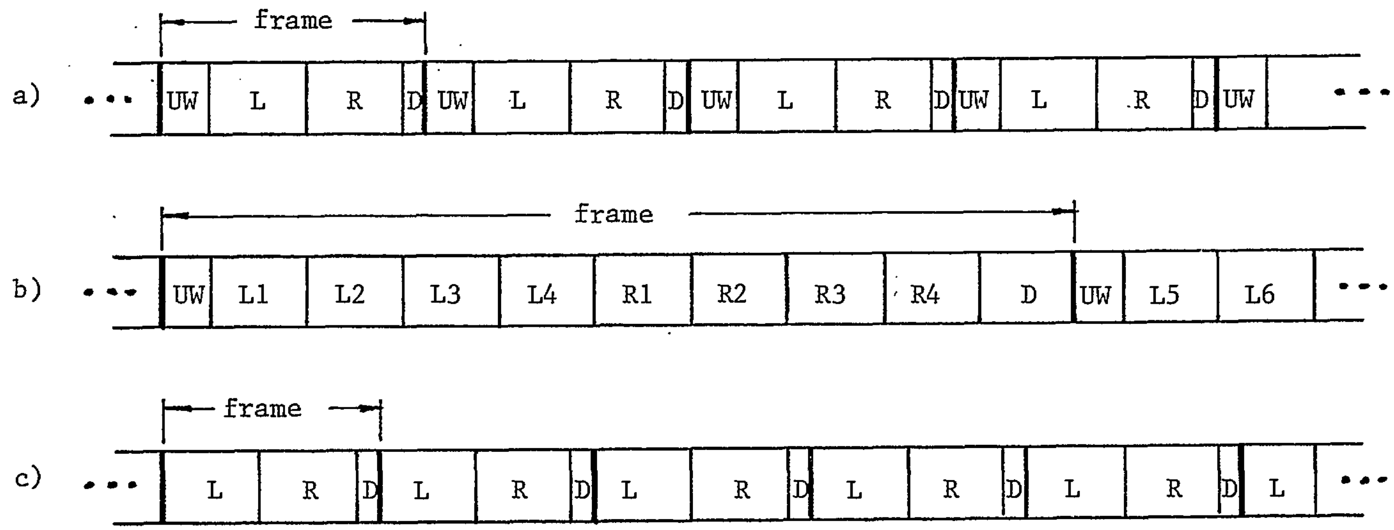
(ii) a frame consisting of multiple digital audio words for each channel, a data component, and a unique word, or

(iii) a frame consisting of only one digital audio word for each channel and a data component (i.e. no unique word - synchronization is determined by the ability to decode the forward error correction coding).

These three frame formats are shown in Figure 3.7. Although method (i) is the most straightforward approach, it will be relatively inefficient in terms of frame overhead (because of the small overall size of the frame). To improve frame efficiency, it may be desirable to group several audio code words together to form a frame. In this respect, it may be advantageous to make such a frame format compatible with the T1 format (i.e. 4 successive audio samples, as is indicated in part (b) of Figure 3.2). With this approach, the synchronization information for the multiplexer would be derived directly from the T1 receive interface.

An alternative to the standard use of a unique word for synchronization purposes is to place all the responsibility for synchronization on the receiver's ability to decode the forward error correcton coding properly. Although this method (shown in Figure 3.7c) eliminates overhead, it is much more complex to implement and will likely have a higher probability of losing synchronization.

In general, the most likely candidate for the frame format is the method in which 4 12-bit audio samples from each channel are included with a 4-bit data component and a unique word for synchronization. However, the tradeoffs associated with such parameters as frame acquisition time, false lock probability and loss of sync probability will have to be evaluated for all three methods before a frame format is defined.



UW= synchronization unique word

L= 4 12-bit samples of left channel of stereo audio program

R= 4 12-bit samples of right channel of stereo audio program

D= data component

Figure 3.7 - Possible Frame Formats

3.4 Implementation Options

In this section, the implementation options required to build some flexibility into the system are discussed.

3.4.1 7.5 kHz Channel Operation

One of the most basic implementation options is to be able to replace a single 384 kb/s 15 kHz audio channel with two 192 kb/s 7.5 kHz audio channels. Figure 3.8 shows the case when each channel of a 15 kHz stereo audio program is replaced by two 7.5 kHz channels. Since these two program sources are represented by data rates that differ by exactly a factor of two, the design of the multiplexer required to handle both is greatly simplified. CCITT Recommendation J.42 [20] specifies that the audio samples from the two 7.5 kHz channels should be alternated, with the first 12-bit sample representing channel no. 1 and the second 12-bit sample representing channel no. 2.

The most important aspect of multiplexing the two different channel formats together is that the receiver be able to distinguish between the different programs. This may be done either on a prearranged basis or by perhaps using a different unique word to identify the 7.5 kHz program channel.

3.4.2 Multiple Stereo Channels

It may be a requirement that more than one stereo audio channel be transmitted from a production studio to the transmit earth station. Although each channel would be transmitted via satellite separately in an SCPC mode, two such stereo channels could be accommodated on the T1 backhaul link. This scenario would make use of all 24

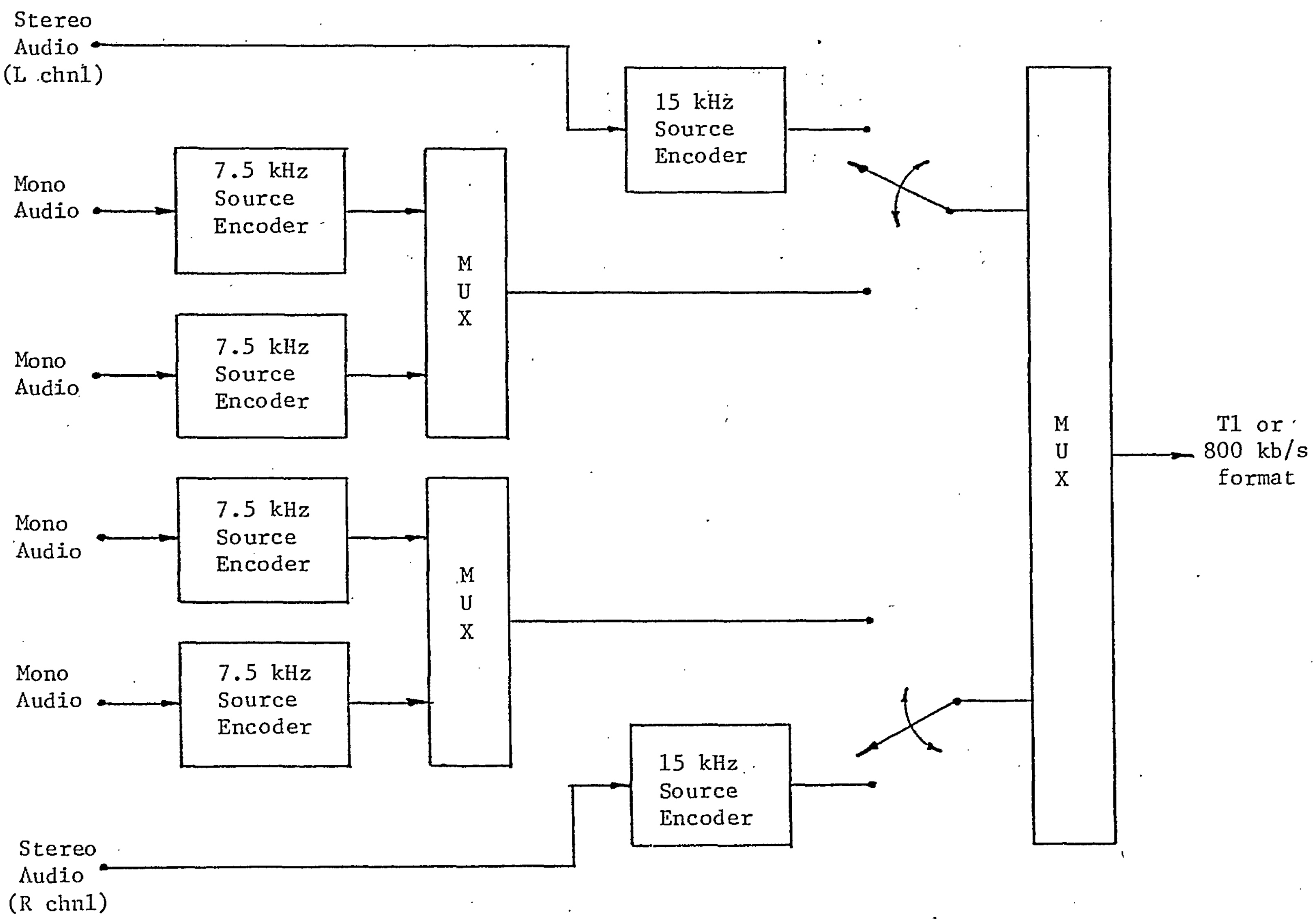


Figure 3.8 - Possible Configuration for 15/7.5 kHz Programs

voice channel slots on the T1 carrier, thus leaving only the first bit of every second frame (4 kb/s) to be used for data transmission. Hence, provided that this reduction in the data handling capability is acceptable, this represents a valid multiplexing scenario.

3.4.3 Compatability with Other Encoding Schemes

Ideally, it would be desirable to design a multiplexer that would be capable of accepting the output from any source encoder and transferring it into a common transmission format. In Section 2.7.4, it was shown that the source coding scheme is compatible with the Scientific Atlanta equipment used in the U.S., provided that the required bit-mapping operation is carried out. It would also be desirable that the multiplexer should be able to accommodate the output from the Dolby encoder. Since this source coding scheme results in a bit rate somewhat less than 384 kb/s, it should be possible to map the Dolby bit stream into the selected frame format with a minimal amount of modifications to the multiplexer.

4.0 TASK 5: MODULATION

The choice of modulation technique for digital SCPC radio distribution terminals is affected by may factors. Among these factors are:

- available C/N_o
- required E_b/N_o
- system phase noise
- interference
- synchronization errors
- use of FEC coding, and
- required BER

The primary issues relate to power efficiency, interference immunity, sensitivity to synchronization errors and implementation aspects. The candidate modulation techniques which can be considered for the SCPC satellite radio program distribution system are:

CBPSK	- coherent binary PSK
DBPSK	- differentially detected binary PSK
CQPSK	- coherent quaternary PSK
DQPSK	- differentially detected qPSK
OQPSK	- offset QPSK
CMSK	- coherent minimum shift keying
DMSK	- differentially detected MSK
Duo-MSK	- Duobinary encoded MSK
TFM	- tamed frequency modulation or double duo-binary encoded MSK.

Speaking in relative terms:

- the binary (CBPSK etc.) schemes are <u>power efficient</u>,

- the correlative schemes (Duo MSK, 1TFM) with constant amplitude are highly bandwidth efficient,

- the quartenary schemes (QPSK, OQPSK, MSK etc.) have power band bandwidth efficiencies in between the binary and correlative schemes,

- the differential detection schemes (DBPSK, DQPSK, dMSK etc.) can be considered because of their implementation simplicity. However, the performance of these schemes are slightly inferior compared to their respective coherent detection versions.

The correlative modulation schemes will occupy relatively narrower bandwidth because of the correlative coding and also will have least spectral spread due to channel nonlinearity because of the constant envelop property. However these schemes will require relativley more E_b/N_o for a given bit error rate e.g. TFM will require about 1 dB more E_b/N_o as compared to QPSK at a BER of 10^{-4}. Further, the modulator-demodulator implementation is more complex than QPSK. For the radio program application the requirements on the power efficiency and the implementation simplicity are more important than the bandwidth efficiency and the immunity to channel nonlinearity. Therefore these schemes are not given further consideration.

The differential detection schemes are also not considered further because of their relatively higher E_b/N_o requirements as compared to the coherent detection schemes. Therefore the remaining coherent schemes which require further consideration are BPSK, QPSK, OQPSK and MSK.

4.1 Comparison of Modulation Techniques

The performance of the above modulation schemes are compared under various filtering and transmission channel conditions in [21-24].

In the rest of this section a brief comparison is made between these techniques with respect to radio program distribution application.

4.1.1 Bandwidth Efficiency

The unfiltered main lobe spectrum of BPSK is twice wider and that of MSK is 1.5 wider as compared to (O)QPSK. However under practical non-ideal filtering conditions, the bandwidth efficiencies may differ. Table 4.1 shows the BT product required to achieve 1 dB E_b/N_o degradation at a BER of 10^{-5} for linear (12 dB backoff) and nonlinear (1 dB backoff) channels.

4.1.2 Power Efficiency

The power efficiency of a modulation scheme is determined by the minimum E_b/N_o required to achieve a specified bit error rate. Under ideal channel and filtering conditions all the above modulation schemes perform the same. Under non-ideal channel conditions the performance of these schemes depend on the type of filtering, adjacent channel spacing etc. Figures 4.1 and 4.2 depict the SNR degradtion at a BER of 10^{-4} for different modulation schemes and for different filter bandwidths [22]. It must be noted that the filtering used is 4-pole 0.5 dB ripple chebychev characteristic both at the transmitter and receiver.

Table 4.1: BT Product for 1 dB E_b/N_o degradation at BER = 10^{-5}. Chebychev transmit and receive filters are assumed.

	BPSK	QPSK	OQPSK	MSK
Linear Channel (12 dB BO)	1.9	1.0	1.5	1.2
Nonlinear Channel (1 dB BO)	2.1	1.0	1.4	1.1

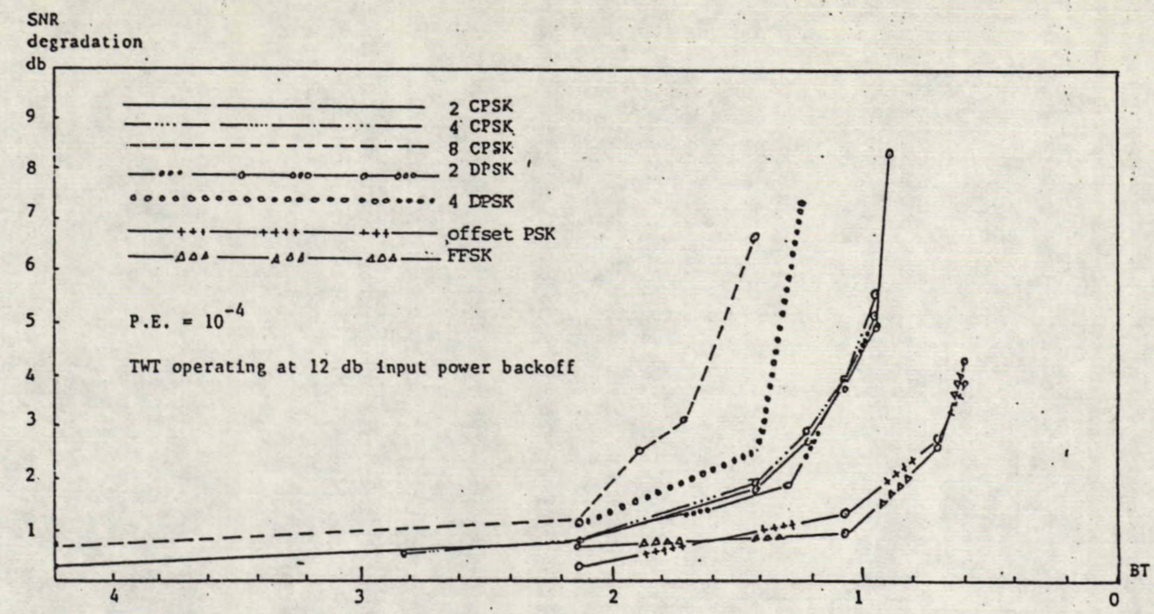


Figure 4.1 : SNR degration as a function of BT where B is bandwidth and T is baud length.

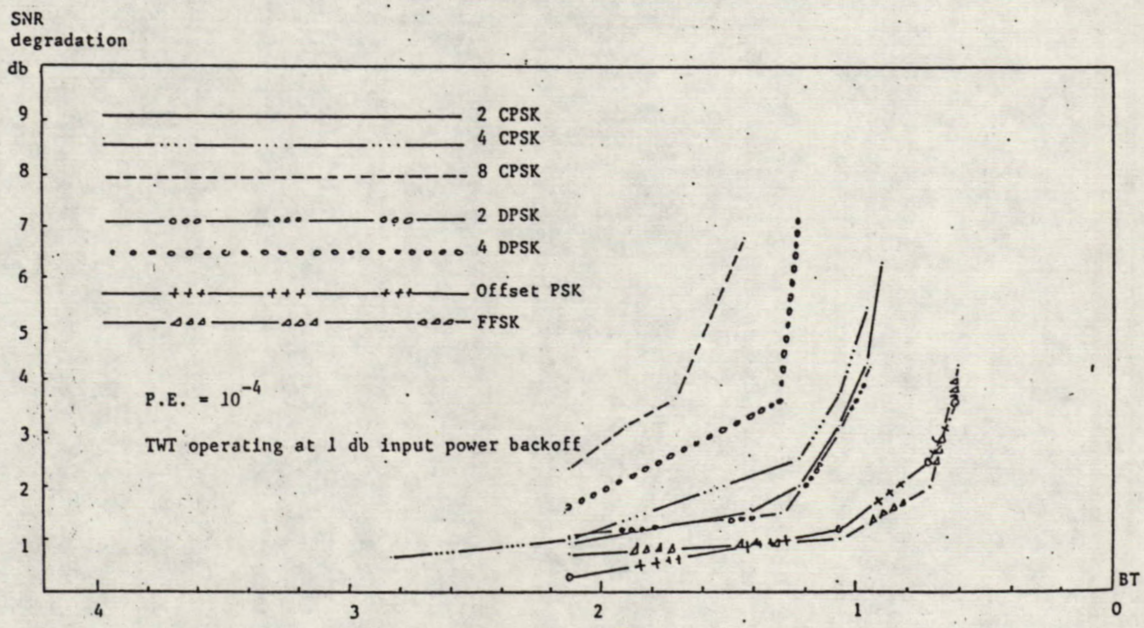


Figure 4.2 SNR degration as function of BT where B is bandwidth and T is baud length.

4.1.3 Effects of Nonlinearity

As the TWTA amplifier may be operated at several decibels backoff for multi-carrier operation to minimize the intermodulation interference. Thus the transponder can be assumed almost linear. The performance of various modulation schemes for different HPA and TWTA input backoff conditions is shown in Figure 4.3 [24].

4.1.4 Effects of Synchronization Errors

The stability of the carrier frquency and the purity of the recovered carrier and bit timing clock are important to the performance of a coherent system. The automatic frequency control (AFC), carrier recovery (CR) circuits and bit timing recovery (BTR) circuits are used to reduce the synchronization errors. The analysis of synchronization circuits and their effects on BER performance are relatively involved. Presented in Figure 4.4 is the power degradation versus carrier phase error.

4.1.5 Interference Effects

The interference to the digital audio distribution terminal can be from a number of sources, notably:

- uplink
- cross polarization
- adjacent satellite
- adjacent channel
- terrestrial
- intermodulation

The effects of this interference should be studied for specific cases depending on the amount of interference.

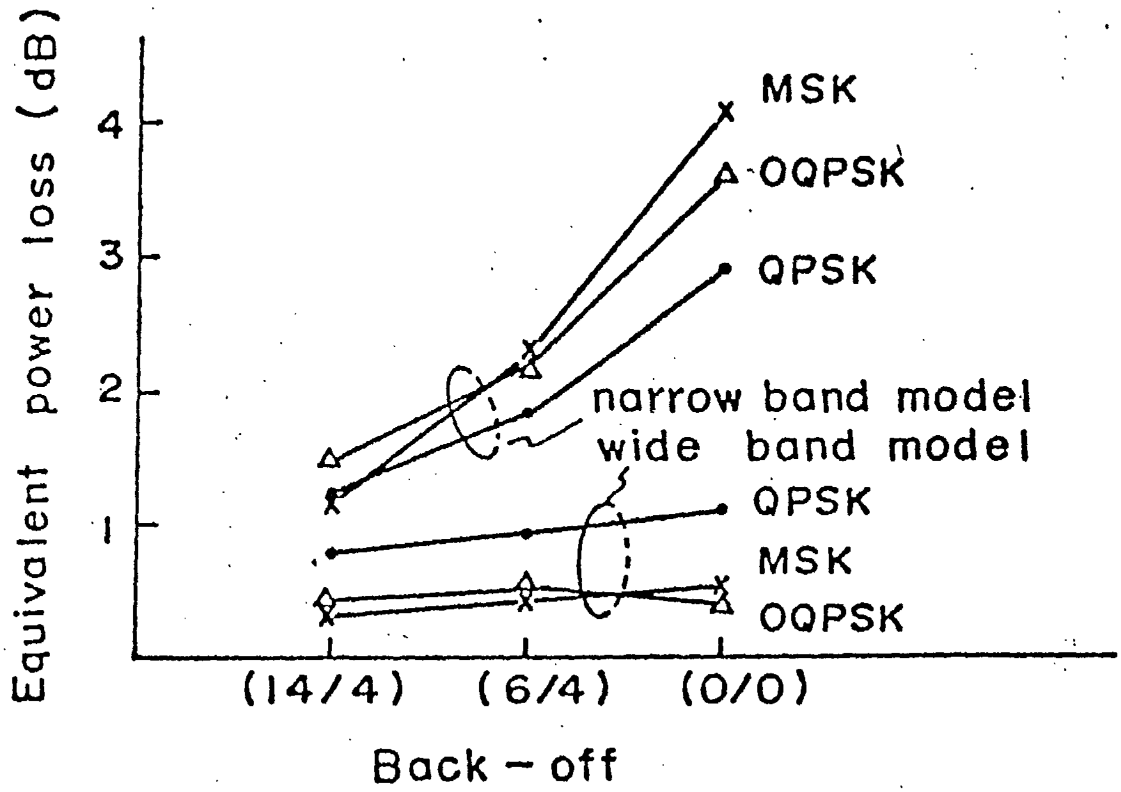


Fig.4.3 Equivalent power loss for various input back-off values for HPA and TWTA (BER: 10^{-4})

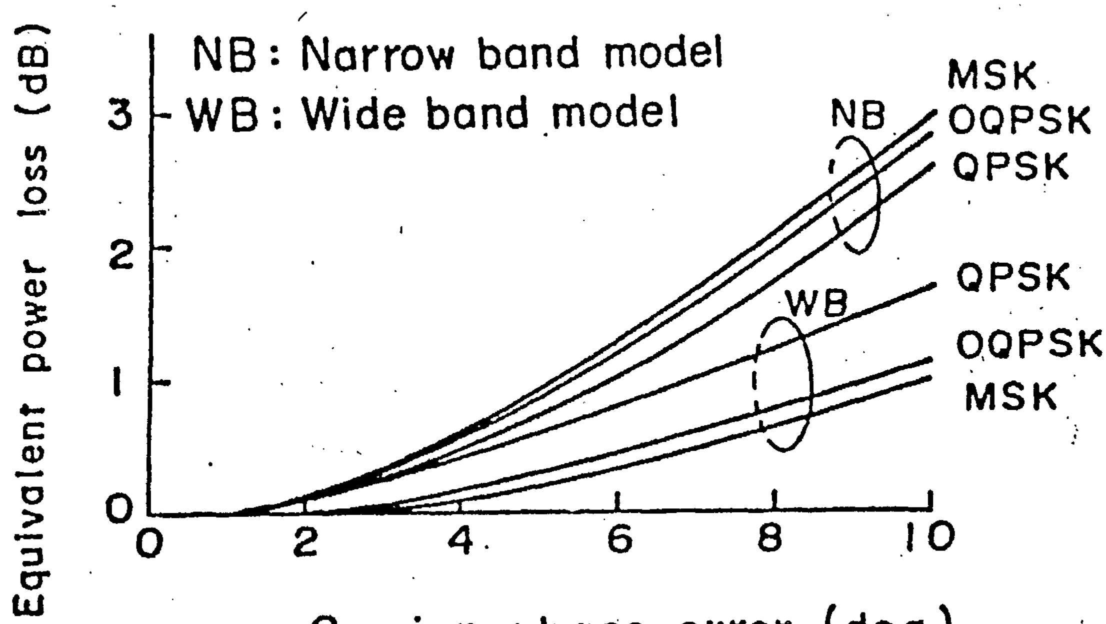


Fig.4.4 Equivalent power loss due to carrier phase error. (BER: 10^{-4})

For the dedicated transponder case, the adjacent channel interference level can be made negligible by using sharp rolloff filters and large spacing between adjacent channels. Optimization of frequency spacing and power has been done for a number of cases [25]. It was found that with an IF receive filter having an amplitude response resembling that of a 4th order Butterworth filter, and a channel spacing of 1.8 times the symbol rate, the adjacent channel interference level is about 40 dB below the wanted signal power. Interference due to intermodulation products can also be minimized or avoided in this case by staggering the radio carriers or by backing off the transponder.

MCS's previous study [25] suggests in the case of shared transponder FM-TV/SCPC that interference to the FMTV signal is not a severe problem. The interference to the radio SCPC carriers is significant because these carriers are of low power. Certain restrictions may be imposed on the SCPC carriers in order to avoid excessive interference to them:

- all SCPC carriers may be grouped to one side of the FMTV carrier to avoid comparably sized $2f_v - f_r$ IM products

- no SCPC carriers may be located less than 13 MHz away from the centre of the FMTV carrier to avoid direct interference from the FMTV signal

- to avoid interference from a cross-polarized staggered transponder, the FMTV signal in the staggered transponder may be located at least 9 MHz from any SCPC carriers in the wanted transponder.

4.2 Simulation Results

Computer simulation has been performed to quantify the performance of QPSK signals subject to different types of filtering and adjacent channel spacings.

4.2.1 Simulation Model

With multicarrier per transponder (MCPT) operation, the digital SCPC carrier is operated in a practically linear channel. The system model used for the computer simulation is shown in Figure 4.5.

The model shows a desired channel and two interfering adjacent channels on either side of the desired channel. The carrier phase and symbol timing are randomized between the wanted and interfering channels. The above computer simulation model is readily implemented by modifying MCS's in-house software 'SATSIM'. The simulation software is validated with the theoretical results the agreement can be seen in Table 4.2.

4.2.2 Filter Optimization

A number of transmit and receive filter combinations were simulated for different bandwidths B. These combinations are shown in Table 4.3.

The performance of QPSK with the above filter combinations are shown in Figure 4.6 - 4.9. It is interesting to note that at comparable bandwidths, the 6th order Butterworth filter performs within 0.5 dB of the raised cosine fitler.

