STUDY OF DIGITAL MODULATION AND MULTIPLEXING TECHNIQUES APPROPRIATE TO THE DISTRIBUTION OF RADIO PROGRAMS BY SATELLITE O

(FINAL REPORT)

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Date: 3 January, 1983
MCS File No.: 8270
DSS File No.: 21ST-36001-3496
 DSS Contract No.: OST81-00256


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$\begin{array}{ll}P & D D 4576462 \\ 91 & D L-46504 \\ C 654 & \\ N 49 & \\ 1983 & \end{array}$
Government Gouvernement ol Canada
Department of Communications
DOC CONTRACTOR REPORTDOC-CR-SP
DEPARTMENT OF COMMUNICATIONS - OTTAWA - CANADA
SPACE PROGRAM
TITLE: Study of Digital Modulation and Multiplexing TechniquesAppropriate to the Distribution of Radio Programs by Satellite
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DEPARTMENT OF SUPPLY AND SERVICES CONTRACT NO: OST81-00256
DOC SCIENTIFIC AUTHORITY:
Eric Tsang .
CLASSIFICATION: Unclassified
This report presents the views of the author(s). Publicationof this report does not constitute DOC approval of the reportsfindings or conclusions. This report is available outside thedepartment by special arrangement.
DATE: 3 January, ..... 1983

# STUDY OF DIGITAL MODULATION AND MULTIPLEXING TECHNIQUES <br> APPROPRIATE TO THE DISTRIBUTION OF RADIO PROGRAMS BY SATELLITE 

## Executive Summary

Prepared by: Miller Communications Systems (MCS) Ltd. DSS Contract No.: OST81-00256

Date: January 3, 1983

In a previous contract, MCS performed a general study of candidate analog and digital modulation and multiplexing schemes suitable for the distribution of mono and stereo radio programs in a direct broadcast satellite (DBS) system. This system would provide both TV and radio programs to cable, community and home receivers. This study evaluates in detail a number of the candidate digital modulation and multiplexing techniques as they impact the following:

- satisfying specified radio performance objectives
- space segment (power and bandwidth) utilization
- selection of subsystem parameters (e.g. oscillator phase noise and frequency stability, receive system ( $G / T$ )
- susceptibility to interference
- operational flexibility
- growth capability
- compatibility with TV direct broadcast systems, TV and radio receive equipments
- compatibility with redistribution media

Two satellites are considered in this study: ANIK $C$ as a potential interim system, and a DBS whose parameters have been specified. For each satellite, two transponder useage categories are considered; specifically, these are dedicated and shared transponders. In the dedicated transponder case a transponder is used by radio signals only, whereas in the shared transponder case TV and radio signals both occupy the same transponder.

Three radio program performance levels are addressed: CATV, High Quality and Low Quality. These levels apply to cable TV, community and home reception applications respectively.

The digital techniques considered include radio program digitization, bit error control and digital modulation.

Pulse code modulation and delta modulation are compared for the digitization of program material for each of the three quality levels. It is found that 12,11 , and 9 bit PCM codecs with $\mu=255$ law companding satisfy the test tone to noise (TT/N) ratio requirements for CATV (50 dB), High (47 $\mathrm{dB})$ and Low Quality ( 36 dB ) applications respectively. These $T T / N$ ratios are satisfied with a bit error rate (BER) of $10^{-7}$ for CATV and High Quality systems and $B E R=10^{-6}$ for a Low Quality system. Delta modulation does not satisfy these $T T / N$ requirements throughout the radio program bandwidth.

Bit error control can be used to relax the BER requirements placed on the digital modem (to reduce link power) without compromising specified PCM performance objectives. Bit error control can be effected by means of forward acting error correcting (FEC) coding and error concealment. It is found that FEC coding need only be applied to the most significant bits (MSB's) in each PCM word. With a bit error rate $(B E R) \simeq 10^{-3}$ at the digital demodulator output and FEC coding providing a reduction in BER to $10^{-6}$ or $10^{-7}$, only $1 / 3$ of the bits in the $P C M$ word need be protected. A r-3/4 convolutional FEC codec using threshold decoding is identified as having potential.

While not investigated in detail, error concealment (EC) is identified as a technique for improving the subjective performance grade associated with PCM codecs subject to bit errors. In EC a parity bit is used to determine whether an error has occurred in each PCM word. Error detection results in a substitution of the erroneous sample by either the previous sample or a value based on interpolation between adjacent samples. For a $B E R=10^{-7}$, $E C$ can reduce the effects of channel errors to an imperceptible subjective grade.

Digital radio program data streams may be time division multiplexed. Issues relating to multiplexing alternatives (continuous versus packet), overhead bits and synchronization are reviewed.

The digital modulation techniques evaluated include binary and quaternary phase shift keying (BPSK and QPSK respectively) and minimum shift keying (MSK). Both coherent and differential detection are considered for each modulation technique.

Coherent binary phase shift keying (CBPSK), differential binary phase shift keying (DBPSK) and differential minimum shift keying (DMSK) are three digital modulation techniques found to be suitable for radio program distribution. These modulation techniques are suggested as their phase noise requirements are the least stringent and their power requirements are moderate (DBPSK and DMSK require about . 6 dB and l.l dB more power than CBPSK, respectively). In addition, their demodulators are simpler to implement. Their only drawback is an increase in bandwidth relative to quadrature phase shift keying (QPSK). In bandwidth limited cases coherent QPSK can be used to increase the bandwidth utilization efficiency. However, note that phase noise requirements for a coherent QPSK demodulator are more stringent.

The satellite scenarios for program distribution include dedicated transponder and shared transponder cases. In the dedicated transponder case one or more radio program carriers can be used. In the shared transponder scenario two options exist. An FMTV signal can share the transponder with radio program carriers (FMTV/SCPC) or, one or more radio programs can be time division multiplexed to modulate a radio subcarrier placed above the IV baseband with the composite baseband signal frequency modulating one carrier (subcarrier case).

System capacities have been determined for all scenarios. In the case of the dedicated transponder with multiple carriers, capacity is power limited. In the shared transponder FMTV/SCPC case capacity is power-limited and, for the DBS satellite, the capacity is also limited by the available RF bandwidth. In the subcarrier case the baseband bandwidth available is constrained by the TVassociated audio subcarrier and the colour subcarrier second harmonic.

In all cases, tables are provided showing the relationship between earth station $G / T$ and radio program capacity. It is shown that in limited capacity requirement situations (1-5 channels) the subcarrier technique requires a smaller earth station $G / T$ than the FMTV/SCPC approach.

Interference mechanisms and their impact on system capacity are reviewed for the various transponder scenarios.

In the case of the dedicated transponder which is power limited, adjacent channel interference (ACI) between SCPC signals can be controlled with suitably sized guard bands. Capacity calculations provided allow for intermodulaton products.

In the $F M T V / S C P C$ case it is concluded that SCPC signals should be placed at least 9 MHz away from the FMTV centre frequency to avoid ACI to the FMTV signal. However, FMTV ACI to the SCPC signals is more important. It is found that the SCPC signals should be placed approximately 13 MHz away (depending on relative carrier levels) from the FMTV signal. Furthermore, SCPC signals should be grouped to one side of the $F M T V$ signal with a guard band at least equal to the composite SCPC signal bandwidth in order to mitigate intermodulation degradation.

In the case of the subcarrier technique, mutual interference between TV-associated audio and radio program subcarriers depends on relative power levels and the filtering used. The only intermodulation products of concern are those falling into the TV baseband. The presence of high IM products in the TV baseband could limit the power associated with the radio subcarrier and thereby limit the quality level and/or the number of radio programs transmitted.

A review of interference in a frequency sharing environment (intra and inter-satellite interference) is also provided. Frequency plans which reduce interference from crosspolarized staggered transponders are presented.

The satellite scenarios have been reviewed with respect to system flexibility, and growth potential. The advantage of the SCPC approach (dedicated or shared transponder) is that radio and $T V$ programs can be transmitted individually. Furthermore, these programs do not have to originate at the same site.

The advantage of the subcarrier technique is that it is a low cost approach that is compatible with FMTV earth terminal equipment. However, capacity is limited in this case. Time division multiplexing of radio programs is suited to the larger CATV capacity requirements.

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In a previous contract*, Miller Communications (MCS) Ltd. performed a general study of candidate analog and digital modulation and multiplexing schemes suitable for the distribution of mono and stereo radio programs in a direct broadcast satellite (DBS) system. The direct broadcast satellite would provide both $T V$ and radio programs to cable, community and home receivers. The purpose of this study is to evaluate in detail a number of the candidate digital techniques as they impact the following:

- satisfying specified radio program performance objectives
- space segment utilization
- selection of subsystem parameters (e.g. oscillator phase noise and frequency stability, receive system $G / T$ )
- operational flexibility
- growth capability
- compatibility with TV direct broadcast systems, TV and radio receive equipment
- compatibility with redistribution media

The study is to be conducted for two satellite transponder scenarios and three radio program performance levels which

[^0]are described in Section l.2. For each of the scenarios and quality levels, suitable radio program digital encoding, error control and modulation techniques are to be recommended based on the results of the technical analysis.

### 1.2 System Descriptions

Table 1.1 sumarizes the system parameters which are to be investigated in the report.

In the dedicated transponder scenario, the full transponder is occupied by radio program carriers only. These carriers can each carry one radio program (single channel per carrier-SCPC) or a number of time division multiplexed (TDM'd) radio programs (probably up to 5 TDM'd radio programs per carrier).

In the shared transponder scenario, the TV and radio programs share the same transponder. There are two ways of sharing. One way is for the radio programs to be transmitted independently from the FM modulated TV signal (FMTV) as SCPC carriers. This is called the shared transponder - SCPC case. In the second way, the radio programs are TDM'd to modulate a subcarrier located above the highest frequency of the TV baseband signal. This is called the shared transponder - subcarrier case.

Other system parameters listed in Table l.l are transponder bandwidth, modulation and source encoding techniques, and radio program quality level. Three radio program quality levels have been specified (see Section l.3) according to signal-to-noise ratio objectives. These quality levels are associated with specific distribution applications as follows

- CATV quality level: for redistribution media (e.g. cable companies, broadcasting stations, etc...).


Table 1.1: List of system parameters which are to be investigated for radio program distribution.
*The analysis for the 23 MHz - DBS case is given in Appendix E.

- High quality level: for community reception or for high quality home reception.
- Low quality level: for home reception or for transmission of multiple sound signals associated with a single video signal.


### 1.3 Terms of Reference

The following are guidelines that have been specified for this study:
(i) Radio baseband signal-to-noise ratio (unweighted)

- CATV quality level: $T T / N \geqslant 50 \mathrm{~dB}$.
- High quality level: $T T / N \geqslant 47 \mathrm{~dB}$.
- Low quality level : TT/N $\geqslant 36 \mathrm{~dB}$.
(ii) Number of Radio Programs
- Per TV : from 2 to 5 radio channels.
- Per beam: up to 40 mono radio channels.