4.2.3 Minimization of Channel Spacing

Two adjacent channels of the same type as the desired channel are considered to be placed on either side of the desired channel at a normalized spacing of d. The carrier phase, symbol timing between these channels are randomized. The spacing between these channels is varied and the degradation is quantified. The BER performance is plotted

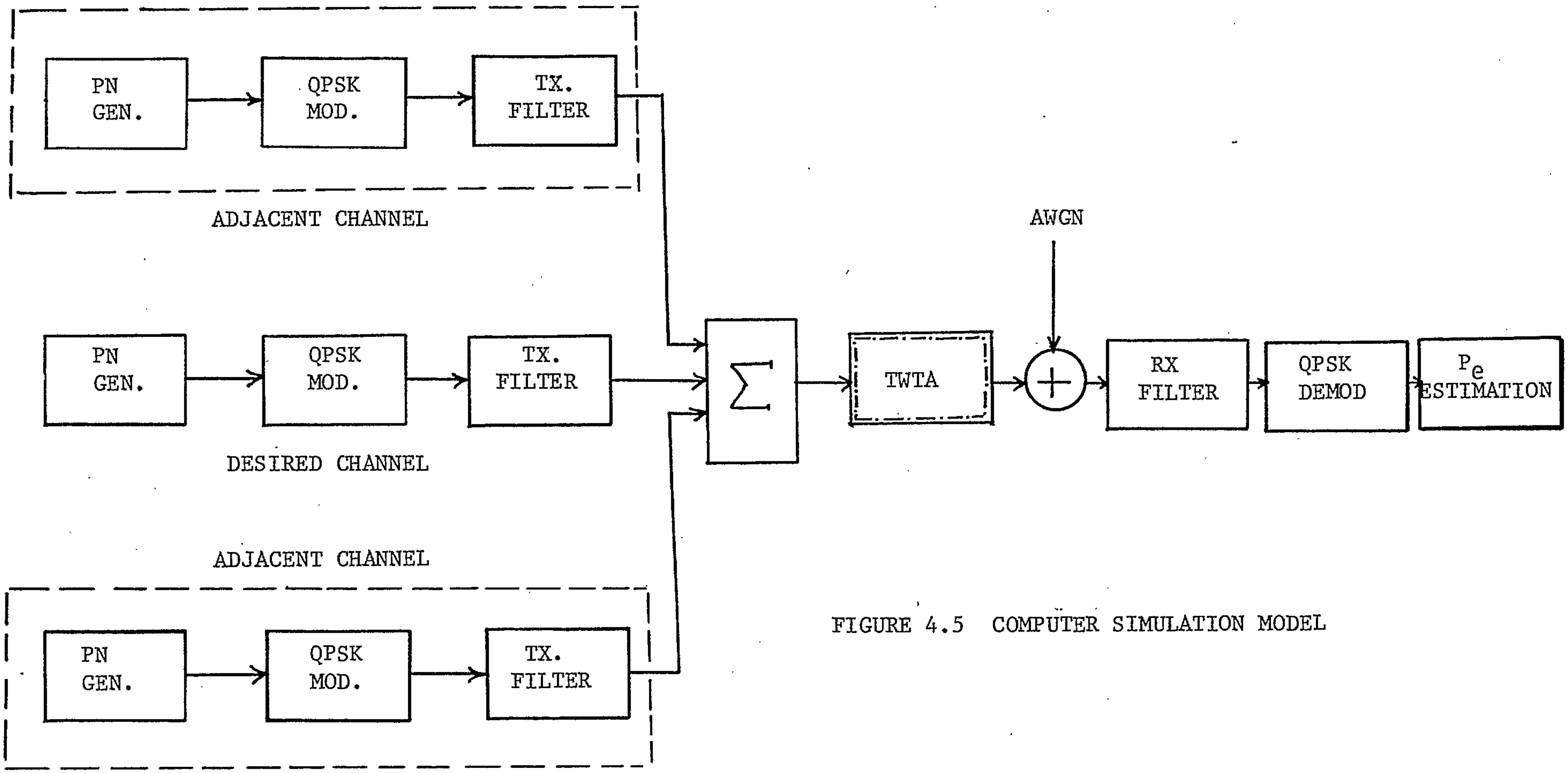


FIGURE 4.5 COMPUTER SIMULATION MODEL

E_b/N_o	Theoretical	SATSIM
7.0	1.54×10^{-3}	1.55×10^{-3}
7.6	6.92×10^{-4}	6.93×10^{-4}
8.4	1.99×10^{-4}	2.00×10^{-4}
9.586	2.00×10^{-5}	2.01×10^{-5}
10.529	2.00×10^{-6}	2.01×10^{-6}
11.3	2.00×10^{-7}	2.06×10^{-7}
11.97	2.00×10^{-8}	2.02×10^{-8}
13.06	2.00×10^{-10}	2.01×10^{-10}
13.936	2.00×10^{-12}	1.99×10^{-12}
14.988	2.00×10^{-15}	2.00×10^{-15}

Table 4.2: Validation of SATSIM Using Differentially Encoded Coherent QPSK.

Tx Filter	Rx Filter	Optimum BT_s at 10^{-5}
1. Square root raised cosine with x/sinx aperature equalization	Square root raised cosine	1.0
2. 2nd order Butter-worth (phase equalized)	Same as Tx filter	1.3
3. 4th order Butter-worth (phase equalized)	Same as Tx filter	1.2
4. 6th order Butter-worth (phase equalized)	Same as Tx filter	1.1

Table 4.3: Filter Combinations and Optimum BT_s for QPSK.

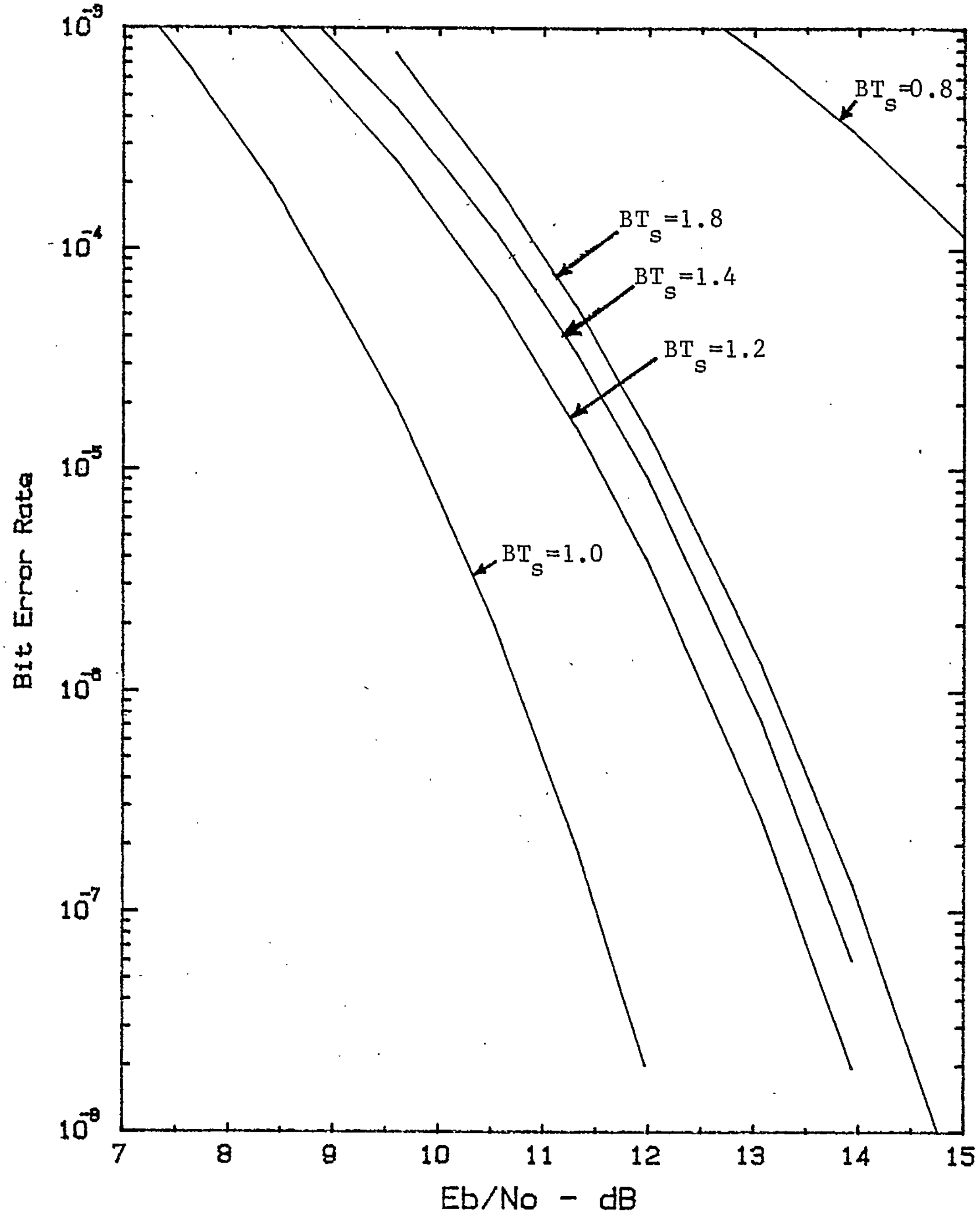


Figure 4.6 Performance of differentially encoded QPSK over channel with square root raised cosine transmit and receive filters (x/sinx aperture equalizer included at transmitter)

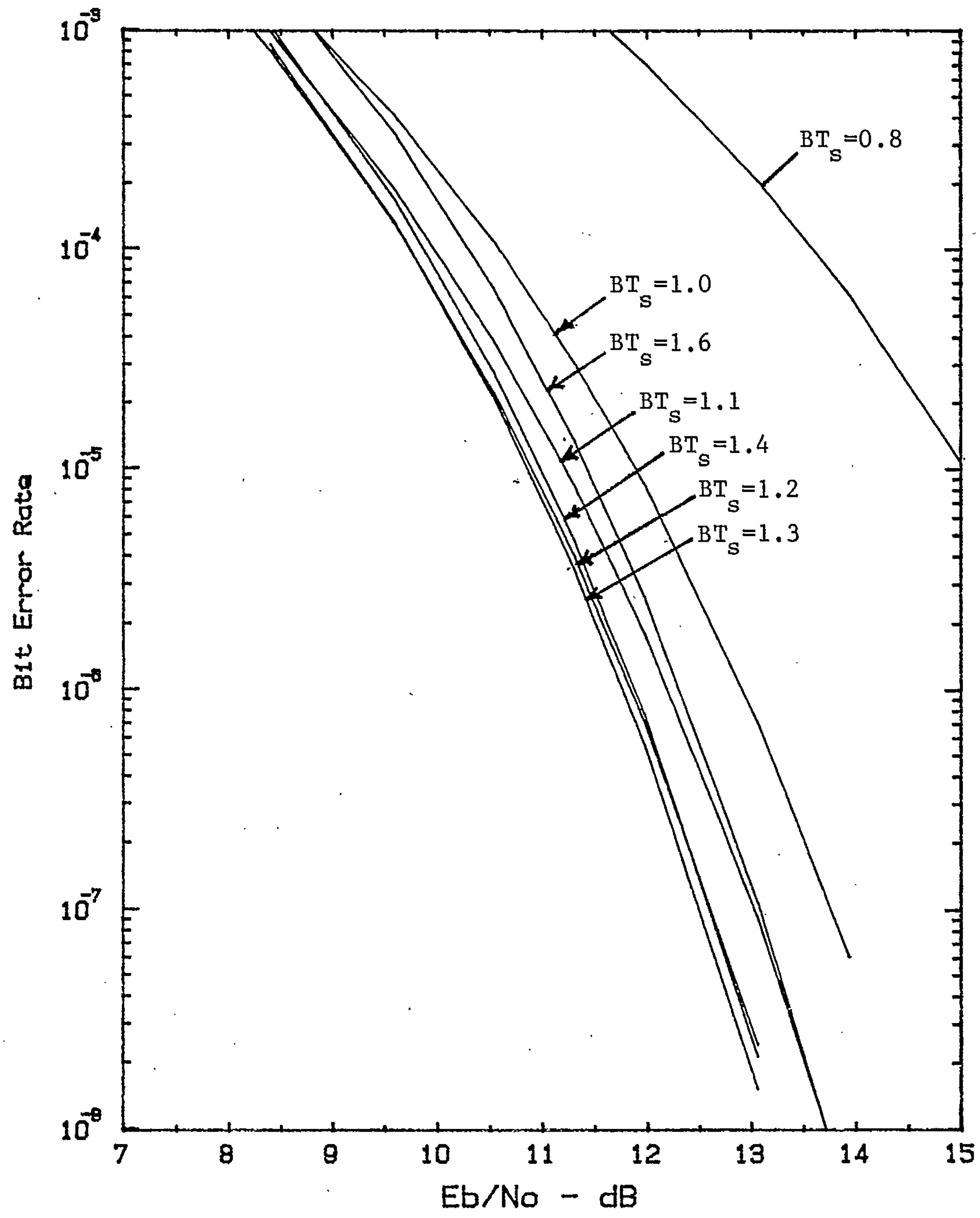


Figure 4.7 Performance of differentially encoded QPSK with 2nd order Butterworth transmit and receive filters (phase equalized)

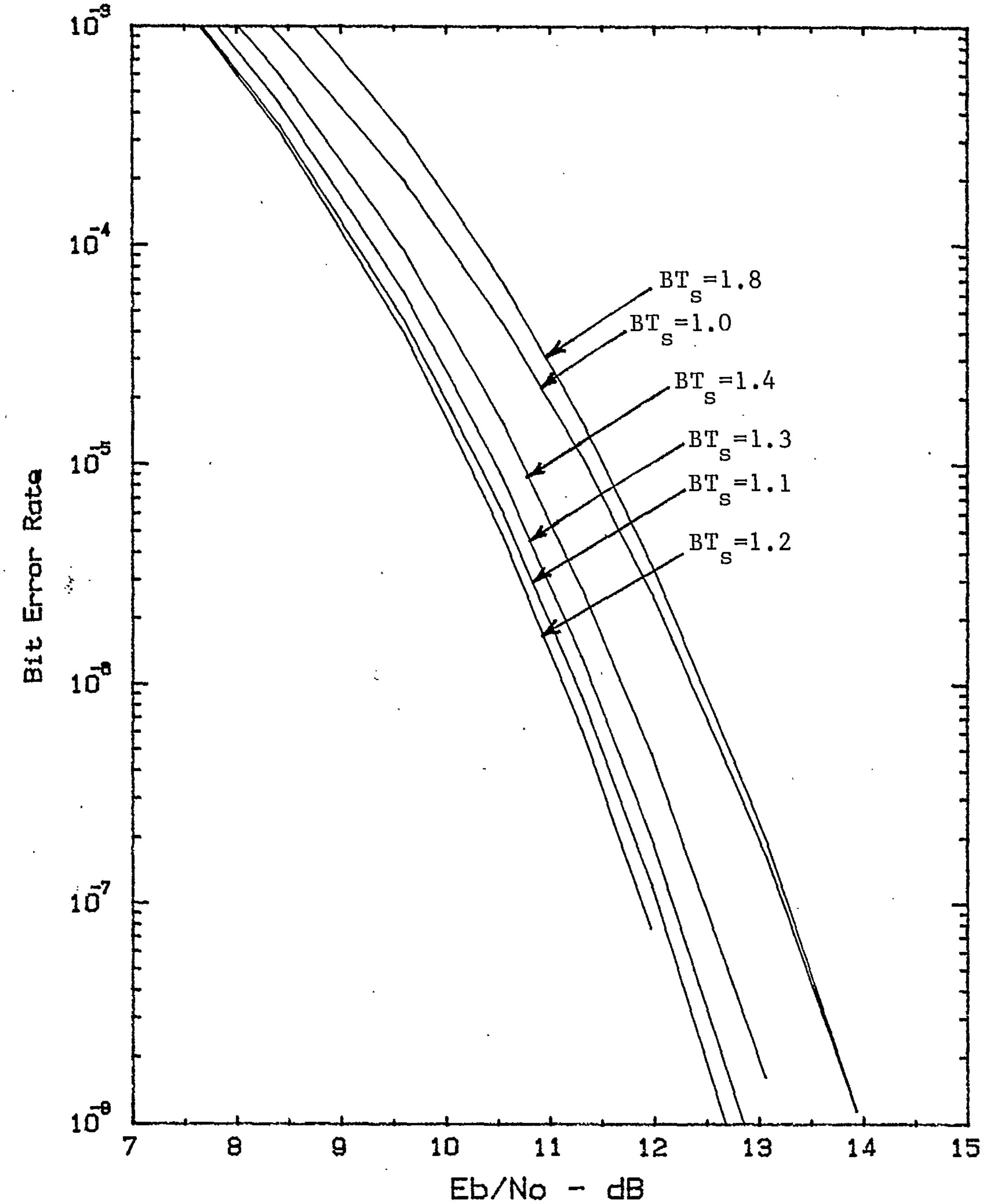


Figure 4.8 Performance of differentially encoded QPSK with 4th order Butterworth filters (phase equalized)

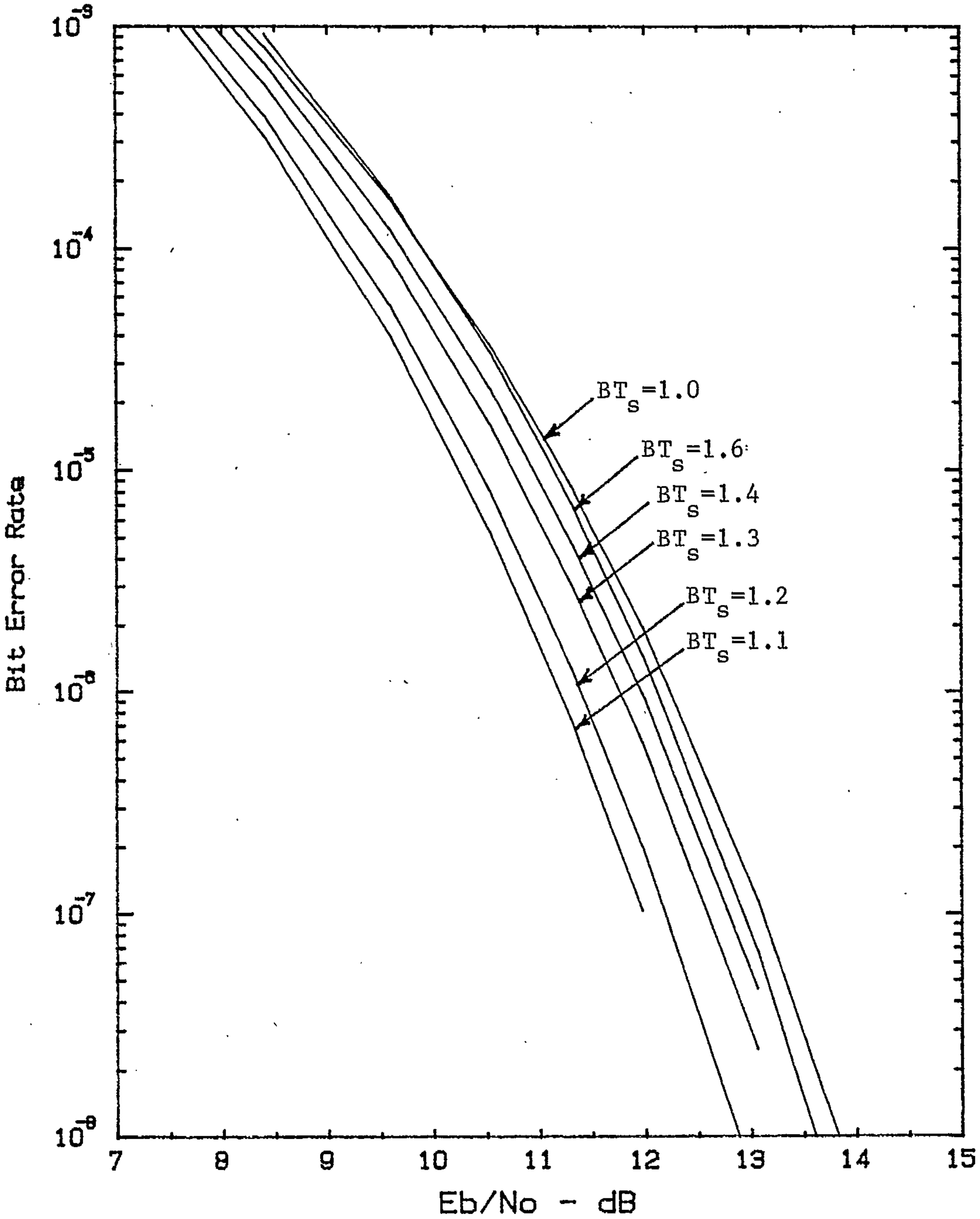


Figure 4.9 Performance of differentially encoded QPSK with 6th order Butterworth filters (phase equalized)

in Figures 4.10 - 4.13 for the filter combinations shown in Table 4.3. In each case both the desired and the interfering adjacent channels use the same type of filter with the optimum BT_s value.

4.3 Implementation of PSK Modems

One of the primary factors affecting the realization of a cost effective system is the implementation complexity. The implementation of the binary and the quarternary PSK modulation schemes are briefly described below:

CBPSK Modem Implementation

The functional block diagram of a coherent CBPSK modulator and democulator is shown in Figure 4.14. The modulator consists of a signal source, a data source whose signal is low-pass filtered, and a double balanced mixer. The data filter provides the bandlimiting, while the double-balanced mixer produces the equivalent of double sideband suppressed carrier modulation with the information primarily contained in the carrier phase. The demodulator is more complex and consists of a decision feedback phase lock loop, which controls the local coherent source in the form of a voltage-controlled crystal oscillator. In this configuration an in-phase and quadrature channel are demodulated and match-filtered with low-pass filters. A single channel is passed through a limiter and timing recovery circuit to reconstruct the diital data. The product of the digital data signal and the output of the other channel provide the phase error information for the tracking loop. This configuration has very good performance in a well behaved channel.

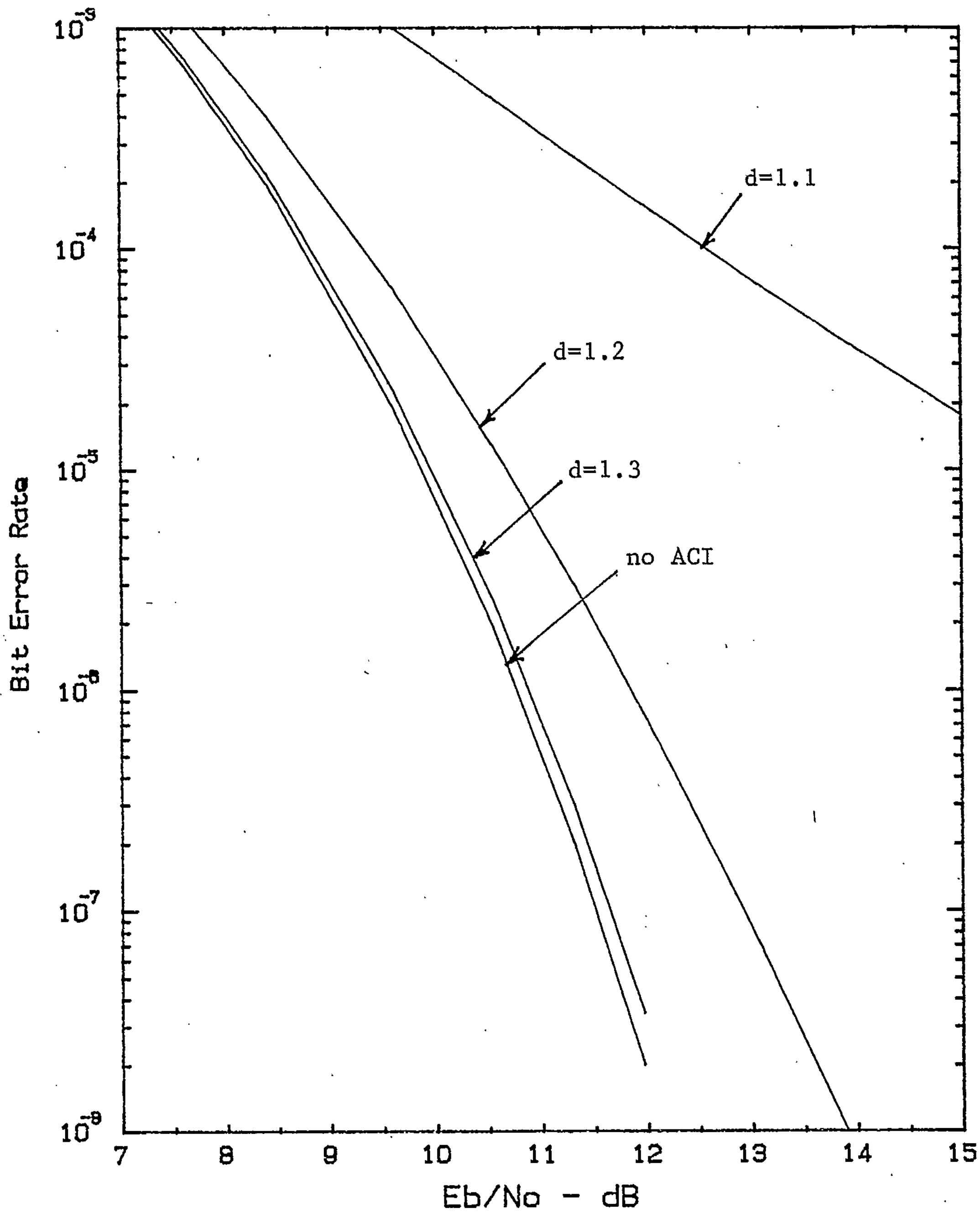


Figure 4.10 Performance of differentially encoded QPSK with adjacent channel interference (sqr. root raised cosine filters, BT_s=1.0)