This guideline is based on the assumption of 4 to 8 TV channels per satellite beam. Although a possibility, it is not to be implied that any radio program is to be associated with a specific TV channel. For example, in the dedicated transponder case, all radio channels for a given beam may be in one transponder dedicated for radio traffic.
(iii) Nominal TV signal parameters

- One FMTV signal per transponder
- TV carrier-to-noise density ratio $=87.0 \mathrm{~dB}-\mathrm{Hz}$
- TV receive IF filter: 4 th order, 0.5 dB ripple Chebyshev filter with equivalent noise bandwidth of 18 MHz
- Peak frequency deviation $=6 \mathrm{MHz}$
- Top video baseband frequency $=4.2 \mathrm{MHz}$ (NTSC).

With the values given above, the nominal carrier-tonoise ratio is 14.4 dB and the unweighted TV signal-to-noise ratio is 31.6 dB .
(iv) Radio program bandwidth $=15 \mathrm{KHz}$. While low quality 5 kHz channels could be incorporated, the baseline evaluation is to be performed for 15 kHz channels only.
(v) Earth station G/T's

- DBS: 6 to $10 \mathrm{~dB} / \mathrm{K}$
- ANIK C: 16 to $22 \mathrm{~dB} / \mathrm{K}$


### 2.1 Source Encoding for Radio Program Distribution

This section assesses two methods for digital encoding of radio programs, namely,
(1) Pulse Code Modulation (PCM) with or without companding

Delta Modulation (DM) with or without companding.

The two encoding techniques are presented and compared on the basis of:
(1) The achievable signal to noise ratio as a function of transmitted bit rate and channel bit error.rate (BER).
(2) Codec Complexity.

The objective is to recommend codecs which can satisfy three program quality levels:

```
Low Quality: TT/N > 42 dB weighted
High Quality: TT/N \geqslant 53 dB weighted
High Quality: (CATV) TT/N > 56 dB weighted.
```

The term "weighted test tone to noise ratio" implies that the instrument used to measure the codec output signal power and noise power has a frequency response which is not constant over the frequency range of interest (DC to 15 kHz in our case). The measurement instrument frequency response or weighting used compensates for the different sensitivity of the human ear to different frequencies; noise components at higher frequencies
( $>1 \mathrm{kHz}$ ) tend to be more noticeable to the human ear than lower frequency components. According to CCITT recommendations in Vol III-2 Rec. J. 21 [1], weighting reduces the audible noise power level by 6 to 10 dB , or equivalently, the codec's output $\operatorname{SNR}$ is improved by that amount. Therefore, the non-weighted $T T / N$ ratios, corresponding to the three quality levels mentioned are:

> Low Quality: $T T / N \geqslant .36 \mathrm{~dB}$ non-weighted High Quality: $T T / N \geqslant 47 \mathrm{~dB}$ non-weighted High Quality: (CATV) $\mathrm{TT} / \mathrm{N} \geqslant 50 \mathrm{~dB}$ non-weighted

In the analysis presented below, possible signal clipping due to finite quantizer input signal range handing capability is not accounted for.
2.1.1 Pulse Code Modulation

### 2.1.1.1 Binary PCM Encoding

In binary PCM the bandimited analog waveform is sampled at least at the Nyquist rate (which is twice the baseband bandwidth) to avoid aliasing noise and converted to a puise amplitude modulated (PAM) waveform. Each pulse is then encoded by an $n$ bit binary code word. The assignment of binary code words to the pulses is performed by a Quantizer. The Quantizer divides the input signal dynamic range ( $\pm v$ volts) into $M$ sub ranges $\left(M=2^{n}\right)$. Each subrange is assigned one of the $2^{n}$ binary code words. If the subranges are all equal, the quantizer is referred to as uniform, and the PCM encoder is of a uniform type. If the subranges are not equal the quantizer is referred to as tapered and the encoder is of a companded type. The binary codes chosen for PCM usually have the property that adjacent quantizer levels are assigned code words which
have minimal number of code bit disagreements (usually only one disagreement is acceptable). The reason behind this strategy is to make the code less susceptible to errors in the $A / D$ conversion. For a given sampling rate, the choice of the code word length, $n$, determines the transmitted bit rate, the channel bandwidth and the signal-to-quantization noise ratio ( $\mathrm{SN}_{\mathrm{q}} \mathrm{R}$ ) of the encoding process, achievable over a specified input signal dynamic range.

It is obvious that with uniform quantization, the $S N{ }_{q} R$ is worse for low level signals than for high level signals. In the case of voice or music signals the probability of long periods of low level signal is very high and, therefore, the tendency is to taper the quantization level structure so as to have finer resolution at lower levels and more coarse resolution at higher levels. Tapering the quantizer levels results in a companding (compressingexpanding) PCM codec. The merits of companding are discussed quantitatively in the following section.

## 2.l.l.2 Quantization SNR And Dynamic Range Improvement With $\mu$-Law Companding

In this section the $\mathrm{SN}_{\mathrm{q}} \mathrm{R}$ improvement with a $\dot{\mu}$-law compander is demonstrated.

A typical companding characteristic has a logarithmic form. A common form implemented in practice for speech (in telephony) is $\mu-l a w$ companding which relates the input and output waveforms by:

$$
\begin{equation*}
y(x)=\frac{\ln (1+\mu x / v)}{\ln (1+\mu)} \quad 0 \leqslant x \leqslant v \tag{2.1}
\end{equation*}
$$

The parameter $\mu$ can be varied to obtain a variety of transfer characteristics as shown in Figure 2.1. It is noted that for $x \ll \frac{V}{\mu}$ the characteristic is almost linear:


Figure 2.1: Logarithmic compression characteristics.

$$
\begin{equation*}
y(x)=\frac{\mu x}{V \ln (I+\mu)} \quad x \leqslant \leqslant \frac{V}{\mu} \tag{2.2}
\end{equation*}
$$

For $\mu \leqslant \leqslant 1$ the nonlinear logarithmic characteristic reduces to:

$$
\begin{equation*}
y(x)=\frac{x}{V} \tag{2.3}
\end{equation*}
$$

and the quantizer is uniform.