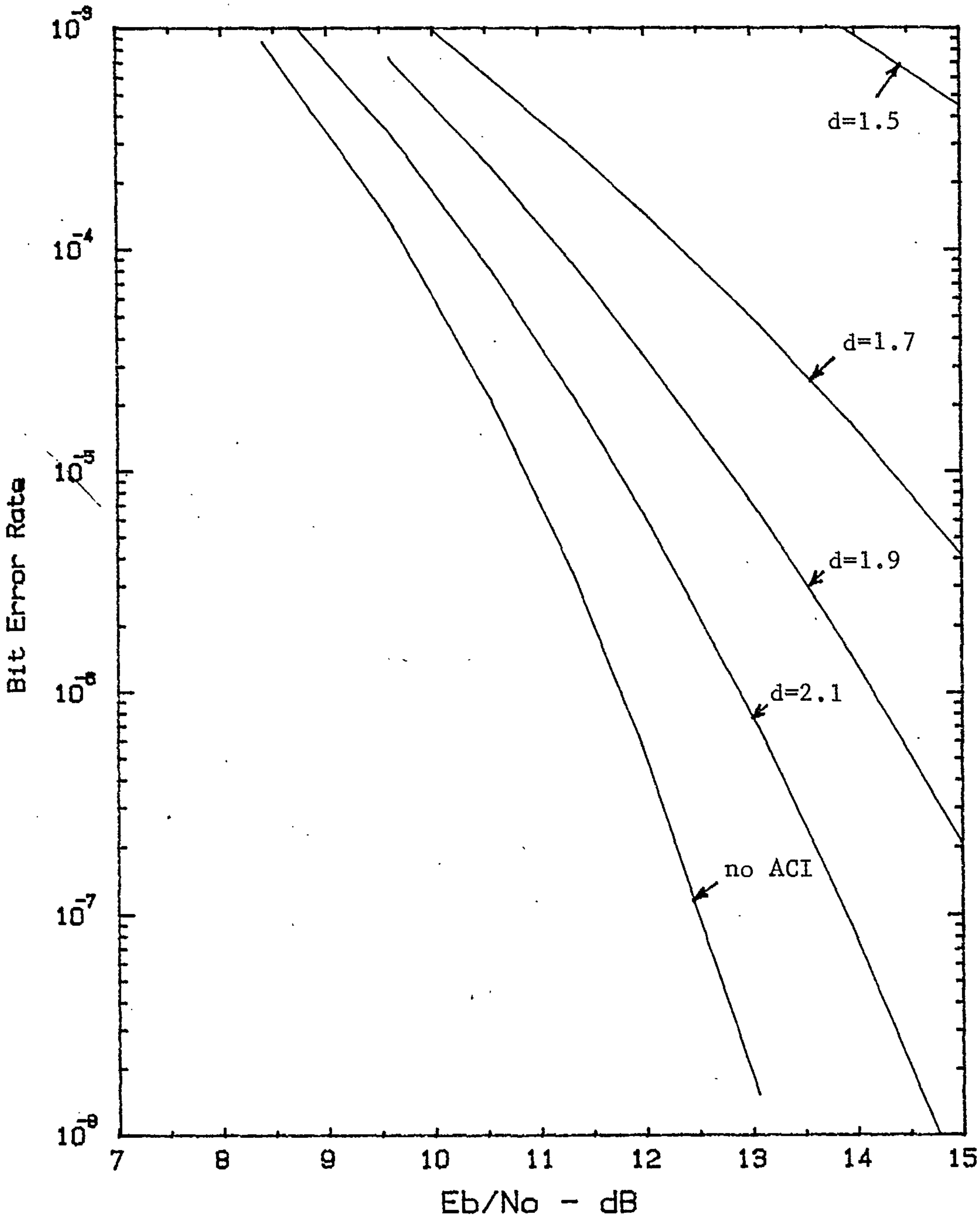


Figure 4.11 Performance of differentially encoded QPSK with adjacent channel interference (2nd order Butterworth filters, $BT_s=1.3$)

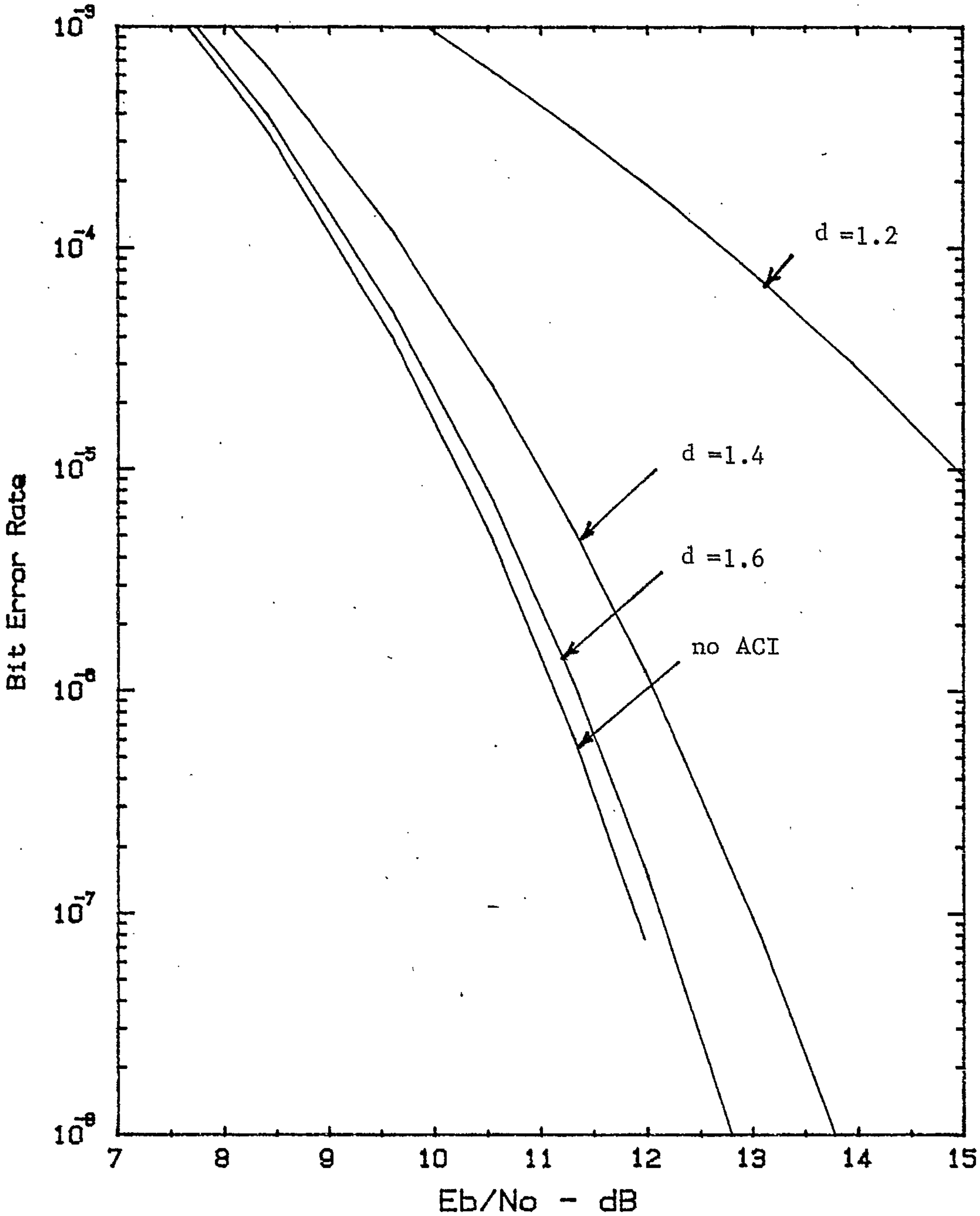


Figure 4.12 Performance of differentially encoded QPSK with adjacent channel interference (4th order Butterworth filters, $BT_s=1.2$)

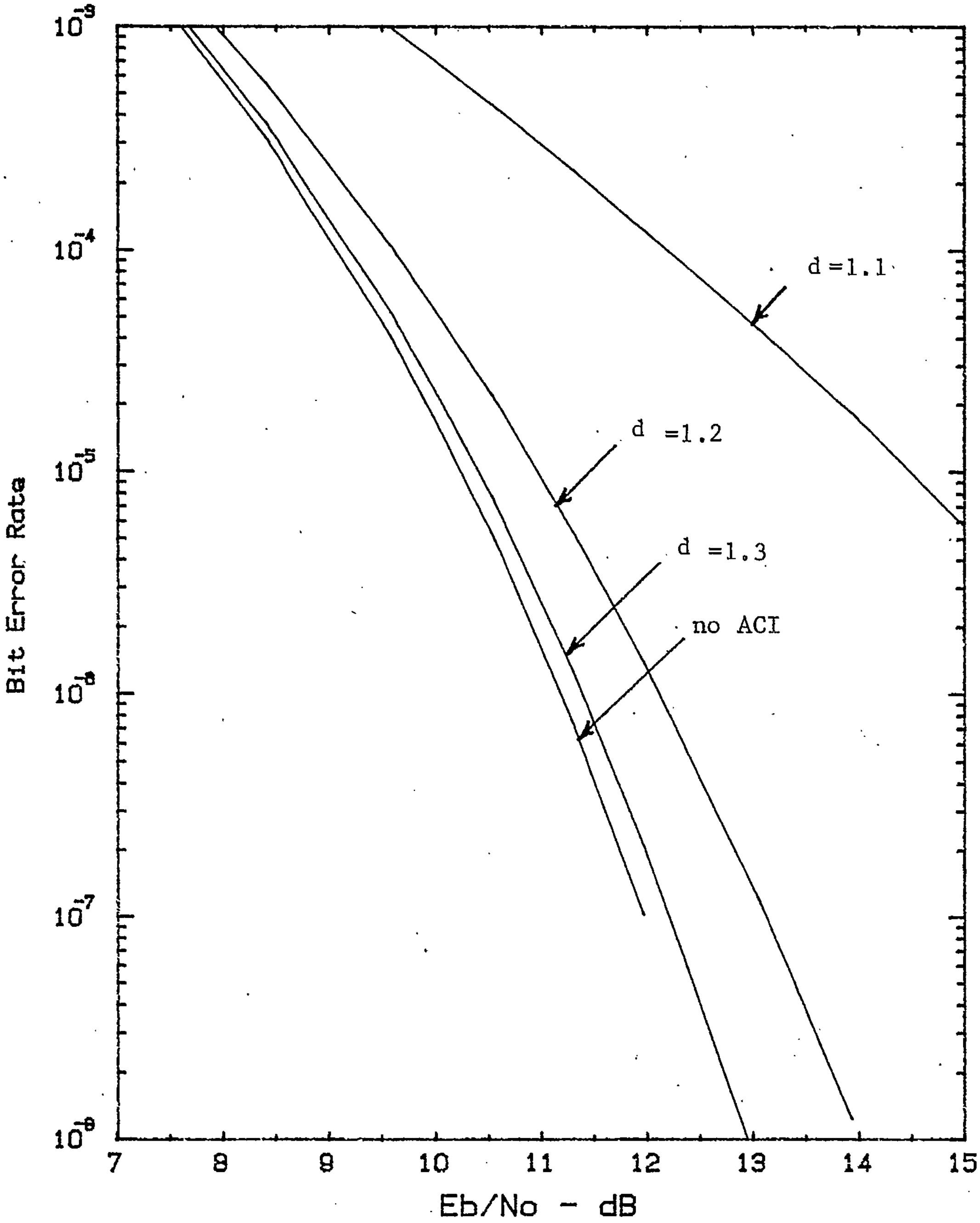
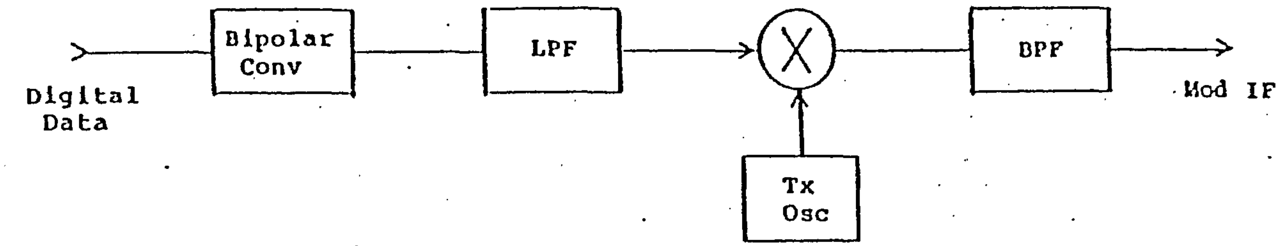


Figure 4.13 Performance of differentially encoded QPSK with adjacent channel interference (6th order Butterworth filters, BT_s=1.1)

BPSK Modulator



BPSK Demodulator

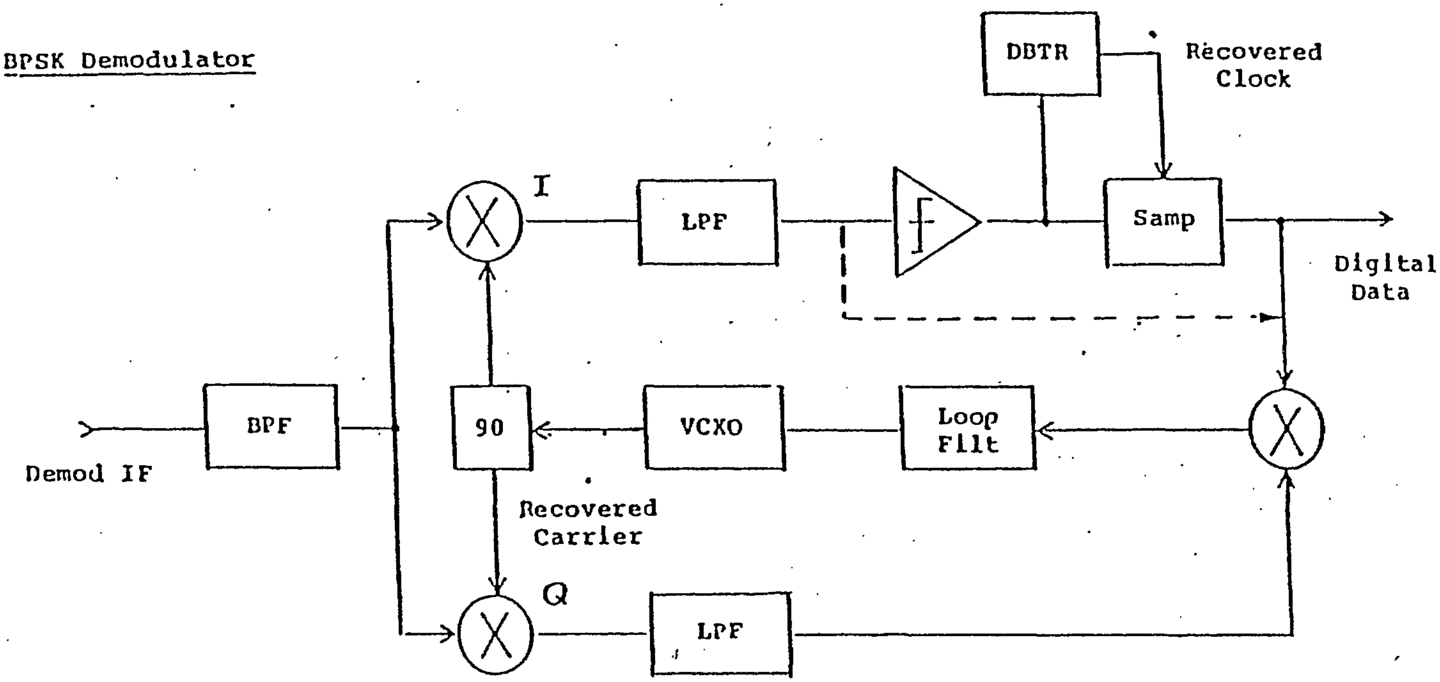


Figure 4.14 BPSK Modulator and Demodulator

QPSK Modem Implementation

A block diagram of a QPSK modulator is presented in Fig 4.15. The FEC encoded data is serial-to-parallel converted to form two data streams. These two data streams are low pass filtered using appropriate filter. The filtered waveforms are multiplied with quadrature carriers using double balanced mixers. The quadrature components are summed and passed through a variable pad which controls the output signal level. The carrier frequency can be selected using a crystal oscillator or a programmable frequency synthesizer. A dmoudulator block diagram is presented in Fig. 4.16. The received IF frequency carrier is mixed with a VCO output which is driven by a phase error voltage generated by the recovered carrier with reference to a stable crystal oscillator. An AGC controls the carrier level. The demodulated signal is low pass filtered, sampled and estimated. The quadrature bit streams are converted into a single bit stream and then fed to an FEC decoder.

4.4 Summary

In section 4, general modulation techniques were briefly compared. The comparison was made on the basis of power efficiency, bandwidth efficiency, effects of nonlinearity, immunity to interference, and sensitivity to synchronization errors. Computer simulated results were presented to optimize the filter bandwidth and adjacent channel spacing. The implementation of binary and quarternary PSK modems were discussed.

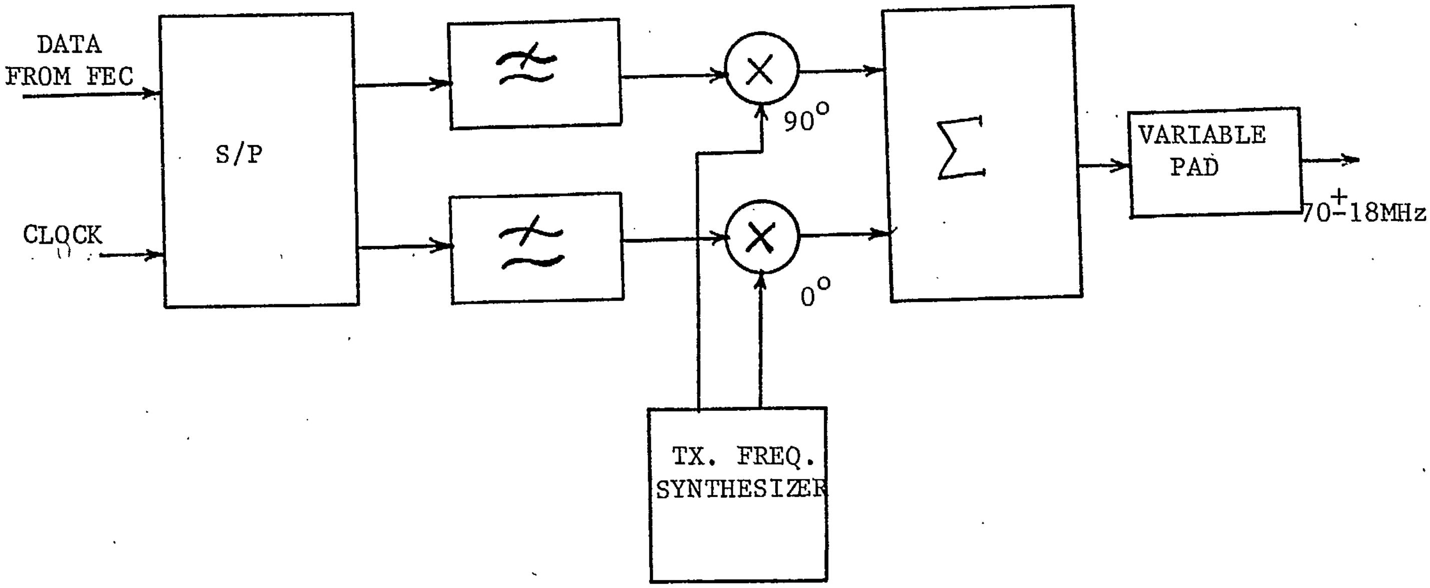


FIGURE 4.15 QPSK MODULATOR

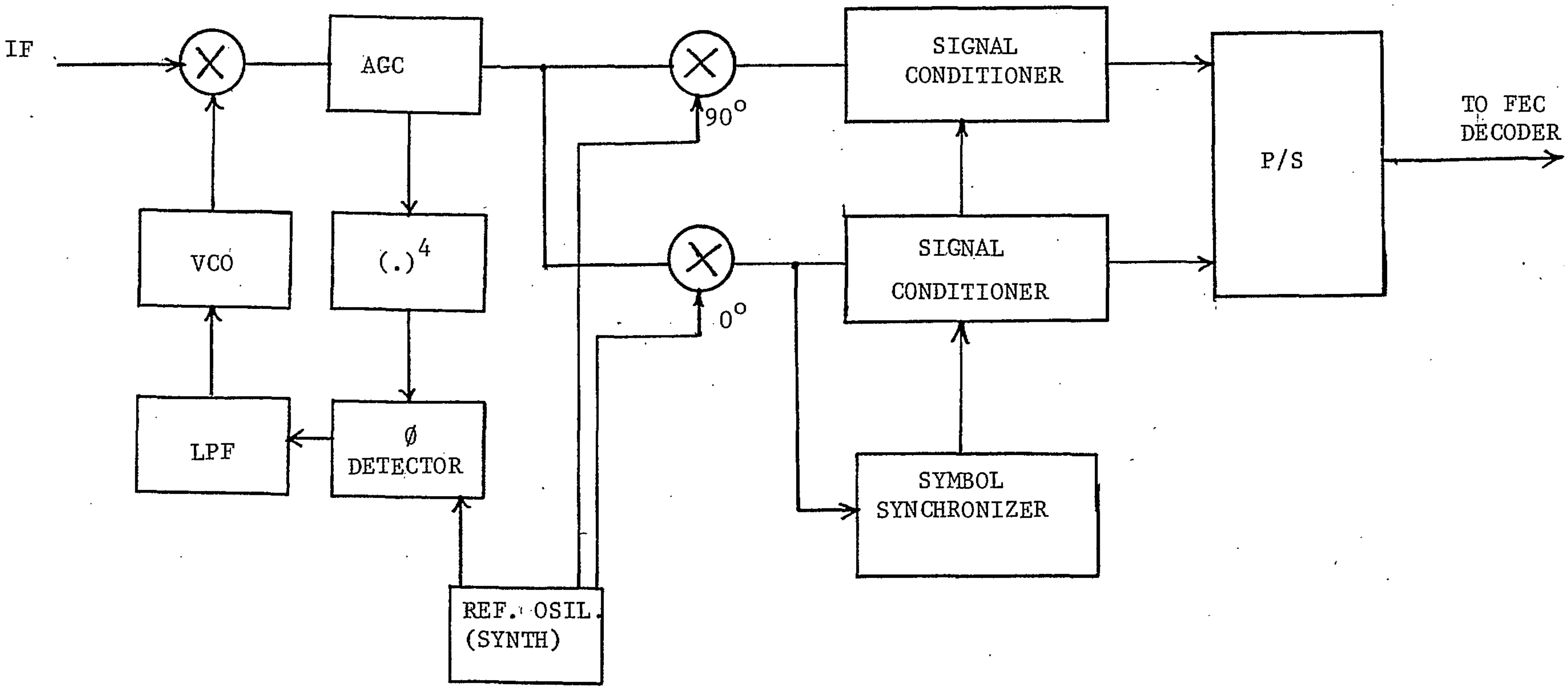


FIGURE 4.16 QPSK DEMODULATOR

5.0 TASK-6: ERROR CORRECTION

The sensitivity of digitally coded radio programs to bit errors is well-established. In order to minimize the effect of bit errors several different approaches are possible:

(a) improvement of received signal-to-noise ratio (C/N)

(b) error detection and concealment

(c) error detection and retransmission (ARQ error control)

(d) forward error correction (FEC) coding

Due to the type of link (satellite), the limitations on improvement based on (a) above are significant (e.g. limited amount of satellite power). Of course some improvement can be realized by optimum design of the channel spacing, etc. and this will be the subject of study of other segments of the project.

(b) can provide significant improvement [26] in the subjective sound quality of a radio program but is only effective for low bit error rates ($<10^{-5}$).

Due to distribution the scenario, (c) is not appropriate. Error Detection and Retransmission (even when used in a hybrid FEC/ARQ scheme) would involve a significant increase in network complexity (i.e. the requirement for a return link and large amounts of buffering).

(d) provides a promising means of improving the bit error rate. It can be used to improve the bit error rate to a point where other means (i.e. error concealment) can be used effectively to further improve the quality of received programs.

The purpose of this section is to outline and compare the various FEC coding schemes.

5.1 FEC Overview

FEC is used to combat the power limitations of digital communications links. With FEC coding, redundant parity bits are added to the data bits, resulting in an increase in the actual transmitted bit rate, but also helping to achieve a net reduction in the C/N_o required to obtain the same error rate performance as in the uncoded case. In general when the code rates are low (i.e. redundancy is high), codes with high gains can be obtained. As the code rate increases, the error correction capability of the codes decrease, and hence the coding gain also decreases.

5.1.1 Benefits of FEC Coding

The raw bit error rate on a digital link is determined by the energy-per-transmitted bit to noise density ratio (E_s/N_o)*, which in turn is determined by the received signal to noise density ratio (C/N_o):

$$C/N_o = E_s/N_o \cdot R_s \tag{5.1}$$

where R_s is the channel bit rate.

If no coding is applied to the system, then

$$E_s/N_o = E_b/N_o \text{ (the energy-per-information bit to noise density ratio)}$$

*Assuming the signal degradation is dominated by White Gaussian noise (true for typical satellite links).

and,

$$R_s = R_b \text{ (the information bit rate).}$$

If coding is applied (at code rate $r = R_b/R_s$) then (5.1) can be written:

$$C/N_o = E_s/N_o \cdot (R_b/r) = (E_s/N_o/r) \cdot R_b = E_b/N_o \cdot R_b \qquad (5.2)$$

Two scenarios are possible:

1) Bandwidth limited system - In this case the channel bit rate can not increase so the information bit rate must decrease. This reduction in information bit rate results in an improvement in BER (at fixed C/N_o) at the expense of reduced information throughput.

2) Power limited system - In this case the channel bit rate can change and no reduction in information bit rate is required. The increase in channel bit rate results in a reduction of the required C/N_o (at fixed BER) at the expense of the increased bandwidth.

In satellite systems the second scenario is the most common. As a result most performance analysis is made using this assumption.

5.1.2 FEC Performance Parameters

The standard measure of code performance is coding gain which is defined as the difference (expressed in dB) in the required E_b/N_o for a given bit error performance, between the ideal uncoded performance and the particular coding scheme (assuming unlimited bandwidth)*. In order to compare coding gains, coding schemes should be compared at the desired output bit error rate.

Another important parameter for coding schemes is the code rate**, i.e. the number of information bits per transmitted bit. This provides a measure of how much the bandwidth must be expanded to achieve the specified coding gain.

The traditional measure of bit error rate performance has major drawbacks when non-Gaussian channels are considered (i.e. non-independant bit errors). For these applications additional performance measurements are desirable which relate to the burst-error correcting ability of the various schemes. These measurements are very dependent on the particular application (transmitted information data structure and channel characteristics). Due to the expected channel characteristics these types of measurements will not be considered in this report.

In addition, performance can be compared in terms of processing speed or delay. For complex schemes, processing delays can be significant. Due to the characteristics of the projected application (no return link), the processing delay is not critical. Of more importance is the processing complexity. The processing speed must be such that codecs at the desired data rate are realizable without significant cost.

5.1.3 Performance of Coding with Non-Ideal Channel

As mentioned in the previous section coding performance is normally compared assuming ideal coherent modulation in a

*From (5.2) it is apparent that, for a power limited system (R_b fixed), E_b/N_o is directly proportional to C/N_o regardless of code rate.