The Bell System in the U.S. has adapted $\mu=255$ for its digital carrier system $[2,3]$ 。

The $\mu$-Law Companded PCM Signal to Quantization Noise Ratio $\left(S N_{q} R\right)$ is given by [2]:

$$
\begin{equation*}
\operatorname{SN}_{\mathrm{g}^{R}}=\frac{3 M^{2}}{[\ln (1+\mu)]^{2}} \frac{1}{1+2 E[|x|] / \mu \sigma_{X}^{2}+1 / \mu^{2} \sigma_{X}^{2}} \tag{2.4}
\end{equation*}
$$

where:
$M$ is the number of transmitted levels,
$\mu$ is the compander's order,
$x$ is the normalized input signal $=x^{\prime} / V_{p e a k} \leqslant \pm 1$,
$\sigma_{\mathrm{x}}$ is the rms value of the normalized input signal.
$E[|x|]$ depends on $f(x)$, the probability density function (pdf) assumed for the signal, and the choice of $f(x)$ depends on the nature of the signal. It should be pointed out, however, that the ratio $\mathrm{E}[|\mathrm{x}|] / \sigma_{\mathrm{x}}$ does not vary much from one paf to another.

The Gaussian pdf given by:

$$
\begin{equation*}
f(x)=\frac{1}{2 \sqrt{\pi}} \exp \left[-\frac{x^{2}}{\sigma_{x}{ }^{2}}\right] \tag{2.5}
\end{equation*}
$$

is useful in modelling signals, such as noise processes, that have relatively low probability of exceeding $\sigma_{x}$. The Laplacian pdf given by:

$$
\begin{equation*}
f(x)=\frac{1}{\sqrt{2} \sigma_{x}} \exp \left[-\sqrt{2}|x| / \sigma_{x}\right] \tag{2.6}
\end{equation*}
$$

is a good model of signals, such as speech, that have a relatively high probability of attaining values larger than $\sigma_{x}$. The probability that $|x|$ exceeds $4 \sigma_{x}$ is .0035 for a Laplacian distributed signal and only . 001 for a Gaussian distributed signal. For the Laplacian and the Gaussian paf's $E[|x|] / \sigma_{x}$ is 1.4 and 1.6 , respectively.

Eq. (2.4) can be used to plot the $S N_{q} R$ vs $\sigma_{x}{ }^{2}$ where $\sigma_{x}^{2}$ is expressed in $d B$ relative to the maximum input signal $\left(\sigma_{\mathrm{x}}^{2}<1\right)$. From the curve obtained, it is possible to estimate:
(1) The input dynamic range for relatively constant $\mathrm{SN}_{\mathrm{q}} \mathrm{R}$ at the encoder output, and
(2) The performance improvement with companding relative to uniform encoding,
for given $M$ quantizer levels ( $\log _{2} M$ bit code) and $\mu$-law companding characteristics.

For a 12 bit encoder the $S N_{q} R$ vs $\sigma_{x}^{2}$ curve is plotted in Figure 2.2 for $\mu=255$, and $\mu=0$ (no companding). It is observed that:
(1) with companding the input dynamic range for relatively constant $\mathrm{SN}_{\mathrm{q}} \mathrm{R}_{\mathrm{r}}$ is 40 dB or more and
(2) the improvement in $\mathrm{SN}_{\mathrm{q}} \mathrm{R}$ gained, through companding is more than 30 dB for $\sigma_{x}^{2} \leqslant-53 \mathrm{~dB}$.

The $\mu=255$ companding law is used throughout the following analyses, because, as will be shown:
(l) It provides a wide input signal dynamic range (~40 $d B$ ) over which the output $S / N$ or $T T / N$ ratios are sufficiently high
(2) It is easily implemented by digital means.

The $\mu$-law companding effects can be achieved by proper digital compression of binary code words generated by a uniform quantizer. The companding effects achievable by the digital compression are identical to those obtainable through analog companding by a multi-segment, piecewise-linear $\mu-l a w$ companding characteristics.

As an example, a 9 -digit 8 -segment law can be implemented by starting with a 13 binary digit uniform encoder. The first digit is the sign digit; the number of leading zeros in the remaining digits when subtracted from 7 determines, in binary notation, the segment number on which the sample belongs. The first 1 following the leading 0's is skipped except for segment 0 ; the next 5 of the remaining digits are copied to determine one of the 32 intervals in each


Figure 2.2: Signal to Quantization Noise with Companding vs. Input Signal Power
segment. Table 2.1 illustrates the translation. From the 9 digit compressed code, a 13 digit linearized code can be obtained by filling the unknown lesser significant digits arbitrarily.

Figure 2.3 illustrates the conversion process for the 13 to 9 bit digital compression. It is observed that the digital compression process is equivalent to an 8-segment piecewise-linear $\mu=255-1$ aw analog compression. Each segment is twice as long as the preceeding one and all have the same number of intervals. It is obvious that this method can be used to compress any linear code of more than 7 bits to an 8 segment $\mu=255$ - companded code of more than 4 bits.

Digitally linearizable compression also lends itself to digital signal processing such as filtering to shape bandwidth and multiplication to effect gain.
2.1.1.3 Performance of PCM Decoders With Digital Errors

In this section the relationship among the PCM codec system parameters is established mathematically as a basis for specifying system parameters which would result in a desired system performance: Performance is given in terms of output signal to total noise power ratio (quantization plus error related noise). The analysis presented is based on the following assumptions:
(i) The program signal is bandlimited to B Hz .
(ii) The program signal has zero mean and r.m.s. value of $\sigma_{S}$. The signal power is $\sigma_{S}^{2}$.
(iii) The signal is sampled at a rate of 2 B samples per second.


I


$\binom{6=}{110}$

0.625
$\binom{4=}{100}$



$0.375-$
$\binom{2=}{010}$

 (10)



 near word ( 13 Bts )

-
 - ${ }^{-3}$ $\qquad$




## 13 Bit Linear Code

9 Bit Companded Code

| 1 | 2 | 3 | 4 | 5 | 6 | 7 | 8 | 9 | 10 | 11 | 12 | 13 | 1 | 2 | 3 | 4 | 5 | 6 | 7 | 8 | 9 |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
| S | 0 | 0 | 0 | 0 | 0 | 0 | 0 | a | b | c | d | e | S | 0 | 0 | 0 | a | b | c | d | e |
| S | 0 | 0 | 0 | 0 | 0 | 0 | 1 | a | b | c | d | e | S | 0 | 0 | 1 | a | b | c | d | e |
| S | 0 | 0 | 0 | 0 | 0 | 1 | a | b | c | d | e | f | S | 0 | 1 | 0 | a | b | c | d | e |
| - |  |  |  |  |  |  |  |  |  |  |  |  | - |  |  |  |  |  |  |  |  |
| - |  |  |  |  |  |  |  |  |  |  |  |  | - |  |  |  |  |  |  |  |  |
| - |  |  |  |  |  |  |  |  |  |  |  |  | - |  |  |  |  |  |  |  |  |
| - |  |  |  |  |  |  |  |  |  |  |  |  | - |  |  |  |  |  |  |  |  |
| S | 1 | a | b | c | d | e | f | g | h. | x | x | x | S | 1 | 1 | 1 | a | b | c | d | e |

Notes: $S$ is a sign bit
$S=1$ for positive signal
$S=0$ for negative signal

Digits $a b c a l a r e ~ c o p i e d ~ f r o m ~ l i n e ̀ a r ~ c o d e ~ t o ~ c o m p r e s s e d ~$ code

Digits xx... are ignored.

Table 2.l: 8 Segment Piecewise Linear $\mu=255$ Compander
(iv) Each sample is quantized to one of $2^{\text {n }}$ levels.
(v) Each sample is coded as a binary code word of $n$ bits.
(vi) Each detected code word is not likely to contain more than a single error induced by channel noise
(vii) All code bits are equally susceptible to error.

The receiver output noise has two components:
(i) Noise due to quantization error
(ii) Noise due to detection errors induced by channel additive noise.

Therefore, the power of the total output noise can be defined as

$$
N=\overline{\varepsilon_{q}^{2}}+\operatorname{Pr}\left[\begin{array}{l}
\text { detection error } \\
\text { in the code word }
\end{array}\right] \overline{\varepsilon_{n}^{2}}
$$

where $\overline{\varepsilon_{q}^{2}}$ and $\overline{\varepsilon_{n}^{2}}$ are the mean square noise contributions due to quantization and channel noise, respectively.

In calculating the noise components, the effect of the compander characteristics and the binary code used should be included.

The PCM binary code chosen affects the encoder quantization noise and the codec performance over a noisy channel. Three typical codes are the natural binary code, the Gray code, and the folded binary code shown in Table 2.2.


Table 2.2: Three 3-bit Codes For PCM

It is shown in Appendix A that the folded binary code is affected by digital errors less then the other two codes.

For a folded binary code ( $n-1$ bits and one sign bit) and a $\mu$-law logarithmic compander, the quantization noise can be expressed as [2]:

$$
\begin{equation*}
\overline{\varepsilon_{\mathrm{q}}^{2}}=\left[\frac{\ln (1+\mu)}{\mu}\right] \frac{1}{12 \cdot 2^{2 \mathrm{n}-2}}\left(1+\mu^{2} \sigma_{\mathrm{sn}}^{2}+2 \mu \mathrm{E}\lfloor|\mathrm{Sn}|\rfloor\right. \tag{2.7}
\end{equation*}
$$

where

$$
\begin{aligned}
& \mu \text { is the compander parameter } \\
& \sigma_{\mathrm{sn}}^{2} \triangleq \sigma_{\mathrm{s}}^{2} / \mathrm{V}_{\mathrm{max}}^{2} \text { is the normalized input signal power }
\end{aligned}
$$

and

$$
\begin{aligned}
& E\left[\left|S_{n}\right|\right]=\int_{-1}^{1} S_{n}^{\prime} f\left(S_{n}^{\prime}\right) d S_{n}^{\prime} \text { where } \\
& E\left(S_{n}\right) \text { is the input signal amplitude pdf }
\end{aligned}
$$

and

$$
\mathrm{s}_{\mathrm{n}} \triangleq \mathrm{~s} / \mathrm{V}_{\max }
$$

With no companding $(\mu \rightarrow 0)$ one obtains $[2,4]$ :

$$
\overline{\varepsilon_{\mathrm{q}}^{2}}=\frac{1}{3.2^{2 \mathrm{n}}}
$$

In calculating $\overline{\varepsilon_{n}^{2}}$ the following arguments are made:

1. An error in the eth least significant bit (LSB) of the code gives rise to an error $\mathrm{x}_{i}$ in the output signal given by
$x_{i}=\frac{1}{\mu}(1+\mu)^{2^{i-1} q_{-1}} \quad i=1 \ldots, n-1$
where $q=\frac{1}{2^{n-1}}=$ the companded signal uniform quantization step size.
2. The noise contributed by all possible single errors in each of the $(n-1)$ LSB of the code is:
$\sum_{i=1}^{n-1} x_{i}^{2}=\sum_{i=1}^{n-1} \frac{1}{\mu^{2}}\left[(1+\mu)^{2^{i-1}} q_{-1}\right]^{2}$
3. The noise contributed by a possible error in the sign bit is:
$\left(2 \sigma_{s n}\right)^{2}$
4. The total channel error noise is then:
$\overline{\varepsilon_{n}^{2}}=\frac{1}{n}\left\{4 \sigma_{\sin }^{2}+\sum_{i=1}^{n-1} \frac{1}{\mu^{2}}\left[(1+\mu)^{2^{i-1} q_{-1}}\right]^{2}\right\}$.
With no companding $(\mu \rightarrow 0)$ one gets:

$$
\begin{aligned}
\overline{\varepsilon_{n}^{2}} & =\frac{1}{n}\left\{4 \sigma_{s n}^{2}+\sum_{i=1}^{n-1} \frac{1}{\mu^{2}}\left[1+2^{i-1} q \mu-1\right]^{2}\right\} \\
& =\frac{1}{n}\left[4 \sigma_{s n}^{2}+\frac{1}{2^{2 n-2}} \sum_{i=1}^{n-1}\left(2^{2}\right)^{i-1}\right]= \\
& \cong \frac{1}{n}\left[4 \sigma_{s n}^{2}+\frac{1}{3}\right]
\end{aligned}
$$

For $\mu \geqslant>1$ it can be shown that:

$$
\overline{\varepsilon_{n}^{2}} \cong \frac{1}{n}\left\{4 \sigma_{s n}^{2}+\frac{1}{\mu}\right\}
$$

If the $P_{e} \leqslant \leqslant l$ it can be shown that:

$$
\mathrm{Pr}\left[\begin{array}{l}
\text { detection error } \\
\text { in the code word }
\end{array}\right]=\mathrm{nP}_{\mathrm{e}}
$$

where $P e$ is the probability of a bit error due to channel noise and $n$ is the number of bits per code word.

With the above results, the output SNR for a PCM system with a $\mu$-law compander and a folded binary code (FB) is given by:
$\left(\frac{S}{N}\right)_{P C M-F B}=\frac{\sigma_{s n}^{2}}{\left[\frac{\ln (1+\mu)}{\mu}\right]^{2} \frac{1}{3.2^{2 n}}\left(I+\mu^{2} \sigma_{S n}^{2}+2 \mu E\left[\left|s_{n}\right|\right]\right)+\left(4 \sigma_{S n}^{2}+\frac{1}{\mu}\right) P_{e}}$ (2.8)

With no companding $(\mu \rightarrow 0)$ :

$$
\begin{equation*}
\left(\frac{\mathrm{S}}{\mathrm{~N}}\right)_{\mathrm{PCM}-\mathrm{FB}} \cong \frac{\sigma_{\mathrm{sn}}^{2}}{\frac{1}{3} \cdot 2^{-2 n}+\left[4 \sigma_{\mathrm{sn}}^{2}+\frac{1}{3}\right] \mathrm{P}_{\mathrm{e}}} \tag{2.9}
\end{equation*}
$$

It can be shown that for a natural binary code $(\mu \rightarrow 0)$ [4]

$$
\begin{equation*}
\left(\frac{S}{N}\right)_{P C M-N B}=\frac{\sigma_{s n}^{2}}{\frac{1}{3} \cdot 2^{-2 n}+\frac{4}{3} P_{e}} \tag{2.10}
\end{equation*}
$$

Thus, it is obvious that $\mathrm{PCM}-\mathrm{FB}$ is superior to $\mathrm{PCM}-\mathrm{NB}$ as long as

$$
4 \sigma_{s n}^{2}+\frac{1}{3} \leqslant \frac{4}{3}
$$

or

$$
\sigma_{s n} \leqslant \frac{1}{2}
$$

i.e. as long as the quantizer r.m.s. loading is \&.5.

If it. is assumed that $\mu=255$ and that the signal has a Laplacian distribution, then, the probability that the amplitude is larger than $4 \sigma_{s}$ is $3.5 \times 10^{-3}$ ( $10^{-4}$ for a Gaussian distribution) and $2 \mathrm{E}\left[\left|s_{n}\right|\right]=1.414 \sigma_{s n}$.

It follows. that (by Eq. 2.8):

$$
\begin{equation*}
\left(\frac{\mathrm{S}}{\mathrm{~N}}\right)_{\mathrm{PCM}-\mathrm{FB}} \cong \frac{\sigma_{\mathrm{Sn}}^{2}}{2^{-2 n_{1}} 1.58 \times 10^{-4}\left(1+255^{2} \sigma_{\mathrm{sn}}^{2}+255 \times 1.414 \sigma_{\mathrm{sn}}\right)+\left(4 \sigma_{\mathrm{sn}}^{2}+\frac{1}{255}\right) \mathrm{P}} \tag{2.11}
\end{equation*}
$$

For $\sigma_{\mathrm{Sn}}^{2}=\frac{1}{16}$ the quantize loading is about $\frac{1}{4}$ and it follows that:

$$
\begin{equation*}
\left(\frac{\mathrm{S}}{\mathrm{~N}}\right)_{\mathrm{PCM}-\mathrm{FB}}=\frac{1}{10.5 \times 2^{-2 \mathrm{n}}+4 \mathrm{P} \mathrm{e}} ; \mu=255 \tag{2.12}
\end{equation*}
$$

The performance for the same loading with no companding is (by Eq. 2.9):

$$
\begin{equation*}
\left(\frac{\mathrm{S}}{\mathrm{~N}}\right)_{\mathrm{PCM}-\mathrm{FB}}=\frac{1}{5.332^{-2 \mathrm{n}}+9.3 \mathrm{P}_{\mathrm{e}}} ; \mu=0 \tag{2.13}
\end{equation*}
$$