**Alternatively the bandwidth expansion factor (the reciprocal of the coding rate) can be used for comparison.

Gaussian White Noise Environment. In practice, actual coding gains often surpass by significant amounts the coding gain predicted using these assumptions [5.2]. In particular the impact of phase and timing errors is about the same for coded as for uncoded PSK* in the ranges of interest. If the code is transparent to phase inversions, differential encoding/decoding can be placed before/after the encoder/decoder and performance is degraded by only 0.1 dB as opposed to 0.3 dB for the uncoded system.

In the case of non-ideal modem performance, the coding gain is typically better than in the ideal case. Take the example given in [17] (see Figure 5.1). Ideally, the uncoded performance is 1×10^{-5} at $E_b/N_o = a$. With coding, ideally, the performance is 1×10^{-5} at $E_b/N_o = d$, giving a coding gain of G. Given that the code rate is R, $E_s/N_o = c$ for the ideal case. The actual modem performance curve shows a loss L_c from ideal at the resultant raw channel bit error rate. The loss reduces the amount of gain of the coded system from the actual E_b/N_o performance curve giving:

$$G_{actual} = G_{ideal} + L_u - L_c \qquad (5.3)$$

Typically $L_c < L_u$ so better than ideal coding gain results.

5.2 System Overview

The environment in which a FEC coding scheme is to be implemented has direct influence on the coding scheme selection process. As a result a summary of the expected characteristics is required.

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*The loop bandwidths should be narrower with respect to the bit rate in the coded case.

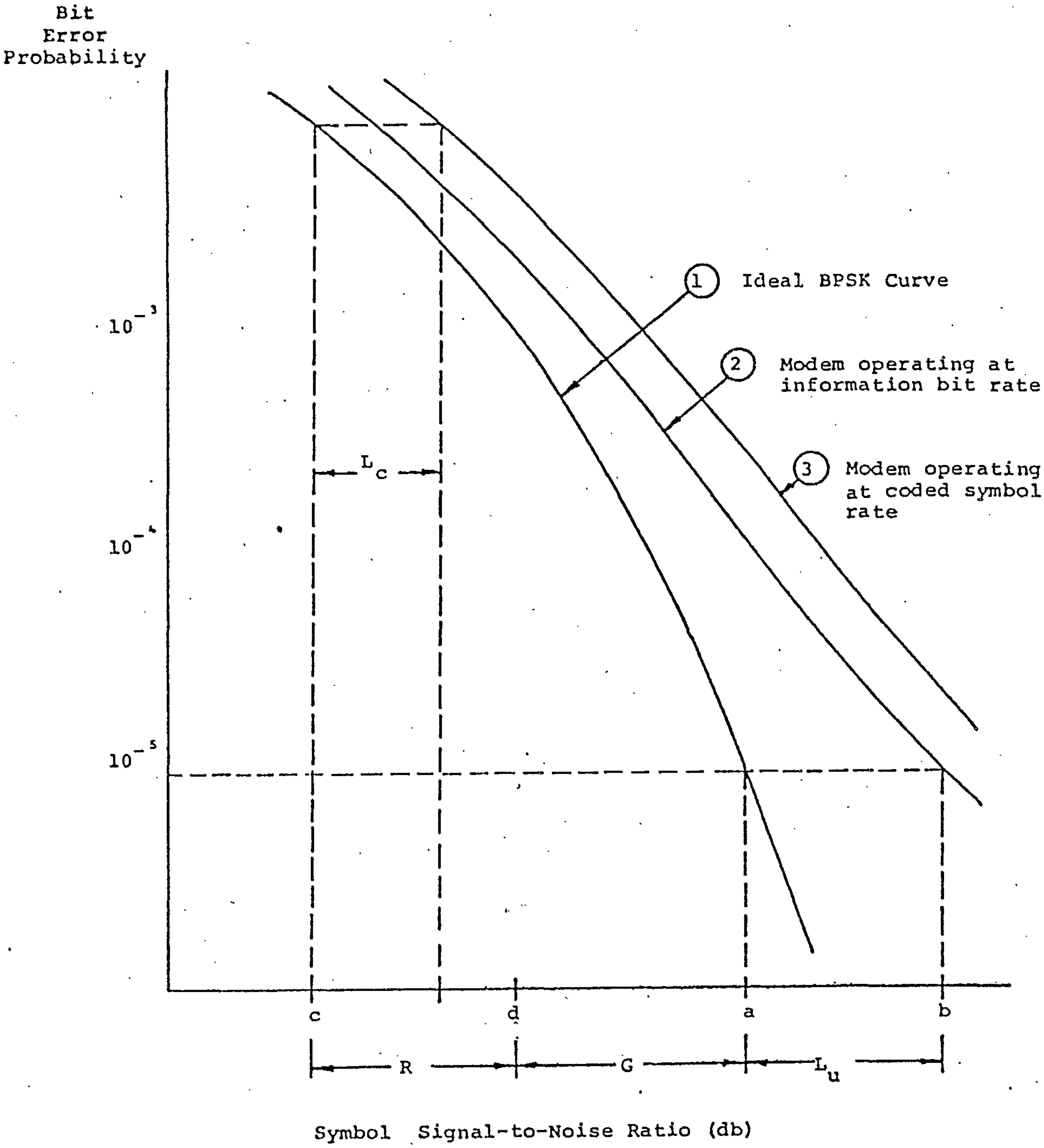


Figure 5.1 Ideal and Typical modem performance at information bit rate and coded symbol rate. Dashed lines illustrate calculation of actual coding gain $L_u + G - L_c$ from ideal coding goin G.

5.2.1 Link Structure

Assuming no subsequent ARQ requirement, the distribution link is one-way. It is expected that data from one stereophonic or two monophonic audio channels and a low bit rate data channel will be multiplexed onto a single transmitted carrier. There is also a possbility that two stereophonic channels could be multiplexed onto a single carrier. In either case some sort of multiplexed data structure is implied. See overall block diagram in Figure 5.2.

5.2.2 Modulation Techniques

Several modulation schemes are being considered [28], including:

- BPSK - coherent and differential
- QPSK - coherent and differential
- OQPSK - offset (or staggered) QPSK
- MSK - coherent or differential
- Duo-MSK
- TFM

The choice of modulation scheme can have a significant effect on which coding scheme is selected. With differential modulation schemes, errors tend to occur in pairs. If differential encoding is used with coherent modulation, errors also tend to occur in pairs*. In either

*If the code is transparent to phase inversions and the differential encoding process takes place before the FEC coding this does not apply.

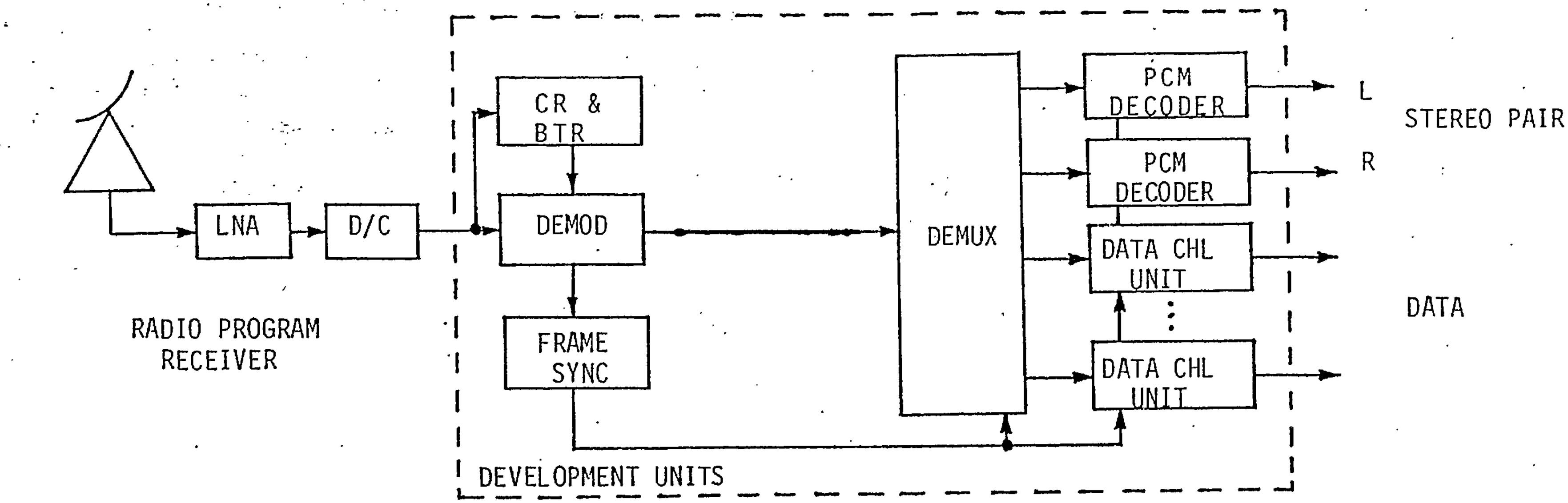


Figure 5.2 - Receiver Block Diagram (without FEC coding)

case single error correcting coding schemes can not be used unless some sort of interleaving is also used.

5.2.3 Channel Characteristics

In the case of coherent modulation the transmission channel can be assumed to be Gaussian. Some burst errors will occur due to fading, loss of carrier, clock or frame synchronization but with proper receiver/demodulator design these effects will be minimal.

5.2.4 Error Detection

Preliminary study in the source coding has indicated the need for one parity bit per PCM word. This will be used for error detection and concealment. This can be viewed as independant from FEC coding. It simply implies a requirement for a specific BER after FEC decoding (1×10^{-5} to 1×10^{-7} considered in this report).

5.2.5 Source Coding

Source coding schemes being considered include [28]:

- Instantaneously companded PCM
- Near-instantaneously companded PCM

With both schemes the effect of bit errors depends greatly on the position of the bit. Bit errors in the most significant bits of a PCM word will result in annoying clicks while bit errors in the least significant bits will result in little if any perceptible degradation of signal quality. One way to reduce this problem is to use error detection and concealment techniques on the most significant bits [28]. Normal FEC coding can then be applied to the full PCM word in order to reduce BER to a level where error detection and concealment can be used.

Near-instantaneously companded PCM introduces an additional factor. A scale-factor word (or range word) must be transmitted with each group of 20 to 30 samples. Bit errors in range words can result in clicks and short term amplitude change. Due to the slow rate of change of range words (for most types of audio programs) the effect of these bit errors is comparable to the effect of bit errors in the most significant bits of the PCM words if concealment is used [29].

5.2.6 Multiplexing

The audio channels (Left and Right) and data channels may be coded together or separately. This can effect the speed at which the codec must operate and the number of codecs required. It must be noted that additional error protection may be required on the data channels. This may be part of the Radio Program Receiver or part of end user equipment.

5.2.7 Integration of Coding into System

The overall system block diagram was given in Figure 5.2. FEC coding can be applied in several places. It can be applied to the multplexed bit stream as a whole (at a comparatively high bit rate) or it can be applied to each channel individually (this allows for a more powerful code to be applied to the data channels).

The coding can be applied to every bit equally or it can be applied to the more critical bits only. In the former case standard codecs can be used and a single codec for the entire multiplexed bit stream is possible. In the latter case, custom codecs may be necessary for each audio channel or complex synchronization schemes will be necessary.

It does not appear that there will be a requirement for coding critical bits. The use of this type of scheme will imply significant increase in the complexity of synchronization circuits and will prohibit the use of most commercial codecs. As a result it is assumed that coding will be applied to all bits equally.

5.2.8 Summary of Parameters

The range of parameters being studied in the rest of the report includes:

r - 1/2 to 7/8

G_c - 2 dB to 5 dB

output BER - 1×10^{-5} to 1×10^{-7}

5.3 Candidate Coding Methods

There exists a large number of coding and decoding techniques for FEC coding. In this section some of the coding techniques suitable for coding of entire bit streams are briefly discussed. The techniques considered include:

Coder	Decoder
BCH Hamming Golay Reed Solomon Cyclic Product Codes	ROM-Look-Up Table Berlekamp-Massey-Chein

Coder	Decoder
Convolutional	Viterbi (hard/soft)
Wyner-Ash Codes	
Convolutional Self Orthogonal Codes (CSOC)	Sequential (hard/soft)
Convolutional Product Codes	Threshold (hard/soft)

5.3.1 Block Codes

5.3.1.1 BCH Codes

BCH codes are probably the most widely known class of multiple-error-correcting block codes which have the cyclic property. Even at very high rates, the implementation can be simple particularly when the Berlekamp decoding algorithm [30, 31] is employed. The practical implementation of a BCH codec can be efficiently done by combining Berlekamp-Massey's iterative algorithm [30, 32] for obtaining the error-locating polynomial, Chien's method [33] for searching its roots, and Burton's procedure [34] for eliminating the inversion step. There is also another implementation [35] approach which is similar. A Simple ROM Look-up table approach can also be used for high rate codes.

Although, in one sense, long BCH codes are asymptotically (as $n \to \infty$ where n is the block length) inefficient (where efficiency is defined as the ratio of minimum distance to the block length), powerful BCH codes can be found in the practical range of n and r.

The performance of some BCH codes is discussed in Section 5.4 and their implementation is discussed in Section 5.5.

5.3.1.2 Hamming Codes

Hamming codes are block codes with minimum distance 3, i.e. they correct all single errors within a block. They are in fact, the single error correcting group of codes within the family of BCH codes. Hamming codes by themselves do not possess the error correction capability required but could be used in a iterated scheme.

5.3.1.3 Golay Code

The Golay code is a perfect* polynomial code. It is in fact a triple-error-correcting (23,12) BCH code. The extended (24,12) Golay code is quite useful as it is exactly Rate 1/2.

5.3.1.4 Reed-Solomon Codes

The Reed-Solomon codes may be viewed as a subclass of the general BCH codes. This class of codes are principally powerful for multiple-burst-correction [36, 37]. A t error correcting code interleaved to degree i is capable of correcting all single bursts of length ti or less, any two bursts of length ti/2 or less, or any t bursts of length i or less etc. Powerful R-S codes can be found which correct both burst and random errors. Commercial codecs using the R-S code with Berlekamp's decoding algorithm has been reported [38]. They have principally been applied in military applications, where jamming is of concern, or mass

*This refers to the property of correcting patterns of t or fewer errors and no others, i.e. optimum redundancy.

storage media applications (computer storage or video disc).

Given that the expected channel will be dominated by random errors, Reed-Solomon codes are not considered further.

5.3.1.5 Cyclic Product Codes or Iterated Single Error Correcting Block (ISECB) Codes

Iterated single error correcting block codes (ISECB) (or cyclic product codes) for high rate applications can be obtained with reasonable code word lengths [39 - 41]. This is achieved by using higher rates for the component codes so that, after multiplication, the resultant code rate is high enough to ensure the required redundancy. These SEC codes can be generated from a primitive generator polynomial of arbitrary degree m and can be characterized by other parameters; block length $n = 2^m-1$, number of information digits $k = n-m$, and rate

$$R = \frac{k}{n} = 1- \left(\frac{m}{2^m-1}\right) \tag{5.4}$$

Very high rate codes $(R \to 1)$ can be achieved for moderate m values. For the iterations of p SEC codes, the resultant code rate is

$$R_p = \prod_{i=0}^{p-1} \left(1 - \frac{m_i}{2^{m_i}-1}\right) \tag{5.5}$$

The implementation of ISECB codes is described in [39, 40].

ISECB have definite advantages for high rate applications (i.e. complexity) but this advantage diminishes for lower rates. These types of codes are therefore more suited for high code rate applications.

5.3.2 Convolutional Codes

Convolutional codes involve the coding of data continuously as opposed to block by block. It has been shown [4, - 43], that non-systematic convolutional codes perform much better than systematic ones. As a result most convolutional codes are non-systematic.

Decoding can be performed by the sequential method, threshold techniques or the Viterbi algorithm. Although the use of sequential decoding can result in very good coding gains, sequential decoding is dropped from further consideration due to the decoder complexity and the buffer overflow problems [37] that occur when the input error rate exceeds the designed capability.

One particular concern with Viterbi decoding is the tendency for burst errors to be produced at the output of the decoder. The burst error statistics* of a practical decoder were investigated in [44]. It was found that (at a output BER of 1×10^{-7}) the average burst error length for a rate 1/2 code was 5. For a rate 3/4 code the average burst error length was 7.8. The average burst length for both these codes increases very slowly as the bit error rate increases to 10^{-4}. The problem of burst errors can be easily overcome by adding a interleaver so that the error detection and concealment equipment can perform properly.

*A burst error was defined as a sequence of bits containing erroneous bits followed by at least 20 correct bits. The average proportion of erroneous bits per burst error was about 50%.

5.3.2.1 Wyner-Ash Codes

Wyner-Ash codes are single-error correcting convolutional codes, characterized by short constraint lengths and simple circuitry. Wyner-Ash codes can have the rate $(2^n-1)/2^n$, i.e. 1/2, 3/4, 7/8,...etc. They have the capability of correcting 2 errors separated by at least 10 bits in the Rate 3/4 case and 29 bits in the Rate 7/8 case.

[45] presents some performance measurements made for the Wyner-Ash code. These results show poor coding gain. The coding gain available from this type of code (at rate 3/4) is approximately 1.3 dB at BER = 1×10^{-7}. As a result Wyner-Ash codes are not considered further.

5.3.2.2 Convolutional Self-Orthogonal Codes (CSOC)

A convolutional code is defined as self-orthogonal if the set of syndrome digits checking each block zero information error digit forms a set of parity checks which are orthogonal on that digit. Robinson and Bernstein [46] derived a construction method for obtaining self-orthogonal convolutional codes by means of a set of distinctive difference triangles. Wu [47 - 49] obtained a large number of threshold decodable CSOC codes. It is well known that CSOC codes are threshold decodable [50] and can correct every error pattern which is guaranteed to be correctable by the minimum distance of the code. The advantages of CSOC codes with threshold decoding are:

(1) simple implementation,

(2) free from error propagation,

(3) guaranteed correction capability beyond the minimum distance, d_{min} (used with appropriate decoder),

(4) capability of operating at very high speed, and

(5) a large number of codes.

As will be shown in Section 5.5, CSOC codes are not extremely powerful but because of property (2) above they are very attractive for concatenated codes [49].

The performance of selected CSOC codes is given in Section 5.4 and their implementation is discussed in Section 5.5.

5.3.2.3 Convolutional Product Codes or Iterated Single Error Correcting Convolutional (ISECC) Codes

Convolutional product codes, like iterated block codes, are obtained by the iteration of single error correcting codes, in this case convolutional codes [51], and can be characterized by the parameters $n = 2^m$, $k = n-1$, and

$$R = 1-2^{-m} \tag{5.6}$$

As in (5.5), the resultant code rate R_p, for the iterations of p SEC codes can be given by

$$R_p = \prod_{i=0}^{p-1} \left(1 - \frac{1}{2^{m_i}}\right) \tag{5.7}$$

It can be seen from (5.6) and (5.7) that the resultant convolutional code gives a higher code rate than the block code for the same amount of parity check digits per block.

ISECC codes are not considered further for the same reason as ISECB codes.

5.3.3 Concatenated Codes

The use of concatenated coding has been pointed out as a good way to increase coding gain while still maintaining low cost [52]. The standard approach is to use a simple block or convolutional inner code and a Reed-Solomon outer code. In this way the simple inner code hardware operates at the channel data rate while the more complicated Reed-Solomon hardware (which operates on blocks of data at a time) only has to operate at a fraction of the channel data rate. In this way the use of expensive high-speed logic components is avoided in high data rate applications (>10 Mbps).

Given the comparitively low data rate expected for the radio distribution system, it is not necessary to use this type of concatenated approach.

5.4 Probability of Error Performance

Performance is presented in terms of decoded bit error rate (BER) versus E_b/N_o, where N_o is the single-sided power spectral density of the channel noise, E_b is the received energy per information bit.

The signalling scheme is taken to be a form of ideal antipodal signalling (e.g. BPSK). The theoretical probability of a signalling bit error with ideal antipodal signalling in an additive white Guassian noise (AWGN) channel is well known and is given by [31, pg. 13]

$$p = Q[(\frac{2E_s}{N_o})^{\frac{1}{2}}] \tag{5.8}$$

$$= Q[(\frac{2rE_b}{N_o})^{\frac{1}{2}}] \tag{5.9}$$

where $Q[X]$ is the area from X to infinity under the tail of the normal distribution, i.e.

$$Q[X] \triangleq \int_{X}^{\infty} \frac{1}{\sqrt{2\pi}} \exp(-\frac{\alpha^2}{2})d\alpha \tag{5.10}$$

The impact of a particular coding scheme on the overall system design depends on certain system requirements and restrictions. For example, if the modulation type and information rate are fixed then the signalling rate and corresponding signalling bandwidth must be increased by a factor of 1/r (where r is the code rate). When this is the case the BER performance is most often plotted versus E_b/N_o. If the signalling rate and bandwidth are fixed then the information rate must be reduced by a factor of r. When this is the case it is often convenient to plot BER performance against E_s/N_o. The former form of presentation has been used.

The decoded BER performance is presented for the following coding schemes:

(a) BCH block codes,

(b) Convolutional codes with Viterbi decoding.

(c) Convolutional self-orthogonal codes (CSOC's) with threshold decoding.

(d) The (24,12) Extended Golay code.

5.4.1 BCH Block Codes

First we present the method used for evaluating the probability of bit error at the output of a block code decoder. Then the performance of a number of BCH block codes is presented for the AWGN channel. Only hard decision decoding is considered.

An (n,k) block code contains k information bits in a coded block of n bits. Assume that the minimum distance, d, of the code is odd. Thus $t = (d-1)/2$ raw bit errors in a block of n bits can be corrected. We will assume the codec is unable to detect all other error patterns. Since the code can correct up to t errors, at most t additional errors are made if the decoder decodes incorrectly. Thus an upper bound on the probability of bit error, P_{odd}, after decoding is given by [31]

$$P_{odd} < \hat{P}_{odd} = \frac{1}{n} \sum_{i \geq t+1}^{n} (i+t) P_r(i) \qquad (5.11)$$

where

$$P_r(i) = \text{Prob}(i \text{ raw bit errors})$$

$$= \binom{n}{i} p^i (1-p)^{n-i} \qquad (5.12)$$

and p is the probability of a raw bit error before decoding.

If the minimum distance d is even then $t = (d/2)-1$ errors can be corrected and t+1 errors can be detected in a block of n bits. Thus an upper bound on the probability of bit error, P_{even}, after decoding is given by

$$P_{even} < \hat{P}_{even} = \frac{1}{n}\Big[(t+1)P_r(t+1) + \sum_{i \geq t+2}^{n} (i+t)P_r(i) \tag{5.13}$$

where $P_r(i)$ is as defined in (5.12).

The accuracy of the above approximations can easily be seen from the following argument. A code which can detect all uncorrectable error patterns is expected to perform better than one which can only detect a limited number or none at all*. The performance of such an unrealizable code would be given by:

$$P_{ideal} = \frac{1}{n} \sum_{i \geq t+1}^{n} iP_r(i) \tag{5.14}$$

Comparing (5.11), (5.13) and (5.14) term by term and noting that (i+t) is less than 2i for i greater than t, we easily obtain the following bounds for even and odd distance block codes.

<u>d odd</u>

$$P_{ideal} < P_{odd} < \hat{P}_{odd} < 2P_{ideal} \tag{5.15}$$

<u>d even</u>

$$P_{ideal} < P_{even} < \hat{P}_{even} < 2P_{ideal} \tag{5.16}$$

---

*This is clearly the case if errors occur independently as for an ideally interleaved error sequence, but may not be true if certain input error patterns are more likely than others.

Thus upper bounds $\hat{P}_{odd}$ and $\hat{P}_{even}$ are expected to be in error by at most a factor of 2. For a decoded bit error rate (BER) on the order of or less than 10^{-5}, a BER factor of 2 will usually, depending on the raw bit error performance of the channel, result in very little difference in terms of required E_b/N_o.

Performance has been evaluated for the odd-distance, BCH codes listed in Table 5.1. Other BCH codes can be found in references [31] and [36] for example. Each code in this table is given a code number for cross- referencing with the plotted performance results. Also given for each code is the code rate.

The decoded BER performance of some of these codes is shown in Figure 5.3. The performance is plotted against E_b/N_o. The coding gain of a particular code, at a given BER, can be obtained directly from these figures by subtracting the E_b/N_o required with coding (in dB) from that required without coding.

5.4.2 Convolutional Codes with Viterbi Decoding

Clark and Cain [31] evaluated the performance of convolutional codes using Viterbi decoding. Curves for code rates of 1/3, 1/2, 2/3 and 3/4 were given. The results for rate 1/2 and rate 2/3 are given in Figures 5.4 and 5.5.

These results are for infinite quantization. 3 bit quantization degrades performance by about 0.25 dB. Hard decisions result in approximately 2.2 dB degradation.

Code Number	n	k	t	Code Rate	Coding Gain[1]	
					@ BER=1×10^{-5}	@BER=1×10^{-7}
B1	63	30	6	.476	2.7 dB	3.4 dB
B2	63	39	5	.619	3.25 dB	3.9 dB
B3	127	64	10	.504	3.4 dB	4.3 dB
B4	127	78	7	.614	3.25 dB	4.1 dB
B5	255	123	19	.482	3.8 dB	5.0 dB
B6	255	139	15	.545	3.8 dB	4.9 dB

Table 5.1 - BCH Codes

[1] Assumes BPSK signalling.