For a test tone:

$$
E\left[\left|s_{n}\right|\right]=\overline{\left|s_{n}\right|}=\frac{1}{\pi} \int_{0}^{\pi} \sqrt{2} \sigma_{\operatorname{sn}} \sin \theta d \theta=2 \frac{\sqrt{2}}{\pi} \sigma_{\mathrm{sn}}
$$

where $0 \leqslant \sigma_{\mathrm{sn}} \leqslant \frac{1}{\sqrt{2}}$
or $\quad 0 \leqslant \sigma_{\mathrm{Sn}}^{*} \leqslant 1$ where $\sigma_{\mathrm{sn}}^{*}=\sqrt{2} \sigma_{\mathrm{sn}}$

Therefore, it follows that for $\mu \rightarrow 0$ (by Eq. 2.9):

$$
\begin{align*}
& \left(\frac{S}{N}\right)_{P C M-F B}=\left.\frac{\sigma_{S n}^{* 2}}{\frac{2}{3} 2^{-2 n}+\left(4 \sigma \underset{S n}{* 2}+\frac{2}{3}\right) P_{e}}\right|_{\sigma_{S n}^{* 2}=1}=\frac{1}{\frac{2}{3} 2^{-2 n}+\frac{14}{3} P_{e}} \text { (2.14) } \\
& \text { and for } \mu=255 \text { (by Eq. 2.8): } \\
& \left.\left(\frac{S}{N}\right)_{P C M-F B} \cong \frac{\sigma_{s n}^{* 2}}{2^{-2 n+1} \cdot 1.58 \times 10^{-4}\left(1+\frac{255^{2}}{2} \sigma_{S n}^{* 2}+255 \frac{4}{\pi} \sigma_{S n}^{*}\right)+\left(4 \sigma_{S n}^{* 2}+\frac{2}{255}\right) P_{e}}\right|_{\sigma_{S n}^{*}=1} \\
& \cong \frac{1}{10.37 \times 2^{-2 n}+4 P_{e}} \tag{2.15}
\end{align*}
$$

Figure 2.4, plotted using Eq. 2.12 and Eq. 2.13, shows the non-weighted $\left(\frac{S}{N}\right)_{P C M-F B}$ vs $n$ for different values of $P_{e}$, with and without companding, $25 \%$ quantizer loading and Laplacian signal amplitude distribution.

It is observed that with 14 to 16 bits and $\mathrm{P}_{\mathrm{e}}=10^{-7}$ the resulting $\left(\frac{S}{N}\right)_{P C M-F B}$ is 63.5 dB (non weighted) with $\mu=255$ companding. Without companding, $n \geqslant 13$ and $P_{e}=10^{-7}$ the non weighted performance is 60 dB . To achieve 60 dB with companding, 12 bits/sample are required at $\mathrm{P}_{\mathrm{e}}=10^{-7}$.

Figure 2.5, plotted using Eq. 2.11, shows the non weighted $\left(\frac{S}{N}\right)_{P C M-F B}$ vs the quantizer loading for different values of $P_{e}$; for $n=14$, with and without companding, and Laplacian signal amplitude distribution. It is observed that for


$P_{e}=10^{-7}$, companding significantly extends (by 20 dB ) the signal dynamic range over which the non weighted performance is better than 60 dB . The dynamic range in this case is $\geqslant 30 \mathrm{~dB}$.

Figure 2.6 is similar to Figure 2.5 except that $n=12$. From Figure 2.6 it is observed that in order to meet a high quality performance of $\geqslant 58 \mathrm{~dB}$ (non weighted) a codec having 12 bits per $\mu=255$-law companded code word should operate with a channel $B E R \leqslant 10^{-8}$. The resulting dynamic range is $\sim 40 \mathrm{~dB}$.

Figure 2.7, plotted using Eq. 2.14 and Eq. 2.15 shows that for a sine wave test tone and full quantizer loading the non-weighted $\left(\frac{\mathrm{T} T}{\mathrm{~N}}\right)_{\mathrm{PCM}}-\mathrm{FB}$ is $\geqslant 60 \mathrm{~dB}$ with $\mu=255$-law companding, $n \geqslant l 2$ and $P_{e}=l 0^{-7}$. It is observed that for $n=12$ and with a full sine wave quantizer loading the performance with companding ( $T T / N=62 ; P_{e}=10^{-8}$ ) is worse than without companding (TT/N $\left.=70 ; P_{e}=10^{-8}\right)$, but with companding and $P_{e}=10^{-8}$, the test tone dynamic range for $T T / N=60 \mathrm{~dB}$ is increased significantly to about 34 dB (less then 14 dB without companding), as shown by Figure 2.8 .

Figure 2.8, plotted using Eq. 2.14 and Eq. 2.15, shows the number of bits required to maintain a desired dynamic range for given $T T / N$ and $P_{e}$ with and without companding. It is observed that with no companding and $P_{e}=0.0,15$ bits are required to maintain $T T / N=60 \mathrm{~dB}$ over a 32 dB dynamic range. If companding is employed, 12 bits are required to maintain $T T / N=60 \mathrm{~dB}$ over a 34 dB dynamic range even if the $P_{e}$ degrades to $10^{-8}$. The dynamic range would be less than 14 dB for $T T / \mathrm{N}=60$ and $\mathrm{P}_{\mathrm{e}}=10^{-8}$ without companding.



By comparing Figure 2.4 and Figure 2.7 it is noted that for the same $P_{e}$, the calculated companded performances (for a 100\% loading test tone and for a $25 \%$ loading Laplacian pdf input signal) are almost identical.
2.1.2 Delta Modulation

The Differential PCM (DPCM) encoder attempts to predict the next sample value and quantizes only the difference between the predicted and actual value. In this way the variance of the quantizer output is greatly reduced below the variance value of the system input. The simplest form of predictive quantizing is delta modulation (DM), in which only a one-bit quantizer is used, and the sampling rate is raised above the Nyquist rate. More complicated DPCM techniques employ $\ell-$ bit quantizers with or without adaptive quantization step adjustments.