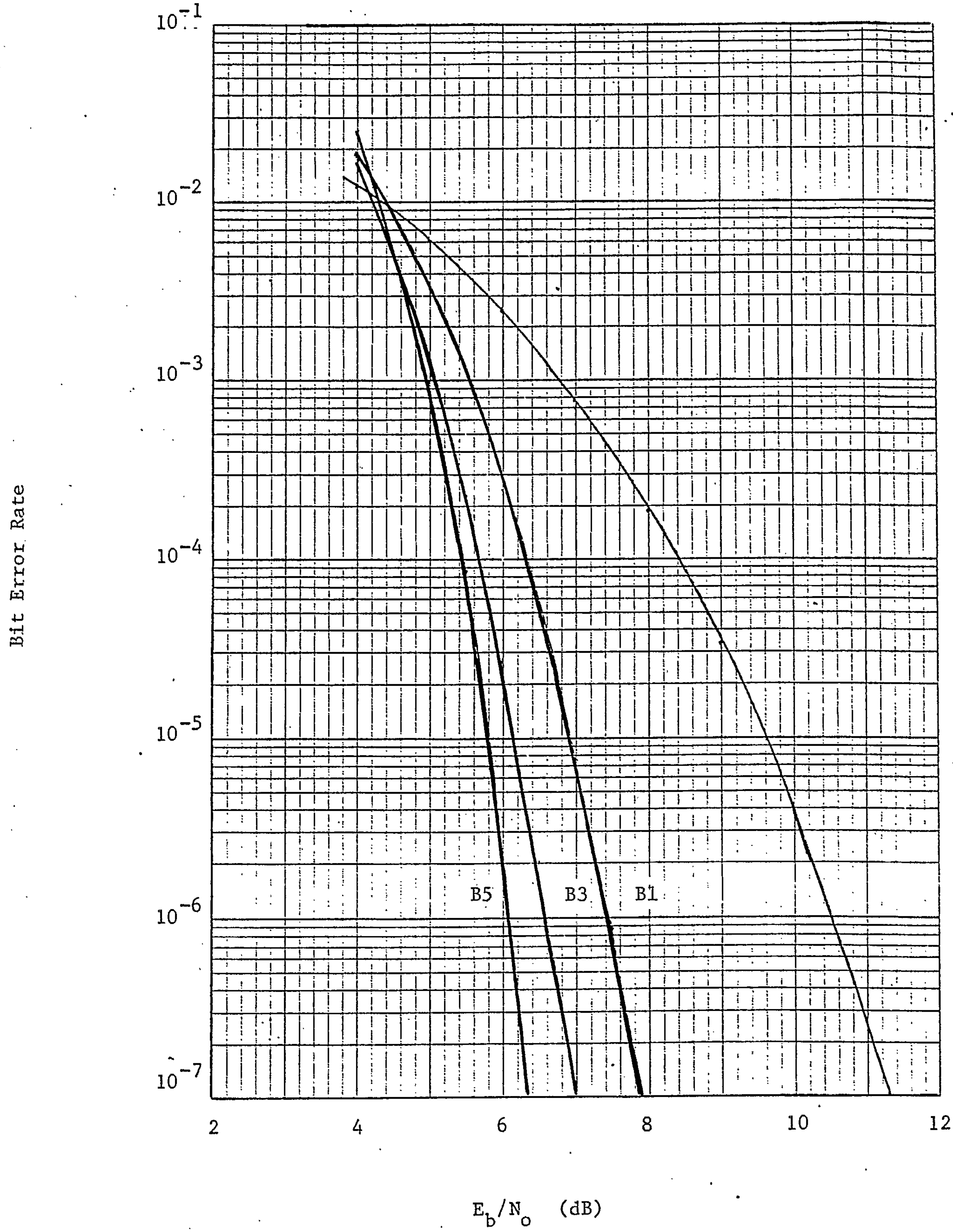


Figure 5.3- Performance of BCH codes

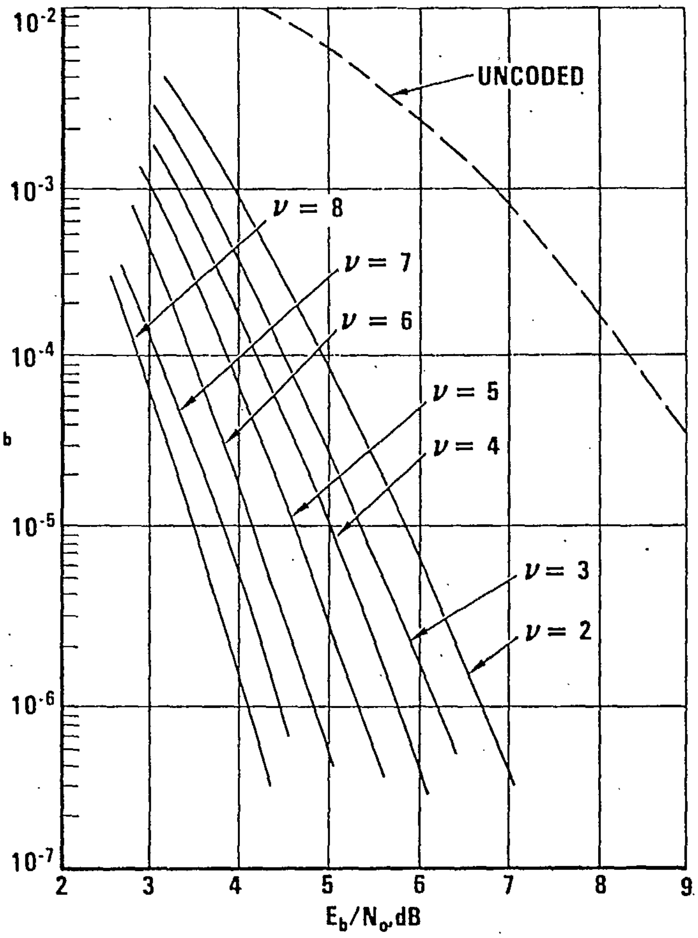


Figure 5.4 Bit error probability for $R = 1/2$ codes with Viterbi decoding (infinite quantization) and PSK modulation.

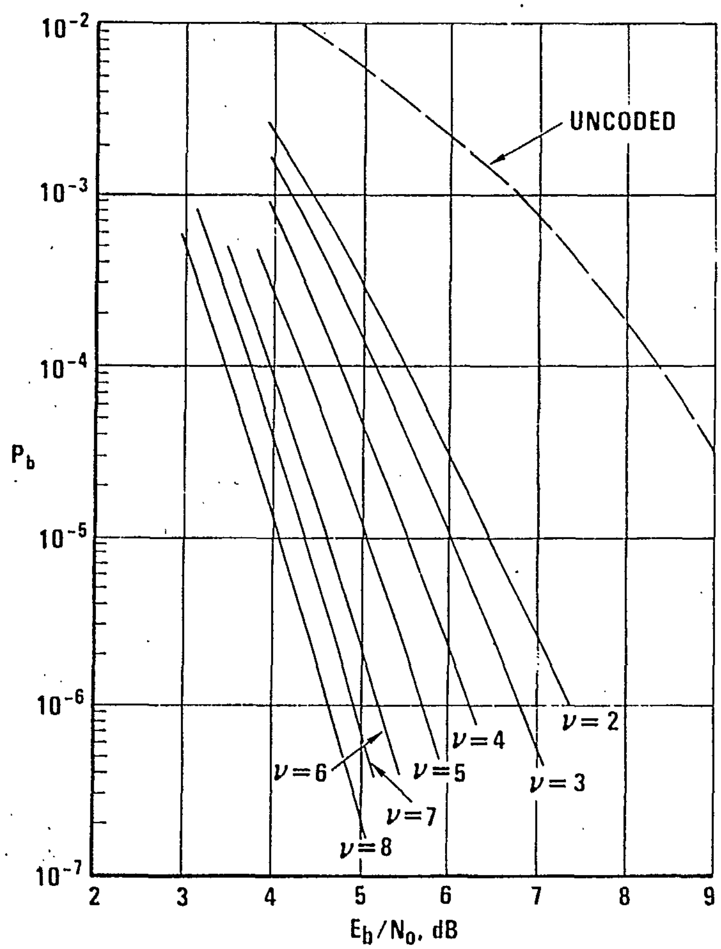


Figure 5.5 Bit error probability for $R = 2/3$ codes with Viterbi decoding (infinite quantization) and PSK modulation.

The performance of some candidates are listed in Table 5.2.

5.4.3 Convolutional Self-Orthogonal Codes with Threshold Decoding

Wu [47 - 49] tabulated a number of convolutional self-orthogonal codes (CSOC's). These codes and are known to be threshold decodable. According to Wu the decoded BER performance can be approximated by (see Equation (46) in [48])

$$P_b = \sum_{i=\frac{J}{2}+1}^{N_e} \binom{N_e}{i} p^i (1-p)^{N_e - i} \tag{5.17}$$

where N_e = effective constraint length of the code

J = d-1 = number of syndrome equations

d = minimum distance

N_A = actual code constraint length

p = raw channel BER

The effective constraint length, N_e, of a threshold decodable convolutional code is bounded by [30]

$$N_e < \frac{1}{2}[k_o J^2 + k_o J + 2] \tag{5.18}$$

where k_o is the number of inputs per block. The code rate, r, is given by

$$r = \frac{k_o}{k_o + 1} \tag{5.19}$$

Code Number	Code Rate	ν	Quantization	Coding Gain[1]	
				@ $1x10^{-5}$	@ $1x10^{-7}$
V1	1/2	3	3-bit	3.95 dB	4.05 dB
V2	1/2	4	3-bit	4.35 dB	4.55 dB
V3	1/2	5	3-bit	4.75 dB	5.05 dB
V4	2/3	3	3-bit	3.35 dB	3.5 dB
V5	2/3	4	3-bit	3.85 dB	4.15 dB
V6	2/3	5	3-bit	4.3 dB	4.65 dB
V7	1/2	6	hard	3.25 dB	3.65 dB
V8	1/2	7	hard	3.6 dB	4.15 dB
V9	1/2	8	hard	3.85 dB	4.4 dB

Table 5.2 - Convolutional Codes with Viterbi Decoding

[1] Assumes BPSK Signalling

The performance of a number of rate 1/2 CSOC's has been evaluated using (5.17) or an equivalent form. Other CSOC's and their generating polynomials can be found in references [47 - 49].

The decoded BER performance for the rate 1/2 code with $J=14$, $N_A = 312$ is shown in Figure 5.6. The coding gain at $BER = 1 \times 10^{-7}$ is 3.2. The performance of rate 1/2 codes with longer constraint length (N_A) are better but not significantly. The rate 1/2 code with $J=24$, $N_A = 1096$ has about 0.4 dB more gain at $BER = 1 \times 10^{-7}$.

The performance of several other CSOC codes are shown in Figures 5.7 and 5.8 (from [53]).

In general the most efficient CSOC's (in terms of error correction) are those with even values for J (i.e. odd minimum distance). The number of errors which can be corrected is J/2.

5.4.4 The Golay Code

Figure 5.9 (from [37]) compares the peformance of the Golay code with several other common codes. The gain of the Golay code is small (2.4 dB at 10^{-7}) but as will be seen in Section 5.5.2 the implementation can be very simple.

5.5 Implementation and Operational Considerations

5.5.1 BCH Codes

5.5.1.1 Encoding

Encoding of BCH codes is performed using a feedback shift register and a multiplexor. The parity bits are generated in the shift register as the information bits within a

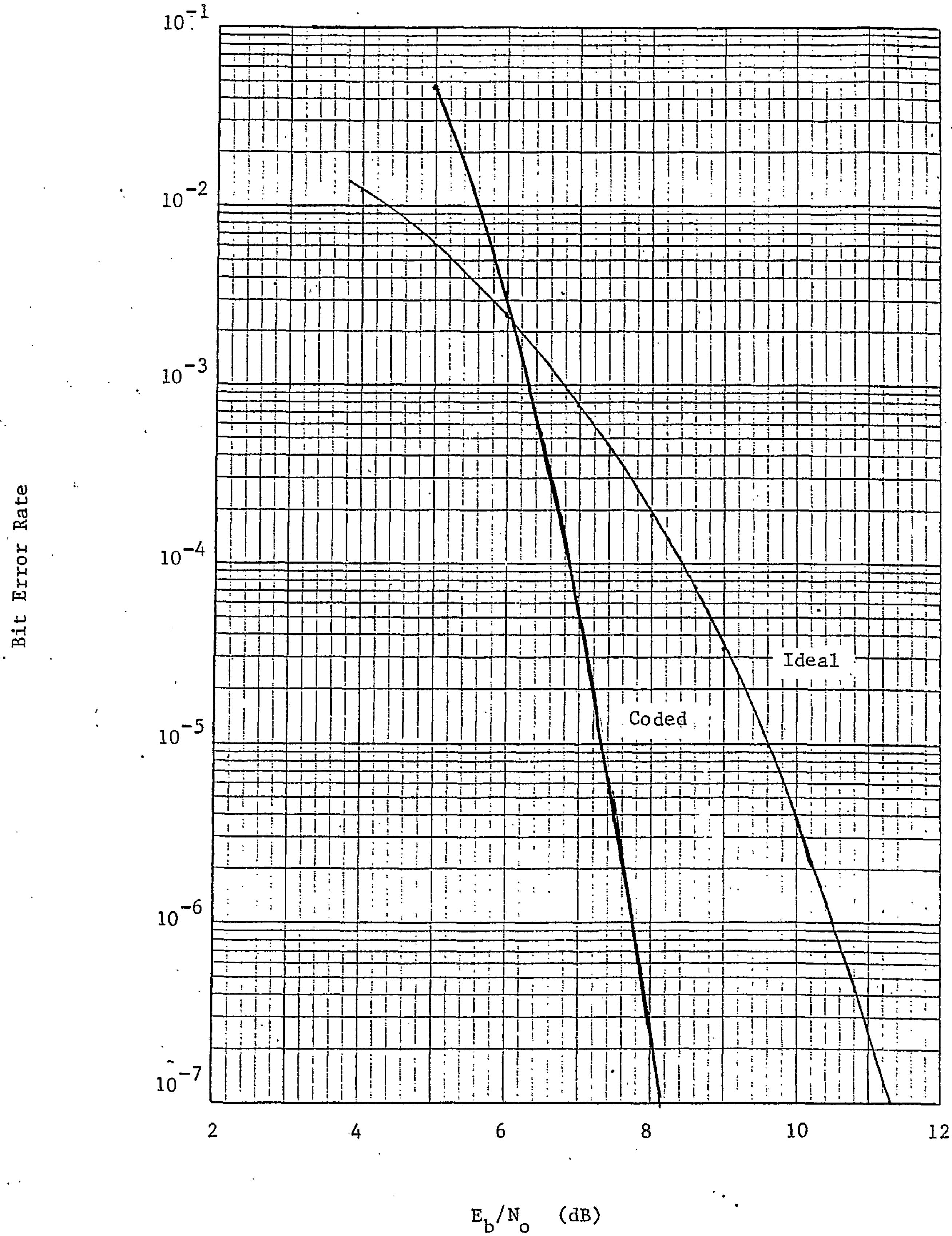


Figure 5.6 Performance of CSOC code

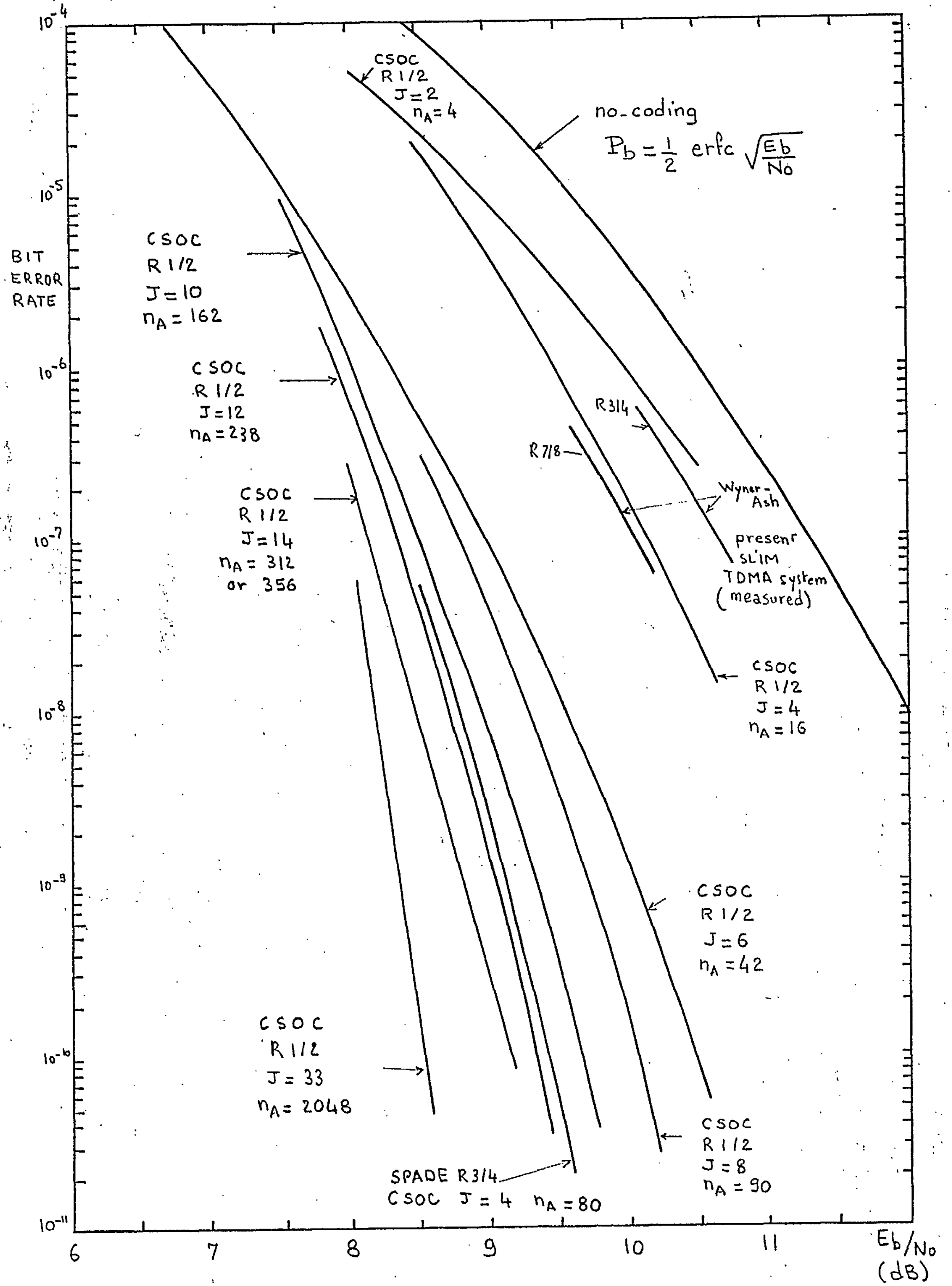


Figure 5.7 -COSC'S AND WYNER - ASH CODE PERFORMANCES

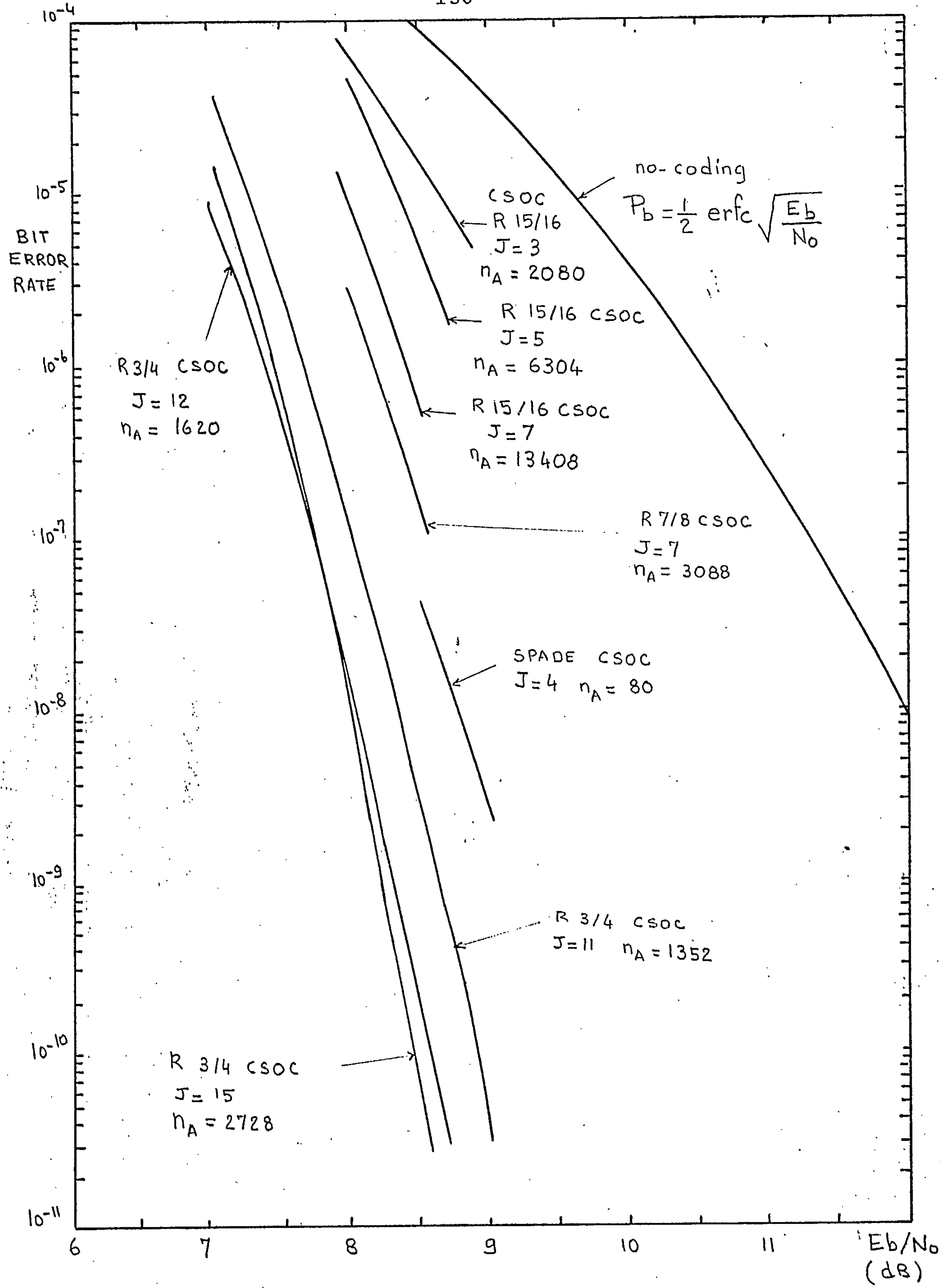


Figure 5.8- CSOC'S PERFORMANCES

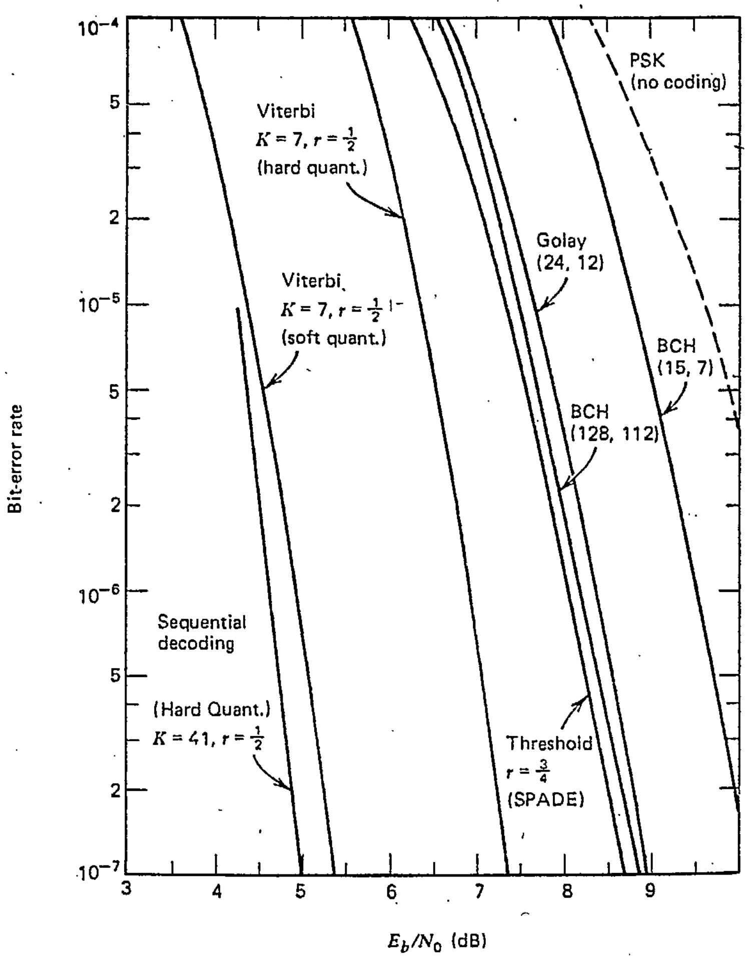


Figure 5.9 - Performance of Several Coding Schemes [5.17]

block are simultaneously added at the shift register taps and shifted out of the decoder. When all information bits have been shifted through, the parity bits are shifted out and the shift register cleared.

5.5.1.2 Berlekamp-Massey-Chien Techniques

A typical time domain decoding technique is shown in Figure 5.10. The transform computer produces the syndrome(s) components. A buffer device stores the received sequence r. The error-locator polynomial $\Lambda(z)$, is computed from the syndrome by solving the key equation. An error-evaluator polynomial $\Omega(z)$ is also generated when solving the key equation. The device labled "Chien search" finds the error locations given the $\Lambda(z)$ polynomial by simply finding the inverse roots. Once the roots of the error locator polynomial are known the errors are corrected readily.

One important step in the decoder is the determination of the minimum-degree error-locator polynomial $\Lambda(z)$ which satisfies the key equation. Earlier procedures for solving the key equation made use of standard matrix inversion techniques. But Berlekamp [30] later devised a simple iterative scheme that provides the most practical implementation and is discussed in detail in [31].

The overall decoder involves three processes:

(1) syndrome calculation,

(2) solution of the key equation, and

(3) Chien search.

A hardware implementation may require a processor with the following units:

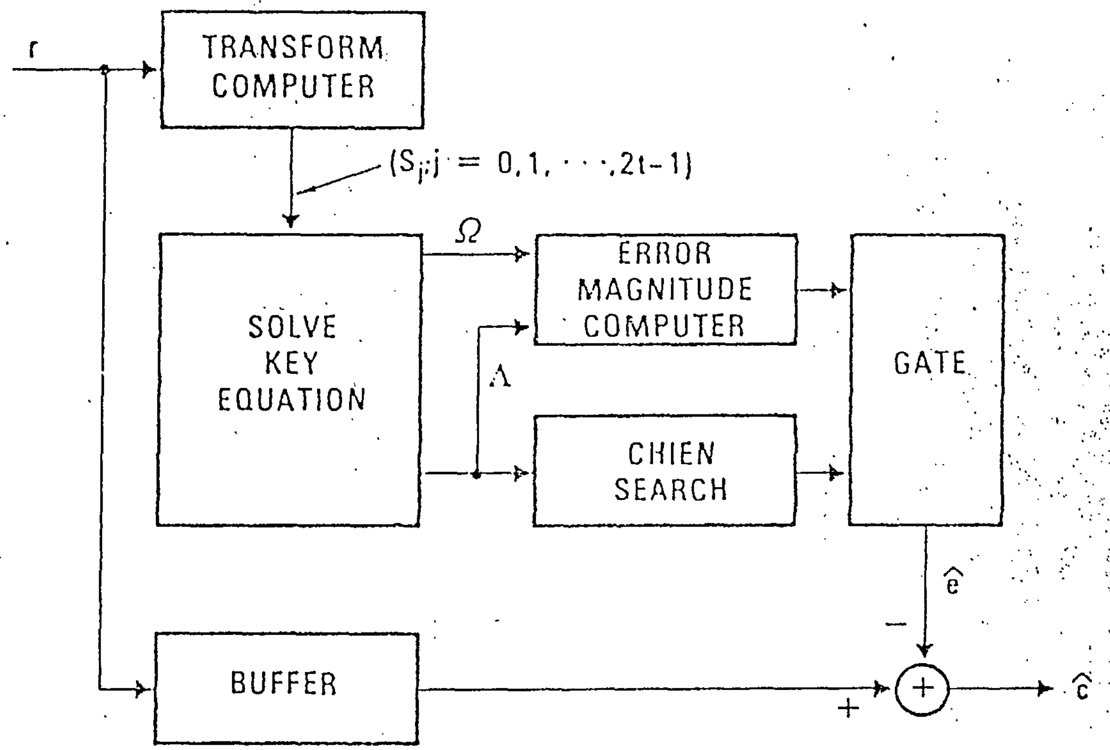


Figure 5.10 Time domain decoder for BCH codes.

(1) central processing unit (CPU),

(2) RAM for buffering and temporary storage of intermediate results,

(3) ROM for storage of computational subroutines and control of decoding process, and

(4) interfaces with the rest of the system.

The interesting aspect of this type of implementation is the inherent flexibility which takes advantage of certain situations which are easy to decode. For example:

(1) When the syndrome polynomial is zero, no error correction need be done. Thus, all the operations associated with Berlekamp algorithm and Chien Search may be skipped.

(2) In addition, the single error case is easily accommodated through special calculations.

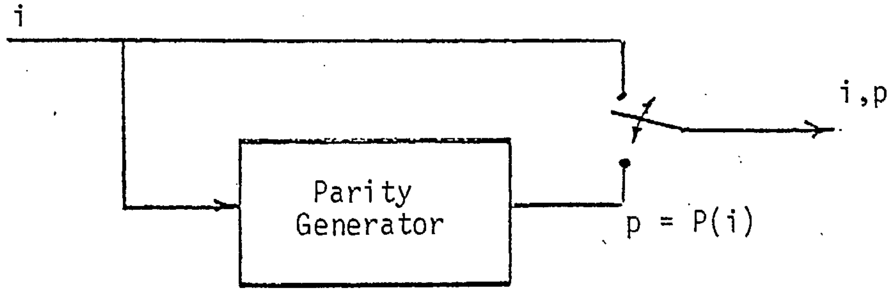
Thus, in applications where the zero or one error case may occur a high percentage of time very little computation is required. At high data rates the syndrome calculation can be done in a hardware peripheral unit. At still higher data rates one can perform the Chien search in a peripheral unit leaving only the solution of the key equation in the CPU. Also a parallel realization of the Berlekamp algorithm suggested by Massey [32] can be used to achieve very high data rates.

An efficient implementation of a BCH decoder for both high code rate and high data rate is reported in [35]. The machine is capable of operating at 34 Mb/s transmission rate and decodes a shortened BCH code (2196, 2136) with only 2.7% redundancy. Since BCH codes are capable of correcting random and burst errors a two stage decoder is used in [35]. These two parallel and independently working stages are the burst decoder and the decoder for random errors. When only random errors are present the burst decoder is not necessary for achieving the required decoded error rate. The burst decoder will also correct single random errors. The first decoding attempt is made by the burst error corrector which is followed by a random error corrector. Both the burst and random error decoders are able to work independently, so that one of the decoder stages can be switched off. Although the implementation is slightly different from the one described earlieer, this decoder algorithms also makes use of a combination of algorithms due to Peterson, Massey, Berlekamp, and Chien.

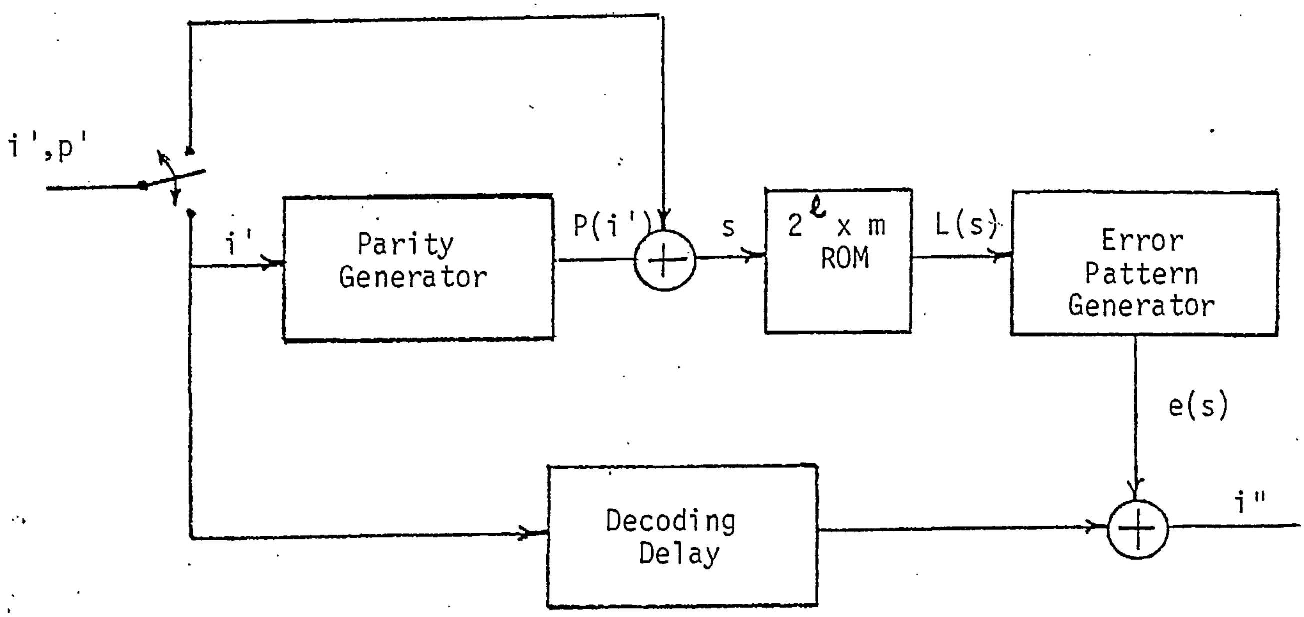
5.5.1.3 A Simple ROM Table Look-up Decoder

Figure 5.11 shows a very simple decoder for an (n, k) BCH code where he number of parity bits, $\ell = n-k$, is small compared to the block length, n. It consists of a parity generator, a $2^{\ell} \times m$ bit ROM, an error pattern generator, a decoding delay shift register, a ℓ-bit modulo-2 adder and a k-bit modulo-2 adder. The parity generator used in the decoder is identical to the parity generator used in the encoder. The scheme uses the parity generator and the first mod-2 adder to calculate the syndrome. The syndrome is used to address the ROM and yields the location of the error(s). The error pattern generator and the second mod-2 adder are used to correct the errors.

Let us assume that the block to be corrected consits of a k-bit information portion i' and an ℓ-bit parity portion



(a) Encoder



(b) ROM Decoder

Figure 5.11: A Simple ROM Decoder For High Rate BCH Block Codes

p'. The primes on these vectors indicate that they may contain errors. To calculate the syndrome we pass the k-bit vector i' through the parity generator. The result is an ℓ-bit vecotr $P(i')$. If we add $P(i')$ to p' we obtain the ℓ-bit syndrome, s. That is

$$s = P(i') + p' \tag{5.20}$$

The ROM is used to store the location of the errors corresponding to each syndrome. The dimension of the ROM must be $2^{\ell} \times m$ where 2^{ℓ} represents all possible values of the syndrome and m is the number of bits required to uniquely determine the location of the errors. The syndrome is used to address the ROM. Call the contents of the ROM $L(s)$. The value of m is given by

$$m = t \times q \tag{5.21}$$

where t is the maximum number of errors to be corrected, and q is the smallest integer such that $n < 2^q$. Each set of q bits can specify the location of one error. The location of the errors, $L(s)$, is then passed to the error pattern generator (which decodes the information $L(s)$ to give the error pattern $e(s)$). The error pattern $e(s)$ is then added to the information vector, i', to give the corrected information vector i''. Thus

$$i'' = i' + e(s) \tag{5.22}$$

As an example, consider the (255,239) BCH code. This code can correct $t = 2$ errors. The number of parity bits is $\ell = 16$, and $m = 2 \times 8 = 16$ bits. Thus the ROM size required is 128 k bytes (8-bit bytes). With todays technology such memory requirements are easily accomodated with only a few chips. On the contrary the (63,39) BCH code would require 63 M bytes of memory.

5.5.2 The Golay Code

The Golay Code, given its simplicity, is most easily decoded using the ROM look-up table approach (see Section 5.5.1.3). For the extended (24,12) Golay code two 4096 by 12 bit ROM's will be required as well as two 12 bit modulo 2 adders. In this case the parity generation is performed using the first ROM and the output of the second ROM is the error pattern (no error pattern generation circuitry required).

5.5.3 Convolutional Codes with Viterbi Decoding

5.5.3.1 Convolutional Encoder

The block diagram of a convolutional encoder is shown in Figure 5.12. It consists of a bank of shift registers, coding polynominal generator and a commutator which samples at a rate higher than the uncoded data input. The coding polynomial generator simply connects the appropriate shift registers' outputs to the inputs of modulo-2 adders based on a generator polynomial representing a specific code.

5.5.3.2 Viterbi Decoding

A functional block diagram of a general Viterbi decoder is shown in Figure 5.13. It accepts at the input successive quantized outputs from the demodulator and delivers the coded information stream at the output. The machine consists of [31]:

(1) synchronizer,

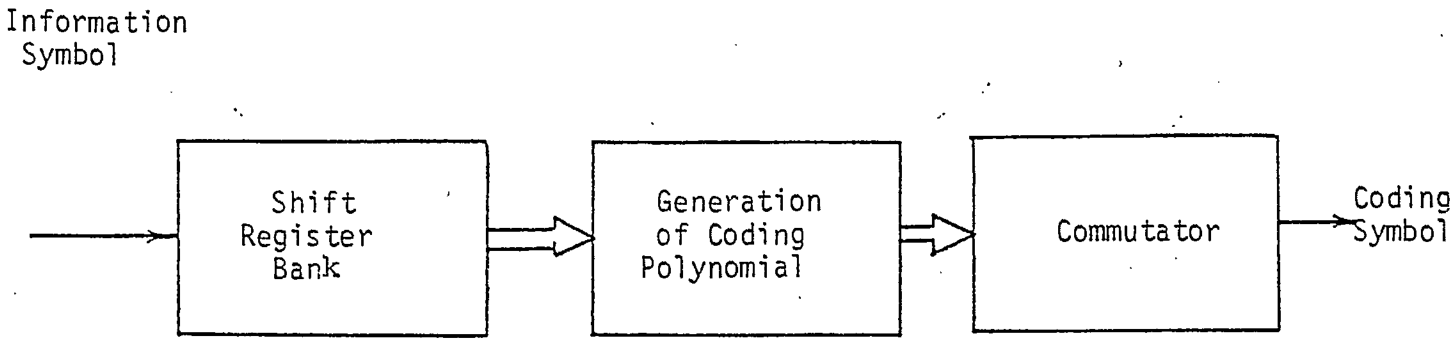


Figure 5.12:A general convolutional encoder

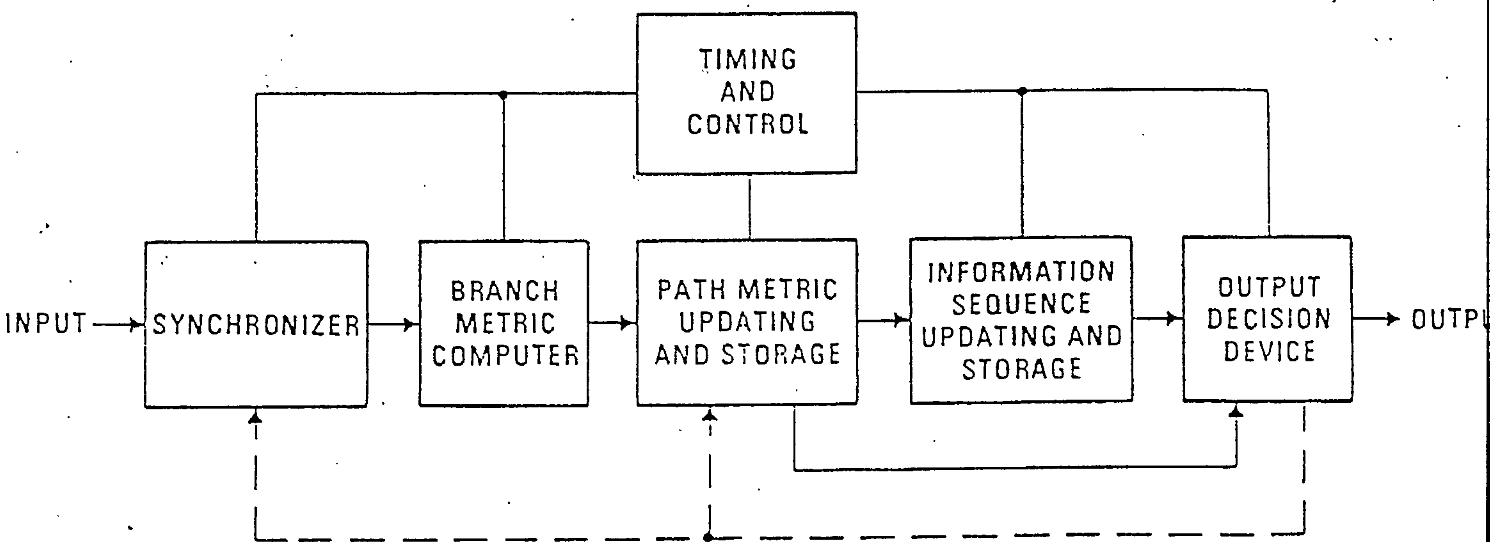


Figure 5.13:Functional block diagram for a Viterbi maximum likelihood decoder

(2) branch metric computer,

(3) path metric updating, comparison and storage,

(4) information sequence (hypothesized) updating and storage, and

(5) output decision device.

The synchronizer is a device for determining the start of a branch in the received symbol stream and resolving symbol parity when required. It is well known that convolutional codes have the property that branch synchronization can be resolved without the addition of synchronization bits because the received data appear to have an excessive error rate when not properly synchronized. Obviously the complexity of the decoder is reduced if synchronization bits (e.g. unique word) are provided.

The branch metric computer determines a new set of metric values for each branch in the code trellis each time a new branch is received. Ideally the branch metric is proportional to the logarithm of the probability that the specified branch was transmitted. The path metrics are computed as the sum of branch metrics along a given path. Only the most likely path into each of a set of possible states (path history) is stored at any given time. The length of the path that must be stored is strongly dependant on the merging properties of the code. Satisfacotry performance is achieved only at decoding depths at which most low weight unmerged paths disappear. Given the accepted decoding length, the job of the output device is to decode the oldest symbol (or symbols) in the sequence having the minimum path metric.

In order to efficiently implement a Viterbi decoder there are several practical design aspects to be considered at every stage. For example, quantizing the branch and path metrics, storing the path metrics, incorporating often used add-compare-select (ACS) techniques, etc. Some of these are discussed in [31].

5.5.4 Convolutional Self-Orthogonal Code (CSOC)

The CSOC encoder makes use of a shift register as in all convolutional encoders. With a CSOC code the information bit(s) shifted in each period are fed directly to the output along with the parity bit(s). The parity bit(s) are generated according the code generator using exclusive-or gates. A block diagram of the encoder is shown in Figure 5.14.

The CSOC decoder makes use of several shift registers and threshold circuits. A block diagram is shown in Figure 5.15.

5.5.5 Interleaving

Most FEC coding methods assume random occurences of bit errors. In the case of transmission channels that result in burst errors care must be taken to ensure that the length of error bursts do not tend to exceed the capacity of the coding method. This can imply the use of multiple-error correcting codes, the use of burst correcting codes or the use of interleaving.

The benefits of interleaving are derived from the fact that all practical coding methods have finite memory. The purpose of interleavers is to separate large error bursts from the channel into smaller (typically single bit) burst

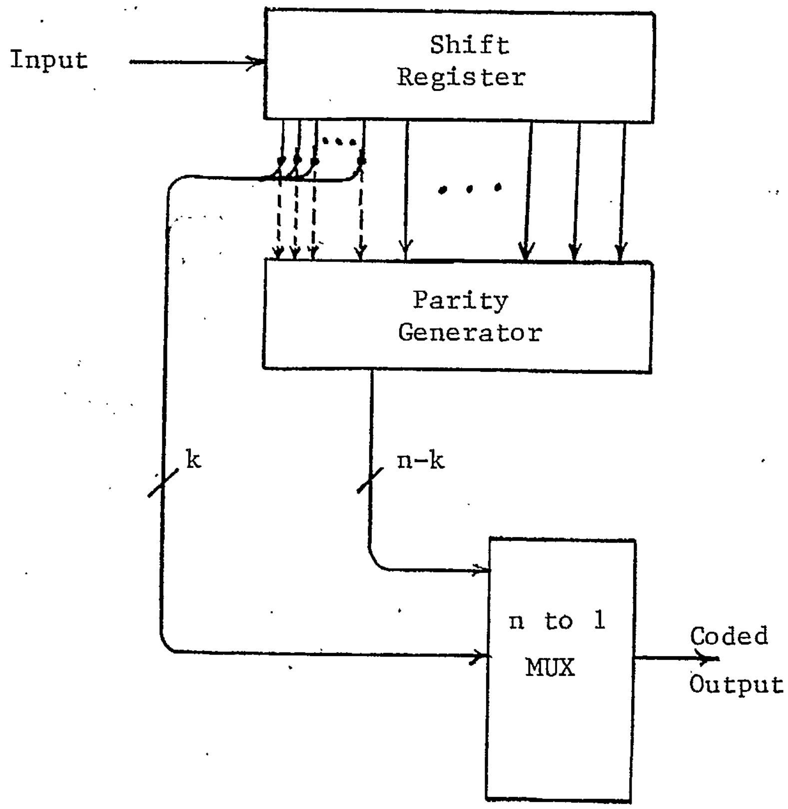


Figure 5.14. CSOC Encoder

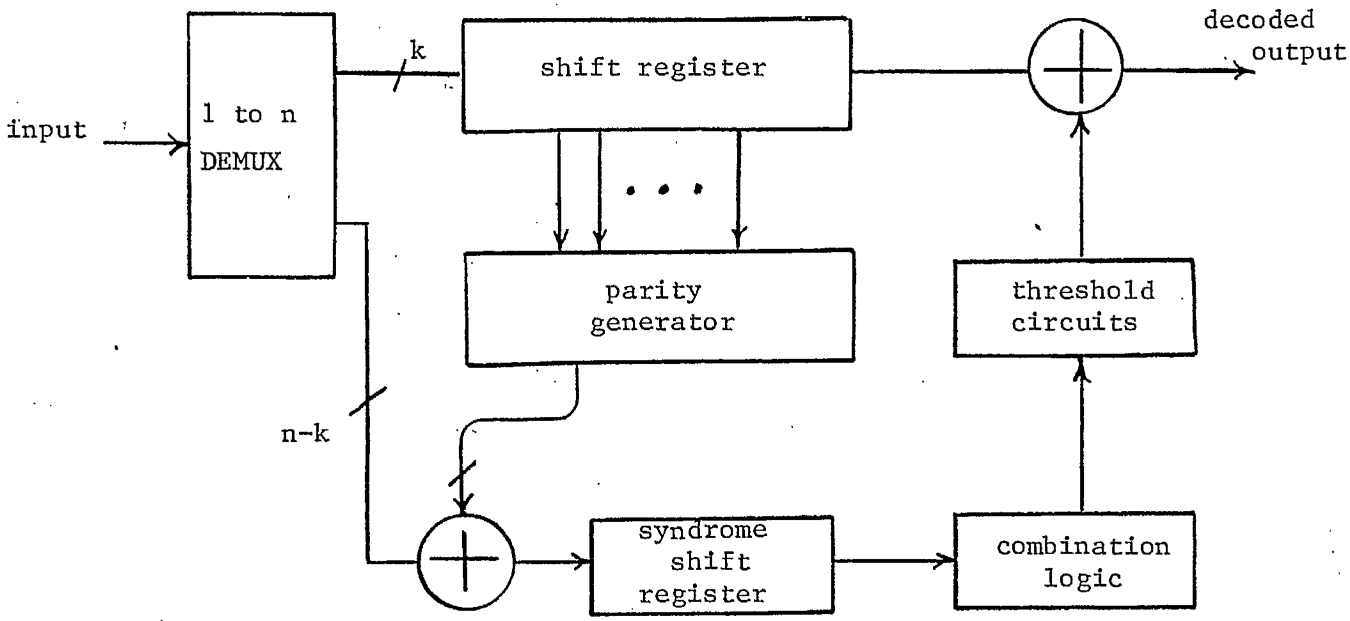


Figure 5.15 - Threshold Decoder

errors where each of the smaller bursts is seperated by a number of bits greater than the memory length of the coding scheme.

[54] investigated several types of interleaving including: Block Interleaving, Helical Interleavers and Convolutional Interleavers. Of these Block Interleaving is the most common and easiest to understand.

Block Interleaving requires no initialization or start up bits, as do the Helical and Convolutional interleavers. Of course this is not critical for the projected application (continuous transmission).

The Helical and Block interleavers are implemented using random access memory (RAM) and a commutator device. The Helical interleaver requires RAM of size $N(N - 1)/2$ while the Block interleaver requires RAM of size NB, where:

> N is the block length of the interleaved words (i.e. block length of code)
> B is the interleaving depth of the interleaver (i.e. number of bits seperating interleaved bits).

The convolutional interleaver can be implemented by shift registers, see Figure 5.16. The parameter M is given by,

$$N = MB \tag{5.23}$$

where N and B are as defined above.

5.5.6 Phase Ambiguity

When BPSK modulation is used, the potential exists for a 180° phase ambiguity in the polarity of the received

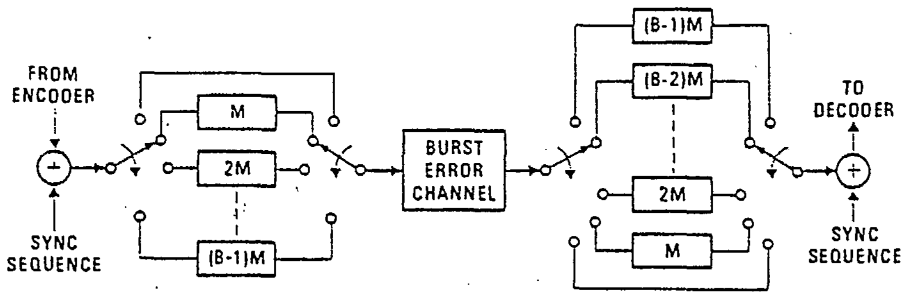


Figure 5.16 : Shift register implementation of convolutional interleaver/deinterleaver [5.7]

channel symbols. This problem can be solved in one of two ways for convolutional codes. First one could use a so-called transparent code, which is transparent to phase inversions. An inversion in the polarity of the decoder input simply produces an inversion in the polarity of the decoder output. When this type of code is used differential encoding/decoding is used outside of the encoder to ensure the proper polarity for decoded date. Note that from the form of this decoder, a signal decoding error produces 2 errors at the differential decoder output. For an AWGN channel this results in a loss of about 0.1 to 0.2 dB.

When non-transparent codes are used a change in polarity at the decoder input results in a word that is not a legitimate code word. As a result, the received data appears to have an excessive error rate and thus the condition is easily detected. For this case there is no performance loss under typical operating conditions ($P_b \approx 10^{-5}$).

One way of combatting phase ambiguity problems with the Golay (24,12) code (or any block code where code word complements are also code words) is to use one less information bit per block, i.e. use only half the code words by eliminating the complement of each remaining code word. The code rate is then reduced slightly, e.g. a (24,11) code results. Consistent block errors in the decoder indicate that something is wrong, and assuming correct timing (i.e. block synchronization), appropriate action can be taken.

With BPSK a quick fix for both block and convolutional codes is to differentially encode/decode the data stream at the channel (inside the FEC codec), perform interleaving (to separate paired errors) and then decode. However, the

E_b/N_o degradation resulting from an effective doubling of random errors can easily be 1 to 2 dB at high error rates expected at the input to the decoder, which is much worse than the solutions stated above.

The phase ambiguity problem is avoided with DPSK signalling. For an AWGN channel the degradation from ideal BPSK is inherent with the DPSK demodulator. Interleaving is also recommended to separate partially correlated errors in adjacent demodulated bits.

5.5.7 Synchronization Aspects

The entire Radio Program Receiver must be synchronized to the incoming data. The demodulator provides bit synchronization for itself and the other units (i.e. provides a clock signal). The FEC decoder must be synchronized to the FEC encoder in order to decode the data. The demultiplexor and source decoder must also be synchronized to the corresponding transmitter units. in order to minimize synchronization overhead it is desireable to, in some way, use the same sycnchronization information for both the FEC decoder and demultiplexor (the source decoder is inherently synchronized to the demultiplexor).

5.5.7.1 BCH Code

If a BCH code is chosen so that its block size corresponds to the multiplexor frame size then synchronization is simple*. A unique word (or synchronization word) is placed at the beginning of each frame. At the receiver a unique word deector is placed before the decoder providing a sync pulse to both the decoder and demultiplexor.

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*This would also work if the multiplexed frame size was a sub-multiple of the code block size and the unique word is provided at the beginning of each coded block.

An alternative method for mutual synchronization would be to use the information provided by the decoder itself. The average output bit error rate could be determined by averaging the number of bits in error over several blocks by means of an accumulator. If the average is less than some threshold (determined by the theoretical performance) then sync is declared. If not, a bit is deleted from the the input bit stream and the next group of blocks is checked. This is repeated until the resultant average bit error rate meets the threshold and sync is declared. Using this method the FEC decoder can provide the demultiplexor with a sync pulse and a lock indication.

The first method is clearly easier to implement. It will require some overhead but it is expected to be minimal. The second method will require increased complexity (although with a microprocessor implementation it should be minimal). The big disadvantage of the second method is the inherently long lock-up time. For example, if the block size is 255 and the averaging period 5 blocks, then on average delay of 1.78×10^6 bits or approximately 1 second delay results. This will require at least 2 seconds of dummy data at the start of transmission and if sync is lost an average 1 second of data is lost. A thorough investigation must be made on the statistics (mean time to lock, mean time to lose sync) for this second scheme could be considered.

5.5.7.2 CSOC Codes

CSOC codes are systematic and, as such, similar techniques to those used by BCH codes can be used.

A unique word synchronization circuit could be used to provide sync pulses to the demultiplexer within the decoder. A unique word is not required every block (or constraint length). In any event the decoder will output incorrect data for a length of time equivalent to the memory length of the code after regaining sync. The same sync information can be used for the data demultiplexor if the frame size is a multiple of the constraint length.

A similar technique to the second method for BCH codes can also be used. A BER monitor and synchronization circuit as described in [55] could be used. In this way sychronization could be established for the demultiplexor as well (if the frame size of the demultiplexor is a submultiple of (or equal to) the constraint length of the code).

Similar comments apply as for BCH codes. The first method is preferred from the viewpoint of complexity while the second method is preferred from the viewpoint of bit effeciency.

5.5.7.3 Viterbi Decoders

Viterbi decoders normally utilize some form of internal BER measurement circuitry to detect synchronization and synchronization circuitry to remedy it. Given that Viterbi decoding normally use non-systematic codes the synchronization information is not easily translated to information useable by the demultiplexor. This especially applies to commercially available Viterbi codecs where this information will not be available at the output of the decoder.

5.5.8 Error Detection

Since the source decoder requires error detection information for error concealment, it would be desireable to utilize the error detection ability of the FEC codes.

The general method for doing this would be to correct all error patterns that are guaranteed correctable by the code. No other error patterns are corrected, instead the error indication would be made available to the source decoder.

The problem with the above method is that the blocks of the code would have to correspond to the PCM code words on a one-to-one basis (or possibly one PCM word from each audio channel in a block). This implies that only short block codes could be used. The only practical code would be the Golay code. This takes 12 bits of uncoded data (i.e. full 12 bit PCM word or two 6 bit partial PCM words).

5.6 Survey of Coding Equipment Manufacturers

Information was obtained from the following sources:

M/A-COM Linkabit, Inc.
3033 Science Park Road, San Diego, California
92121
Tel: (619)457-2340

Harris
P.O. Box 1700, Melbourne, Florida, 32901
Tel: (305)724-3932

Cyclotomics
2120 Haste St., Berkeley, California, 94707
Tel: (414)548-1300
Tel (East Coast): (703)356-6736

Digital Communications Corporation
19 First field Road, Gaithersburg, Maryland, 20760
Tel: (301)948-0850

Space Research Technology
17317 El Camino Real, Houston, Texas, 77058
Tel: (713)480-3086

Scientific Atlanta:
Tel (Mississauga Office): (416)677-6555
Tel (Atlanta Office) : (404)449-2025

Aydin Monitor Systems
Tel (Brampton, Ont. rep.): (416)459-4397

Linkabit Corporation

Linkabit offers a number of encoders/decoders with various code rates, maximum input data rates and decoding techniques. All use convolutional schemes. The unit suitable for the Radio Program Receiver is described below.

LV7017B:

This is a constraint length 7 convolutional encoder and Viterbi decoder which can operate at data rates up to 10 Mbps (at rate 1/2). Code rates available include 1/3, 1/2, 3/4. An optional interleaver/deinterleaver provides burst error correction capability. The interleaving depth is 512. The cost of a single device is approximately $15,000 US (quantity 1) or $10,000 US (quantity 100).

Harris Corporation

Harris Corporation can provide a rate 1/2, k = 7 convolutional encoder/Viterbi decoder using soft decisions. It is built from 60-70 IC's and would cost in the neighbourhood of $5,000. They also have a K = 3 rate 1/2 single chip Viterbi decoder which was part of a military system they developed. It should be available for commercial use in the near future at somewhere around $500 per chip in small volumes. The military system itself utilized interleaving but this was not built into the chip. The maximum speed would be about 20 kbps and therefore the two decoders are not suitable.

Harris is presently developing a 10 Mbps, rate 1/2, K = 7 single chip convolutional encoder/Viterbi decoder. It has not yet been tested but hopefully would be available in the near future. The characteristics of the 10 Mbps coded will be similar to the characteristics of the unit presently available for up to 256 kpbs. The 256 kbps unit has a guaranteed gain of 3.3 dB over ideal uncoded PSK performance when used with a Harris BPSK or QPSK modem or about 5.3 dB over the actual guaranteed uncoded performance (see Figure 5.17).

The cost of the high rate codec is expected to be in the range of $3000-4000 US.

Cyclotomics Inc.

Cyclotomics presently offer coding devices in several code rates and data rates. All codecs use Reed-Solomon coding methods and are therefore not suitable for the Radio Program Receiver.

Digital Communications Corporation

No information has been obtained to date.

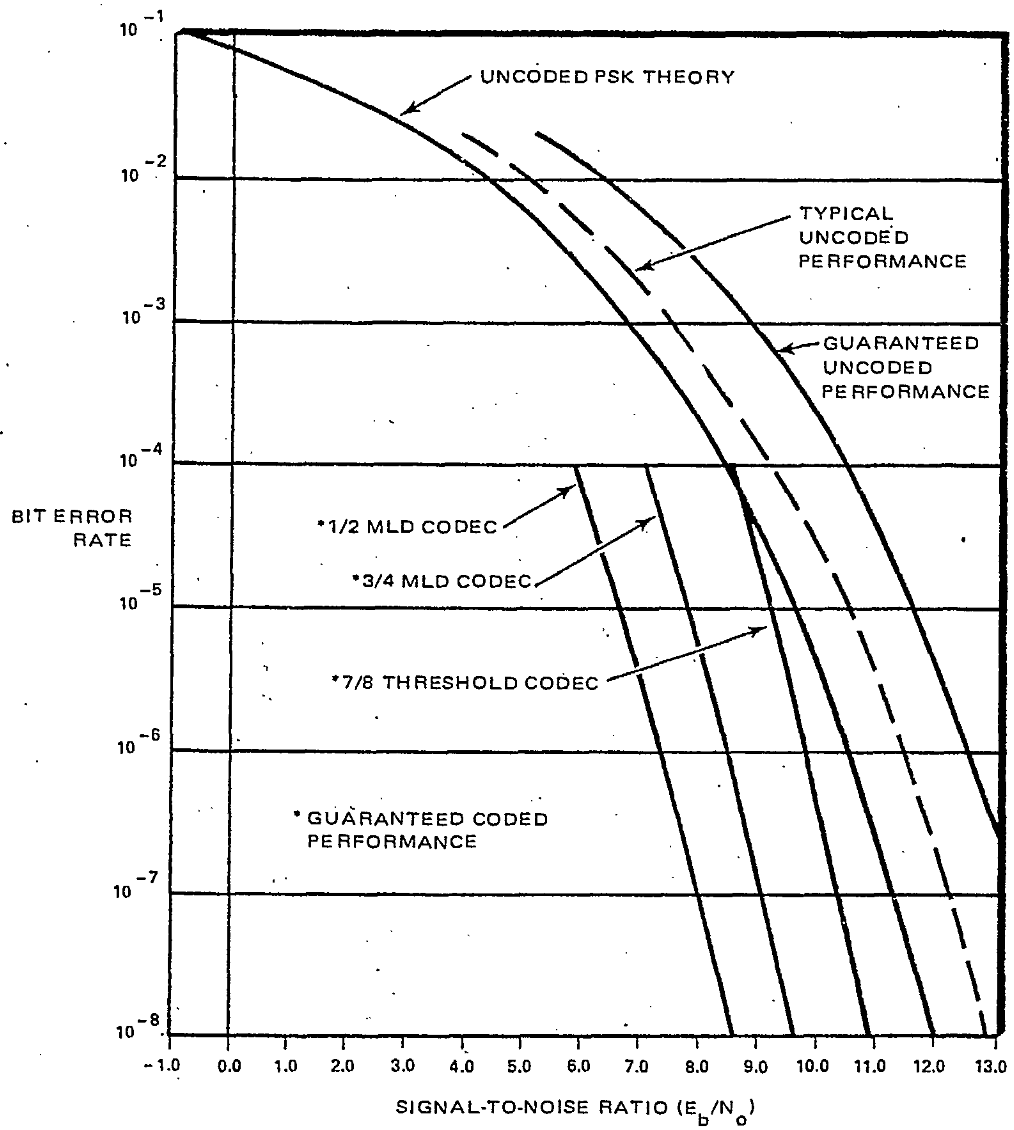

Figure 5.17 - Performance of Harris
256 kbps modem/codec

Space Research Technology

Space Research Technology has developed a rate 1/2 codec. The code used is a (24,12,3) Block code. A VLSI version of the codec is being developed and should be in volume production by the summer of 1985. These units operate up to 2.5 Mbps and have a coding gain of 2.9 dB at 10^{-7}. They will cost about $500 U.S. per chip or $1200 U.S. for the complete module on a board.

Scientific Atlanta

Scientific Atlanta offer a rate 7/8 CSOC code/threshold decoding (with feedback) codec with their 8800 series of modems. The data rates available range from 56 kbps to 10 Mbps (for QPSK). The theoretical gain at 10^{-7} is 3.3 dB for this codec.

Aydin Monitor Systems

Aydin offer a rate 7/8 CSOC/threshold decoding codec with their 2700 series of modems. The data rates available for the model 2707 are 1 - 50 Mbps (models with lower rates are available). The total cost of the 2707D (demodulator/ Decoder) is about $20,000 CDN (without decoder the cost is about $17,000 CDN).

5.7 Coding Scheme Selection

In Section 5.3, the list of coding schemes was narrowed down to four:

(1) BCH codes

(2) The Golay code

(3) Convolutional coding with Viterbi Decoding

(4) CSOC codes with Threshold Decoding

Section 5.4 and 5.5 each of these schemes were examined more closely. A summary of the results of these sections as well as other aspects is included in Table 5.3.

Characteristic	Coding Technique			
	BCH	Golay	Viterbi	CSOC
(1) Code Type	Block	Block	Convolutional	Convolutional
(2) Decoding Technique	Berlekamp-Massey-Chien	Look-up Table	Viterbi (soft or hard)	Threshold
(3) Practial Code	~1/2	1/2 (for(24, 12) Extended Golay)	1/2, 3/4	1/2, 3/4
(4) Coding Gain (@ BER = 10^{-7})	3.4 - 5.0 dB	2.4 dB	4.0 - 6.0 dB for rate 1/2, soft decisions 3.6 - 4.2 dB for rate 1/2, hard decisions	2.3 - 3.3 dB for rate 1/2
(5) Synchronization	requires external synchronization (i.e. unique word)	same as BCH	self-synch-ronizing	can be self-synchronizing
(6) Phase ambiguity	must be resolved before decoding or must use 1 bit less per block	same as BCH	resolved by transport code or extra circuitry	same as Viterbit

Table 5.3 (a) - Summary of Code Characteristics

Characteristic	Coding Technique: BCH	Golay	Viterbi	CSOC
(7) Error Detection Capability	not practical	simple to implement	not practical	not practical
(8) Complexity	complex	very simple	complex	simple
(9) Commercial Availability	code using (24,12) block code similar to Golay code available		one definite supplier, another possible	several available at rate 7/8 (use feedback decoding)
(10) Commercial Price (U.S.)	$1200 (large quantities)		$15,000 (quantity 1) $10,000 quantity 100) possibly $3000 - $4000 from other supplier	$3000 when bought with $17,000 demodulator
(11) Integration into system	-will require unique detector before decoder -same sync information can be used by demultiplexor	-can be integrated with error detection function (i.e. placed after demultiplexor) -will require sync information (same as demultiplexor)	-may require deinterleaver after decoder -synchronization independent of demultiplexor	-if built could be self-synchronizing with information available to demultiplexor -if bought then sync information not available

Table 5.3 (b) - Summary of Code Characteristics

6.0 CONCLUSIONS

In this report, the performance tradeoff and the implementation considerations of various digital techniques used for (1) source coding, (2) multiplexing, (3) modulation, and (4) channel coding were presented.

As far as the source coding techniques are concerned consideration was given mainly to PCM derived techniques. The instantaneous companding (ICOM) appears to have an edge over the near-instantaneous (NICOM) techniques. The PCM technique employing 14 bit to 11 bit instantaneous A-law companding is a CCITT standard. Codec implementation is simple and straightforward. In NICOM different error correction techniques are suggested for bits of different importance. Thus, the error correction implementation is relatively complicated and the use of highly powerful FEC codes becomes difficult. In ICOM, except for the 1 bit parity protection, the error correction (FEC) is done separately and is independent of the source codec. This enables one to use commercially available FEC codecs to achieve large coding gain. Larger coding gains are important because the transponder is power limited. For the same reason, the higher bit rate compression achieved in NICOM as compared ICOM has less practical significance. While the multiplexed NICOM channels easily interface with the 2.048 Mbps hierarchy, ICOM channels can easily interface with 1.544 Mbps hierarchy. Further digital conversion of the 14 bit to 11 bit A-law compressed data stream to a 15 bit to 11 bit μ-law compressed data stream and the reverse process can be hardware implemented relatively easily. These advantages coupled with a North American market target heavily favour the 14 - 11 - ICOM-PCM technique. The implementation of this technique was discussed in significant detail.

The relative merits of continuous and packet multiplexing techniques were assessed and the former approach was chosen for detailed investigation. The channel multiplexer and demultiplexer will have design features suitable to accommodate different program (15 kHz, 7.5 kHz) and data channels. The T1 link mux and demux can be used in conjunction with the channel mux-demux to backhaul the digitized program from the studio center to the earth station. The implementation aspects of the channel mux-demux were presented.

Several modulation techniques (BPSK, QPSK, OQPSK, MSK, Duo MSK, TFM, etc.) were considered. For multicarrier operation, the transponder has to be operated at several decibels backoff (i.e. operation is nearly linear). The transponder is power limited rather than bandwidth limited. Because of this, highly bandwidth efficient, constant envelope schemes (Duo MSK, TFM) which have relatively poor power efficiency and higher implementation complexity were not considered for further investigation. Although the implementation of differentially demodulated schemes is relatively simple, those schemes normally perform inferior to their coherently demodulated counterparts. The remaining coherently detected BPSK, QPSK, OQPSK and MSK schemes were compared.

The primary criteria of comparison were the probability of error performance and the implementation complexity. Based on this comparison the choice has been made on QPSK (Note: BPSK has similar performance and implementation complexity). Computer simulations were performed to select a practical filter combination (transmit and receive). Filter bandwidth was optimized by minimizing the E_b/N_o requirement. Simulation results were also presented for the case when the desired channel is placed inbetween two identical interfering adjacent channels. The performance degradation in the desired channel was evaluated for different channel spacings and for the optimized filter bandwidth. The implementation schemes of BPSK and QPSK were presented.

As far as forward error correction is concerned, a high coding gain is of primary importance. As the channel is largely power limited a coding gain of at least 2.5 dB has been found necessary when a typical earth station receiver (3M, 120°K LNA) is used. A coding gain of 4 to 5 dB will enable the transponder to deliver a larger number of audio program channels (15 kHz mono) using the digital technique as compared to the number of channels using the analog FM technique. The code rate is not of great importance especially when the QPSK technique is used. Both block and convolutional coding and several decoding schemes were compared on the basis of their performance and complexity. Several commercially available codecs (threshold, Viterbi, Sequential) can meet the coding gain requirement, the cost slightly increasing with the increase of coding gain.

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APPENDIX 2A

Tau-Tron Digital Program Channel Modules

tau-tron
A UNIT OF
GENERAL SIGNAL

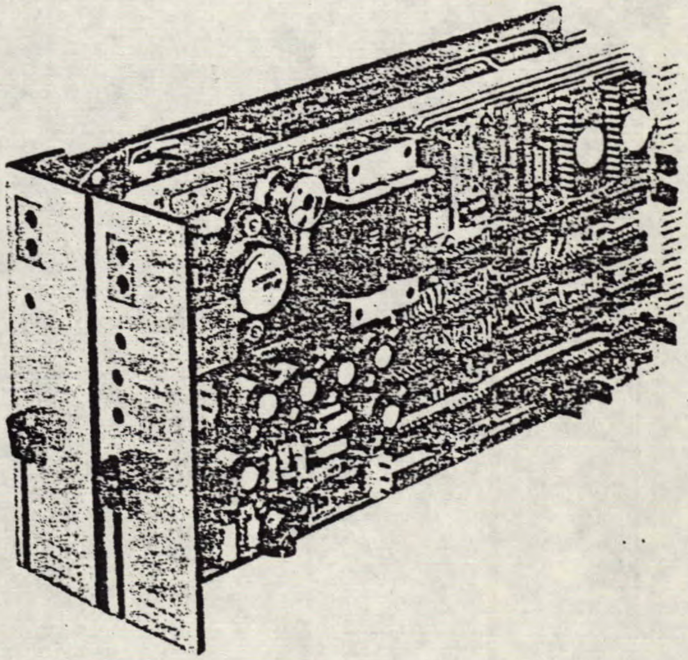
intraplex

PT-15 Transmitter
PR-15 Receiver

15 kHz Digital Program Channel Modules (TDM-150 System)

The PT/PR-15 digitally encoded 15 kHz audio program channels are plug-in units for use with Tau-tron's versatile INTRAPLEX TDM-150 Digital Multiplexer which allows the transmission of high quality program over T-carrier systems. Up to four channels can be inserted, dropped, or originated in one- or two-way configurations.

- **Equipment and Monitor Bantam Jack Access Points**
- **Receive Level Adjustment**
- **Mute and Parity Error Indicators**
- **Emphasis Selectable**
- **High Performance 15 kHz Program Bandwidth for AM/FM, Stereo/Monaural Applications**
- **Drop and Insert using Existing T1 Transmission Paths while Maintaining Integrity of Existing Circuits**
- **High Resolution Linear Coding with Digital Companding. Error Mitigation to Effectively Remove Clicks Associated with Digital Errors**
- **Flexible Channel Assignments Compatible with D1D, D2, D3 or D4 Channel Banks**
- **Simple Installation and Maintenance and Complete T1 System Compatibility**



PT-15 PR-15

SPECIFICATIONS

BANDWIDTH	40 Hz to 15 kHz
TRANSMISSION LEVELS	+21 dBm full-load levels at input and output. Transmit gain range adjustable in 0.2 dB increments from -1 to +14 dB. Receive gain range adjustable -6 dB to 0 dB.
IMPEDANCE	600 ohms or 150 ohms, balanced and floating for input and output. Return loss 26 dB, 50 Hz to 5 kHz and exceeding 20 dB, 40 Hz to 15 kHz.

LOGITUDINAL BALANCE	40 Hz to 4 kHz More than 70 dB 4 kHz to 7 kHz More than 65 dB 7 kHz to 15 kHz More than 60 dB
DIGITAL PROCESSING	32,000 samples/s, 384 kb/s transmission rate. 14-bit coding digitally companded to 11 bits/sample plus 1 bit for error mitigation.
FRAMING	D1D, D2, D3, D4 compatible formats. 6 VF channels per 15 kHz program channel.
TOTAL DISTORTION	Less than 0.5%, 40 Hz to 125 Hz, less than 0.35%, 125 Hz to 15 kHz, for -6 dBm0 to +18 dBm0 sinusoid input.
TOTAL HARMONIC DISTORTION	Less than 0.4%, 40 Hz to 125 Hz, less than 0.3%, 125 Hz to 15 kHz, for -6 dBm0 to +18 dBm0 input.
QUANTIZING NOISE	Signal-to-quantizing ratio more than 52 dB for a -6 dBm0 and +18 dBm0 sinusoid at 1004 Hz measured in a flat 15 kHz band.
INTERMODULATION DISTORTION	Total intermodulation distortion less than 0.2% with simultaneous +12 dBm0 inputs at 500 Hz and 2 kHz.
OVERLOAD	Total distortion at +21 dBm0 less than 2.0% from 40 Hz to 15 kHz. Recovery from a 27 dBm0 signal at 1004 Hz is less than 100 ms.
IDLE-CIRCUIT NOISE	More than 76 dB below full-load level measured in a flat 15 kHz band.
SINGLE-FREQUENCY INTERFERENCE	40 Hz to 1 kHz Less than 65 dBm0 1 kHz to 12 kHz Less than 75 dBm0 12 kHz to 15 kHz Less than 65 dBm0 With audio input terminated.
SPURIOUS TONES	Spurious audio output tones other than the fundamental and harmonics for inputs from 20 Hz to 20 kHz and -5 dBm0 to +18 dBm0. 0 to 300 Hz Less than 55 dB below input 300 Hz to 1 kHz Less than 60 dB below input 1 kHz to 12 kHz Less than 65 dB below input 12 kHz to 15 kHz Less than 60 dB below input Above 15 kHz Less than 55 dB below input
FREQUENCY RESPONSE	40 Hz to 125 Hz +0.3 dB to -0.7 dB 125 Hz to 10 kHz ±0.3 dB 10 kHz to 14 kHz +0.3 dB to -0.7 dB 14 kHz to 15 kHz +0.3 dB to -1.0 dB
LEVEL TRACKING	Output level tracking is within ±0.2 dB for input levels from -20 dBm0 to +18 dBm0.
GROUP DELAY	Less than 4 ms, 40 Hz to 15 kHz with respect to in-band minimum.
INTERCHANNEL GAIN DIFFERENCE	40 Hz to 125 Hz Less than 0.4 dB 125 Hz to 10 kHz Less than 0.3 dB 10 kHz to 15 kHz Less than 0.8 dB
INTERCHANNEL PHASE DIFFERENCE	40 Hz to 125 Hz Less than 6 degrees 125 Hz to 10 kHz Less than 3 degrees 10 kHz to 15 kHz Less than 6 degrees
CROSSTALK	Attenuated by more than 75 dB, 40 Hz to 15 kHz within a pair of associated terminals.
ERROR MITIGATION	Single-bit parity for 7 most-significant sample bits. Previous sample held for parity error. Muting for parity error in a majority of consecutive samples.
SURGE PROTECTION	Audio input and output tolerance tip to ring, tip and ring to ground greater than 600 V transient in 10 μs maximum decaying to 300 V in 1 s minimum.
SYSTEM COMPATIBILITY	TDM-150 system compatible. Full-width plug-in. Power consumption 3.5 w, PT-15; 3.3 w, PR-15. Time slots manually selectable for 1 thru 6, 7 thru 12, 13 thru 18 or 19 thru 24.
ENVIRONMENTAL	Full performance 0° to 40° C. Operational to 50° C.

tau-tron
A UNIT OF
GENERAL SIGNAL
Outside Mass., 1-800-TAU-TRON;

Tau-tron Inc.
27 Industrial Ave.
Chelmsford, MA 01824
(617) 256-9013
Telex 750245

APPENDIX 2B

Bayly Engineering Digital Audio Program Terminal

OMNIPLEXER TECHNICAL SPECIFICATION: T1 PROGRAM (15 kHz)

The Omniplexer specifications comply with the criteria outlined in AT&T. Technical Advisory 74, Issue 1, December, 1982. They are repeated here for convenience. 7.5 kHz Program is as in T.A.79, Sept. 1983.

Frequency Range	:	40 Hz - 15 kHz
Saturation Level	:	Output - adjustable +15 dBm to +21 dBm Input - adjustable -5 dBm to +22 dBm Nominal - +21 dBm
Clipping Level	:	1.2 dB above saturation level
Impedance	:	Input and Output - switchable 600 Ohms/150 Ohms balanced and floating
Return Loss	:	Input - ≥ 30 dB, 40 Hz - 15 kHz Output - ≥ 26 dB, 40 Hz - 10 kHz ≥18 dB, 10 kHz-15 kHz
Balance	:	40 Hz - 4 kHz - ≥ 70 dB 4 kHz - 7 kHz - ≥ 65 dB 7 kHz - 15 kHz - ≥ 60 dB
Protection	:	Input and Output audio lines protected against 600V surge with 10 sec rise time and ≥ 1 msec fall time to 50%
Level Stability	:	≤ ±0.1 dB variation (24 hours)
Idle Channel Noise	:	≤ -55 dBm0, 15 kHz flat
Spurious Tones	:	The level of all spurious tones at the following frequencies when a single tone from 20 Hz to 20 kHz at -5 dBm0 to +18 dBm0 is applied to the channel: 20 Hz to 300 Hz - ≥ 55 dB below the applied tone 300 Hz to 1 kHz - ≥ 60 dB below the applied tone 1 kHz to 12 kHz - ≥ 65 dB below the appled tone 12 kHz to 15 kHz - ≥ 60 dB below the applied tone >15 kHz - ≥ 55 dB below the applied tone
Overload Recovery:	:	Instantaneous from +24 dBm ≤ 100 μsec from +27 dBm

Gain/Frequency Response	:	40 Hz - 300 Hz - +.25 dB to -.67 dB 100 Hz - 10 kHz - ±.25 dB 10 kHz - 14 kHz - +.25 dB to -.67 dB 14 kHz - 15 kHz - +.25 dB to -0.75 dB
Envelope Delay Variation	:	40 Hz - 15 kHz ≤1.8 mSec.
Single Frequency Interference	:	40 Hz - 300 Hz - ≤ -55 dBm0 300 Hz - 1 kHz - ≤ -65 dBm0 1 kHz - 12 kHz - ≤ -75 dBm0 12 kHz - 15 kHz - ≤ -65 dBm0
Harmonic Distortion	:	40 Hz-125 Hz - ≤0.4%, -20 dBm to +18 dBm 125 Hz-15 kHz - ≤0.3%, -20 dBm to +18 dBm 40 Hz-15 kHz - ≤1.5% at +21 dBm
Intermodulation Distortion	:	≤0.1%, 500 Hz & 2 kHz each at +12 dBm0
Total Distortion Ratio	:	≥49 dB, 125 Hz to 15 kHz and -6 dBm0 to +18 dBm0.
Signal to Quantizing Ratio (15 kHz flat)	:	≥52 dB, 1004 Hz, at -6 dBm0 to +18 dBm0
Audio Pre/De-emphasis Options	:	1) None (flat) 2) CCITT J.17 (6.5 dB loss at 800 Hz)
Access Jacks	:	Terminated and Bridged audio bantam jacks
Level Tracking Error	:	±0.2 dB, -10 dBm to +10 dBm
Interchannel Gain Difference	:	≤ .4 dB, 40 Hz - 125 Hz ≤ .2 dB, 125 Hz - 10 kHz ≤ .4 dB, 10 kHz - 15 kHz
Interchannel Phase Difference	:	From 5° at 40 Hz to 3° at 200 Hz From 3° at 200 Hz to 5° at 15 kHz

Interchannel Crosstalk Attenuation	:	$\geq$ 75 dB, 40 Hz to 15 kHz
Error Mitigation	:	Parity Detection/Level Hold 1 bit parity per T.A.74
Compression Law	:	14/11 A-law (A = 43.8)
Power Requirements	:	-42 to -60 VDC (-48VDC nominal)
Battery Noise Level Accepted	:	56 dBrnC, or single frequency of 100mV rms, 10 Hz - 20 MHz
Operating Temperature	:	0^{o}C to $+40^{o}$C to meet spec. 0^{o}C to $+50^{o}$C operational
Relative Humidity	:	Up to 55% for continuous operation. Up to 80% short term exposure.
Alarm	:	The following alarms and functions are provided: 1) Red Alarm 2) Yellow Alarm 3) PCM Simulator Fail 4) Channel Alarm 5) Alarm Cutoff 6) Bypass Alarm 7) Channel Mute

Alarm Contacts (IC) are provided for the following:

1) Red Alarm
2) Yellow Alarm
3) Channel Alarm (1 per channel)

<u>1.544 MB/sec Line Interface</u>

<u>INPUT</u>

Line Rate	:	1.544 Mbits/sec $\pm$130 ppm
Line Code	:	Bipolar AMI
Pulse Amplitude	:	Nominal 3V peak signal per Technical Advisory #34, with attenuation equivalent to 0-750 ft. of inter-office cabling, or with over-equalization equivalent to 0-750 ft. of inter-office cabling.

Input Impedance	:	100 Ohms nominal at 772 kHz, or 1000 Ohms
Return Loss	:	$\geq$16 dB, 50 kHz to 1.544 MHz @ 100 Ohms
Jitter Acceptance	:	Per PUB 41451

<u>OUTPUT</u>

Line Rate	:	1.544 Mbits/sec $\pm$30 ppm in the stand-alone mode
Line Code	:	Bipolar AMI
Pulse Amplitude	:	3.0V $\pm$0.3V peak, with plug-in equalizers for 0-150 ft., 150-450 ft., and 450-750 ft.

APPENDIX 2C

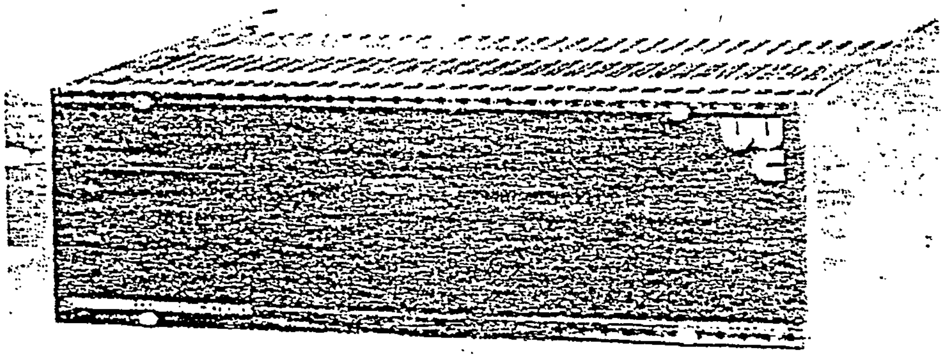
Wegener Communications Series 1600 Audio SCPC Products

WEGENER
COMMUNICATIONS

TECHNICAL
DATA

SERIES 1600 AUDIO SCPC PRODUCTS

The Wegener Series 1600 Audio SCPC Products are modular components which install in the Series 1600 Mainframe. Configuration of the equipment is determined by user requirements. The system utilizes the PANDA™ II noise reduction system to provide the end user with a transparent audio link.



Series 1600 SCPC Receiver

EQUIPMENT FEATURES

- PANDA™ II noise reduction provides a transparent audio link
- High transmission efficiency minimizes transponder costs
- Audio channels available for 15 kHz and 7.5 kHz bandwidth applications
- Modular packaging for easy service expansion

EQUIPMENT DESCRIPTION

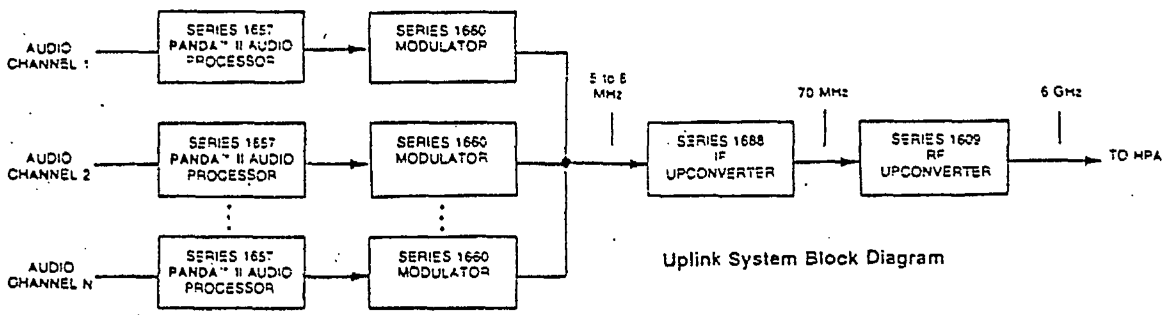
The Wegener Communications SCPC systems provide a transparent audio link utilizing the PANDA™ II noise reduction system.

Source audio is processed by the Series 1657 PANDA™ II Audio Processor and is modulated on an FM carrier by the Series 1660 Modulator. The modulator output is upconverted first to 70 MHz IF by the Series 1688 IF Upconverter and then to RF by the Series 1609 RF Upconverter. The 6 GHz RF output is routed to an HPA for transmission.

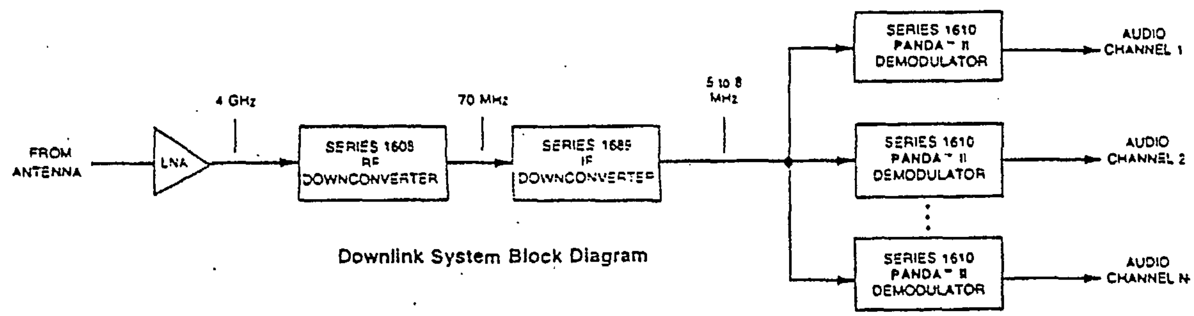
The received signal is amplified by an LNA and downconverted to 70 MHz IF by the Series 1608 IF Downconverter. The IF signal is downconverted to the 5-8 MHz spectrum by the Series 1689 Downconverter. The signal is then demodulated and processed by the Series 1610 PANDA™ II Demodulator.

SPECIFICATIONS

Audio Characteristics		
Channel Bandwidths	15 kHz and 7.5 kHz	
Frequency Response	±1.0 dB (50 Hz – 7.5 kHz)	
	±1.0 dB (50 Hz – 15 kHz)	
THD at Peak Deviation	< 0.5% (125 Hz – 15 kHz)	
	< 1.0% (50 Hz – 125 Hz)	
Peak Program Level	− 18 dBm	
Audio Signal to Noise Ratio	> 90 dB, relative to + 18 dBm program level	
Audio Interface	600 ohms, balanced	
Modulation Characteristics	<u>15 kHz Channel</u>	<u>7.5 kHz Channel</u>
Deviation	50 kHz peak	25 kHz peak
Occupied Bandwidth	130 kHz	65 kHz
Receive Noise Bandwidth	150 kHz	75 kHz
Channel Spacing	180 kHz	90 kHz
C/No, Operational	66 dB	64 dB
C/No, Degraded	60 dB	58 dB
Power Requirements		
Standard Power Supply		
Voltage	105 – 130 VAC	
Frequency	47 – 63 Hz	
Optional Power Supplies		
	210 – 260 VAC	
	− 42 to − 56 VDC	
	− 20 to − 28 VDC	
Mechanical		
Size	5¼"H x 19"W x 13"D	
Weight	20 lbs.	



Uplink System Block Diagram



Downlink System Block Diagram

For further information and application assistance, contact:

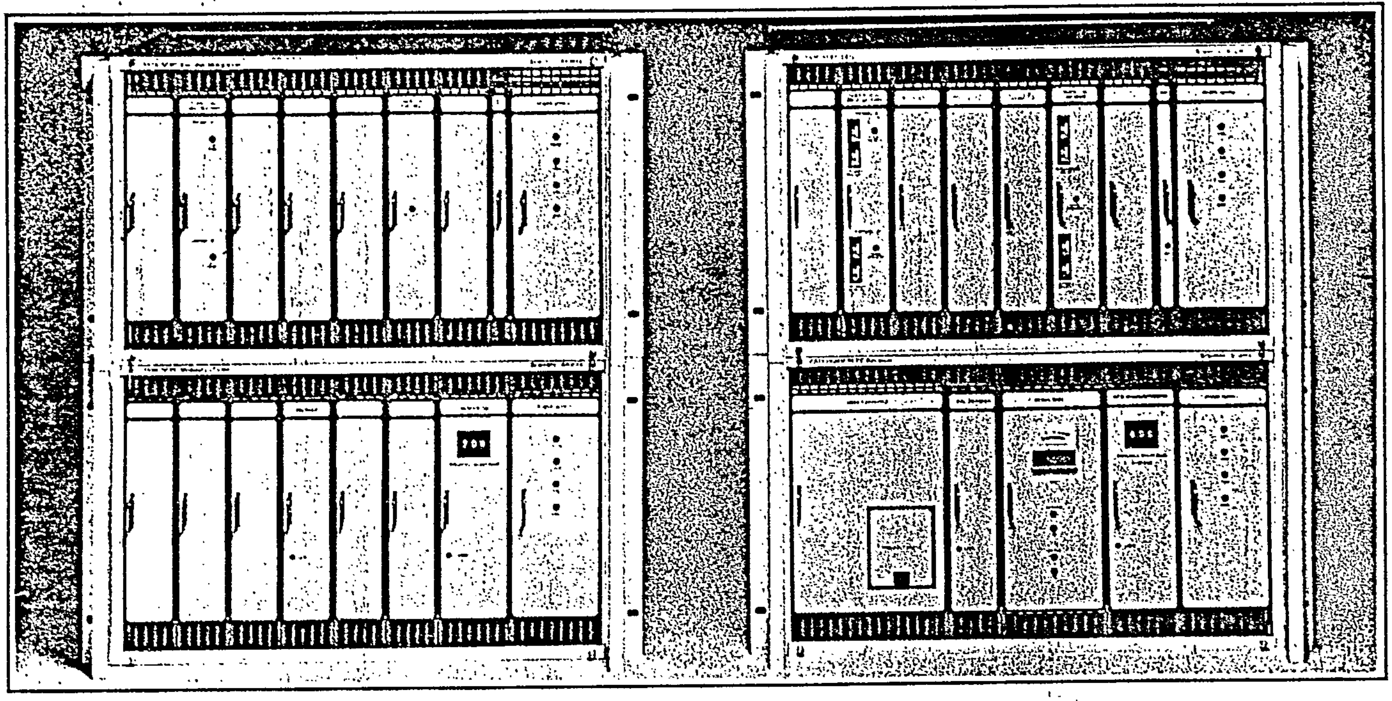
WEGENER COMMUNICATIONS
150 TECHNOLOGY PARK/ATLANTA
NORCROSS GEORGIA 30092

1600-049-984

APPENDIX 2D

Scientific Atlanta DAT-800 SCPC Receive Terminal

DAT-800 SCPC



System Description

The Scientific-Atlanta DAT-800 SCPC Program Audio and Data Distribution System facilitates the relay of media information in digital format over satellite channels. Operating at 800 kb/s, a single digital SCPC (Single-Channel-Per-Carrier) channel accommodates several combined broadcast audio and data services:

- Digital Program Audio (15 kHz or 7.5 kHz) Channels
- Voice Cue (32 kb/s) Channels
- Data Channels

The system utilizes a bi-phase shift keying (BPSK) carrier modulated by an 800 kb/s Time Division Multiplex (TDM) digital data stream. Its receiver unit demodulates and demultiplexes this data stream into separate audio and/or data channels.

Features

- Satellite Multiple Access
- Digital Link Performance
- Advanced Technology
- Channel Configuration Flexibility
- Data and Audio Program Expansion

Satellite Multiple Access

The SCPC technique allows one satellite transponder to furnish economical simultaneous distribution of numerous broadcast services. Several business locations can share a satellite transponder from separate uplink facilities. When the transmitting sites are located free of RFI and within line of sight of the satellite, compatible installations within view of the satellite can receive the signals.

Selection among the pre-assigned SCPC signals within the satellite transponder is made by front panel controls on the equipment. This agility is provided in 100 kHz steps across the entire 36 MHz transponder bandwidth.

Digital Link Performance

For digital satellite communications links, Forward Error Correction (FEC) coding is employed to improve the accuracy rate of the received data. In the DAT-800, one additional bit is added to every seven information bits of the transmitted stream, to improve the bit error rate. High quality audio is achieved at a bit error rate of 10^{-7} where the link improvement is over 3 dB due to the FEC coding gain. This coding gain can be used to reduce the size of the receive antenna area by a factor of two, or to decrease the satellite EIRP requirements. Economic considerations for the total communication network will determine the most cost-effective utilization of this FEC coding gain.

Scientific
Atlanta
Satellite Communications Division

Advanced Technology

Sophisticated digital technology is incorporated within the SCPC Program Audio and Data Distribution System for reliable operation and service. Integrated circuit (IC) design including large-scale integrated (LSI) components are employed for long life. Reduced size and power consumption further increases system reliability. Forward Error Correction (FEC) improves signal quality within the system configuration, actually reducing errors present in the received data stream. Fifteen-bit encoding of the audio signal plus digital compression provide an audio dynamic range in excess of 80 dB. Digital companding techniques are used and reduce instantaneously a 15-bit word into an 11-bit word, thereby reducing the transponder bandwidth required. Nyquist filtering similarly reduces the bandwidth, resulting in space segment cost reduction.

Channel Configuration Flexibility

Modular construction of the uplink and downlink equipments within the SCPC System allows easy configuring for programming flexibility. Standard configurations include:

- 2 each 15 kHz audio channels and one each voice cue or 32 kb/s data channel
- 4 each 7.5 kHz audio channels and one each voice cue or 32 kb/s data channel
- One each 15 kHz audio channel and 13 each voice cue or 32 kb/s data channels
- Two each 7.5 kHz audio channels and 13 each voice cue or 32 kb/s data channels
- 24 each voice cue or 32 kb/s data channels

Special configurations can be accommodated to meet individual data distribution requirements within the 800 kb/s format.

Uses for the digital data system include:

- Point-to-point transmission of time sensitive audio
- Point-to-multipoint distribution of program audio for rebroadcast
- Point-to-multipoint desemination of digital data for business and industry

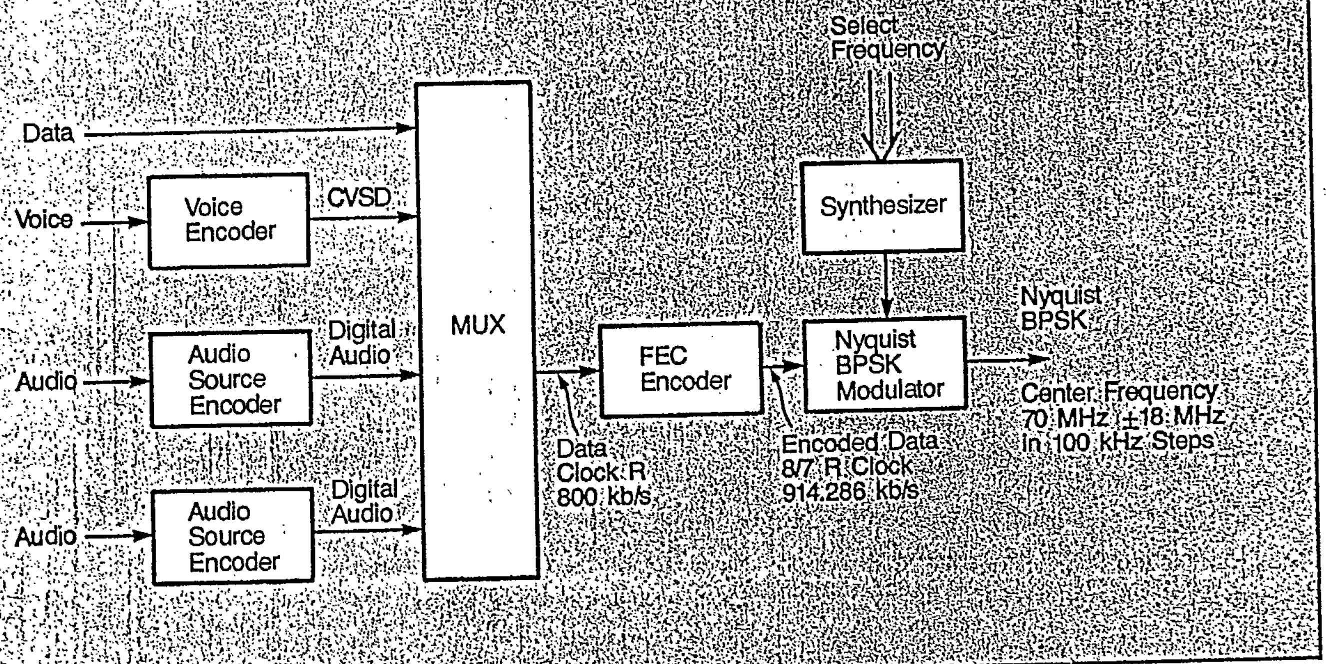
Configuration of the 800 kb/s data is flexible.

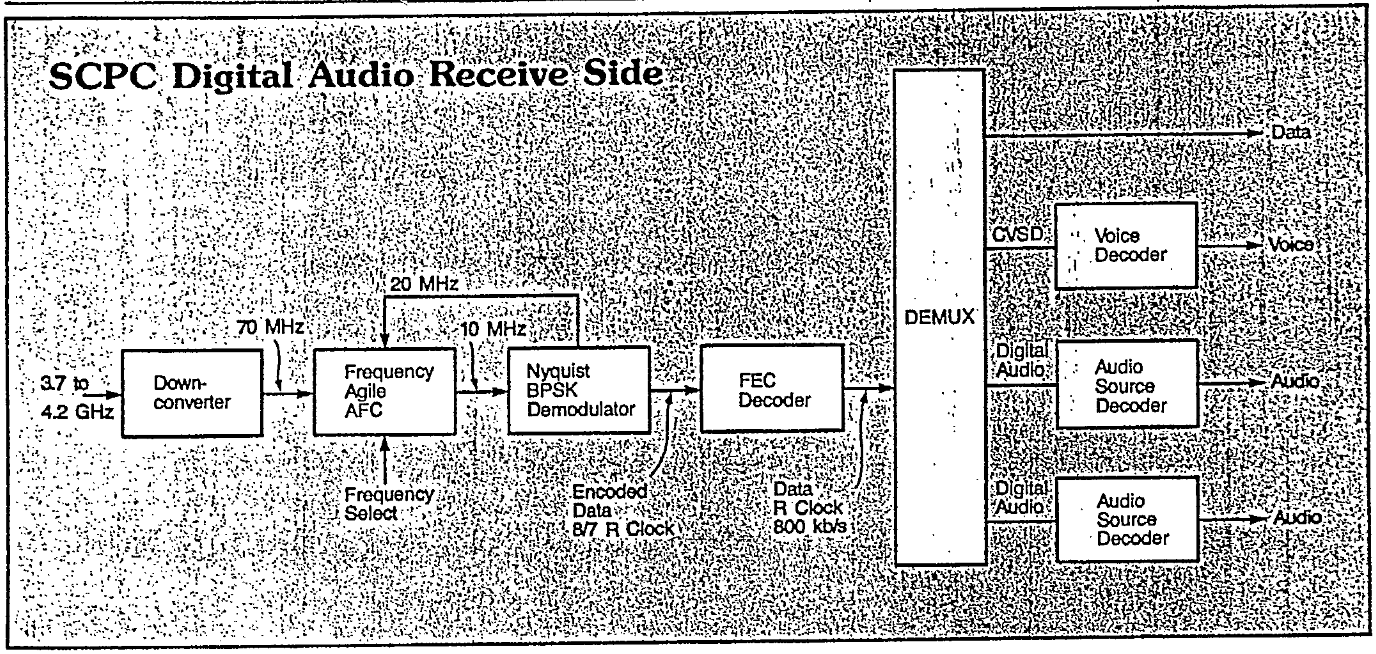
Data and Audio Program Expansion

Growth of distribution requirements within the data and audio program industries can be accommodated cost-effectively. Initial equipment purchases can be limited to the capacity needed for the near term. Later expansion needs can be met by purchase of 800 kb/s increments. For example, a broadcaster wishing to distribute a high quality stereo pair can limit his purchase to only that capacity.

As new programming is added for network distribution, a second SCPC system can augment the first system even if the originating studios are not co-located. Receiving equipment is then compatible with either program being uplinked. The same holds true for a growing business as its data distribution requirements increase.

SCPC Digital Audio Transmit Side





Non-Redundant Transmit and Monitor SCPC 800 kb/s System

The source encoding converts each 15 kHz audio channel to a 384 kb/s data stream, each 7.5 kHz audio channel to 192 kb/s and each voice cue or standard data channel to 32 kb/s. The voice cue channel uses a continuously variable slope delta (CVSD) modulation scheme to convert the 32 kb/s of data into an audio circuit similar to telephone quality. These are multiplexed together to form an overall 800 kb/s data rate. The data is processed by a rate 7/8 FEC encoder. The coded data is routed to the Nyquist BPSK modulator at a 914.286 kb/s rate. The output IF center frequency can be selected in 100 kHz increments over the 52 MHz to 88 MHz bandwidth in the frequency agile modulator. The ninety-nine percent transmit envelope is 1.28 MHz wide. The transmit IF spectrum is upconverted to a transponder frequency in the 5.925 GHz to 6.425 GHz range which is then amplified and transmitted using the appropriate size HPA and antenna combination. The upconverter, HPA, and transmit antenna are external to the 800 kb/s transmit electronics.

The 3.7 GHz to 4.2 GHz downlink signal is received by a suitably sized antenna. This RF carrier is amplified in the LNA such that an adequate signal level is input to the crystal controlled double conversion downconverter. The 70 ±18 MHz output from this unit is routed to the AFC module which downconverts to the final 10 MHz IF. Frequency selection in this unit is front panel controlled in 100 kHz increments. The carrier regeneration 20 MHz output from the demodulator is coupled back to the AFC module and is compared to a 20 MHz crystal oscillator (XO) in a phase detector. The resultant error signal is filtered and drives a synthesizer so that the second IF frequency is one half the 20 MHz XO frequency or 10 MHz. This AFC loop allows narrow band operation and overcomes uplink, satellite and downlink frequency offsets. The Nyquist BPSK demodulator detects the data from the 10 MHz IF carrier. The output 914.286 kb/s data stream inputs into the Forward-Error-Correction (FEC) module where threshold decoding with syndrome feedback is applied. This results in a 3.55 dB coding gain. The resultant 800 kb/s serial data stream is then input to the demultiplexer where decommutation into individual data streams for audio, voice cue and data channels takes place. The source decoding modules convert digital program or cue audio back to their respective analog forms.

Program Audio Channels

The 15 kHz program channel is sampled at a 32 kHz rate (16 kHz for a 7.5 kHz channel) and digitized into a 15-bit word. Digital companding techniques are used to instantaneously compress the 15-bit word into an 11-bit word to reduce the transmissions bandwidth. A parity bit is included resulting in a word length of 12 bits.

The parity bit is used as an additional error concealment encoding that allows the bit-error-rate (BER) to degrade to 10^{-5} before errors are "just perceptible" in the 15 kHz program audio output. Each program encoder and decoder contains two separate 15 kHz (or 7.5 kHz) channels which may be used independently or as a stereo pair. The balanced output level is +24 dBm maximum into a 600 ohm load. This allows for 16 dBm of headroom when +8 dBm average program level (APL) is employed. Maximum deviation from a flat frequency response is 1.0 dB over the 40 Hz to 15 kHz (7.5 kHz) band. The unit provides in excess of 80 dB peak signal-to-idle channel noise ratio and a signal-to-quantization noise ratio of 52 dB.

Voice Cue Channel

The voice cue channel is used for network closed circuit broadcast and event coordination. This circuit provides telephone quality, noise free, one-way voice communications capability. A CVSD encoding scheme is employed for this service at a 32 kHz rate.

Data Channel

The standard data decoder will simultaneously process three asynchronous RS-232 data streams using a single 32 kb/s channel. This unit is fully addressable from the originating studio. The data channel can be used for electornic mail and other distribution of one way data. Special, customer tailored, data channels can be designed to address specific requirements which fit within the 800 kb/s information data rate format.

A 32 kb/s data throughput capability is provided within the DAT-800 when multiplex and demultiplex equipment is customer provided. Interface is through a standard 25 pin, D-female connector in RS-232C format.

For additional information contact:

Broadcast Audio Marketing
Scientific-Atlanta, Inc.
3845 Pleasantdale Road
Atlanta, Georgia 30340
(404) 449-2381

Specifications

800 kb/s SCPC Receive Terminal

BPSK Receiver

Downconversion
  Dual
Received Data Rate
  914.3 kb/s
Information Data Rate
  800 kb/s
FEC Decoding
  Rate 7/8 hard decision threshold decoding
Bit Error Rate
  Better than 10^{-7} when available E_b/N_0 is 10 dB or greater

Program Audio 15-kHz Channel (7.5 kHz Channel)

Sample Frequency
  32 kHz (16 kHz)
Digital Compression
  15 to 11
APL
  +8 dBm into 600 ohms.
Peak Unaffected Signal
  +24 dBm into 600 ohms
Signal-to-Quantizing Noise Ratio
  52 dB at +22 dBm
Idle Channel Noise
  81 dB below +24 dBm
Total Harmonic Distortion
  0.3% maximum at +24 dBm

Voice Cue Channel

APL
  +8 dBm into 600 ohms
Signal-to-Quantizing Noise Ratio
  25 dB at +8 dBm
Output Impedance
  600 ohms ±10%

System Power Required

  105 to 130V ac, 57 to 63 Hz, 80W

Scientific-Atlanta, Inc.

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Canada: 1640 Bonhill Road, Unit 6, Mississauga, Ontario L5T 1C8, Canada; Telephone 416-677-6555-Telex 06983600
Europe: Horton Manor, Stanwell Road, Horton, Slough SL3 9PA, England; Telephone Colnbrook (02812) 3211; Telex 849406

DATE DUE