Two types of quantization noises are involved:

1. slope overload noise and
2. granular quantization noise.
(PCM suffers granular quantization noise and quantizer amplitude saturation or clipping noise).

Assuming slope overload is avoided for optimum performance, the maximum test-tone to total noise ratio for a DM codec is given by $[5,6]:$

$$
\begin{equation*}
\left(\frac{T T}{N}\right)_{D M}=\frac{3 \pi /\left(\omega_{T} \tau\right)^{3}}{1+\left\{24 \mathrm{P}_{e^{2}} / \omega_{T}{ }^{\omega} o^{\tau}\right\}} \tag{2.16}
\end{equation*}
$$

where

$$
\begin{aligned}
\tau= & \frac{l}{f_{S}} \text { the sampling interval } \\
\mathrm{P}_{\mathrm{e}}= & \text { bit probability of error due to channel noise, } \\
\omega_{T}= & 2 \pi f_{T} \text { test tone frequency or upper limit of the } \\
& \text { baseband frequency range. } \\
\omega_{0}= & 2 \pi f_{O} \text { decoder reconstruction filter lower } \\
& \text { frequency cutoff (assumed } \left.f_{0}=50 \mathrm{~Hz}\right) \text { ). }
\end{aligned}
$$

Eq. 2.16 was used to plot Figure 2.9 which shows the $T T / N$ ratio vs . $\mathrm{f}_{\mathrm{s}}$ for a l kHz test tone for $\operatorname{BER}$ of $10^{-6}, 10^{-7}$ and $10^{-8}$. Also shown are the $T T / \mathrm{N}$ ratios for 5 kHz and 15 kHz test tones for $B E R=0.0$.

It is observed that for a 1 kHz test tone and for $\mathrm{f}_{\mathrm{s}} \geqslant 200 \mathrm{kHz}$ the $T T / N$ ratio meets the CATV high quality requirement ( $T T / \mathrm{N}$ $\geqslant 50 \mathrm{~dB}$ non weighted) for $\mathrm{P}_{\mathrm{e}} \leqslant 10^{-6}$. However, for a tone of 15 kHz and $\mathrm{f}_{\mathrm{S}}=200 \mathrm{kHz}$ it is observed that $\mathrm{TT} / \mathrm{N}=20 \mathrm{~dB}$ even for $P_{e}=0.0$.

In Figure 2.9 the number of PCM bits per sample ( $N$ ) resulting the same transmission rate as delta modulation (assuming 32 kHz PCM sampling rate) is indicated.

Comparing Figure 2.9 to Figure 2.7, indicates that for a 1 kHz test tone $\mathrm{n}=12$ ( 384 kbps transmission rate) and $P_{e}=10^{-7}$ the performances of the two techniques are similar but for the 5 kHz test tone, PCM is superior, since in PCM the performance does not depend on the test tone frequency.


Figure 2.10 [4] shows the output signal to noise ratio vs. the normalized bit rate $R_{b} \triangleq \frac{\ell f_{S}}{2 f_{m}}$ where $f_{S}$ is the sampling rate, $f_{m}$ is the tone frequency or baseband bandwidth and $\ell$ is the number of bits ( $\ell=n$ for $P C M$ and $\ell=1$ for $D M$ ). It is observed that PCM is superior even for $n=8$ and moderate sampling rates ( $\leqslant 200 \mathrm{kHz}$ ).

The merit of delta modulation is that no overhead bits are required for framing of code words as in PCM, since in a DM receiver the baseband is reconstructed directly from the input bit stream. In PCM, some overhead bits are required for each channel for code word framing. The PCM framed bit stream structure enables incorporation of simple error control schemes in the basic design which may enhance its performance even further. These will be reviewed in subsequent sections.

It is shown in [7] that the dynamic range of DM codecs can be extended up to 40 dB by employing adaptive step size in the feedback predictor. Such DM codecs are referred to as Continuously Variable Slope Delta Modulator (CVSDM). It is noted that CVSDM is a technique aiming at mitigating overload effects in DM; the previous assessment of DM performance (in comparison to PCM with no signal clipping) assummed no overload at all, as indicated.

In terms of complexity, PCM codecs and CVSDM codecs are similar. In PCM codecs there is a requirement for code word framing overhead bits to be transmitted and detected in the receiver to establish timing synchronization. In CVSDM codecs there is a requirement for a feedback predictor and stepsize control logic in both the transmitter and the receiver.


Figure 2.'.0: Output SNR for PCM and one-bit $\triangle M$ and "white" or $R C$ input spectra versus the normalized bit rate $R_{b}=\ell 1_{s} / 2 f_{m}, P_{e}=0.0[4]$

Table 2.3 compares the l2-bit companded PCM codec with a CVSDM codec both having the same bit rate. It may be stated that for the three quality levels of program reproduction companded PCM is superior to CVSDM since it has high performance potential over the full signal bandwidth for the three quality levels specified. It is noted that even for the low quality level CVSDM does not meet the requirement of $T T / N \geqslant 36 \mathrm{~dB}$ (non-weighted) over the full 15 kHz bandwidth.
2.1.3 Comparisons Of The Proposed PCM Codecs With Prototype/ Commercial Radio Program Codecs

Base on the analysis in the previous sections it is recommended that digitally companded ( $\mu=255$ ) PCM codecs be used for digital radio program transmission.

Table 2.4 compares the codec performance characteristics of four existing companded $P C M$ systems $[8,9,10,11]$ with those of three PCM codecs proposed for the three required quality levels. The three quality levels required are:
(a) CATV reception $T T / N \geqslant 50 \mathrm{~dB}$ non-weighted
(b) High Quality Home Reception $T T / N \geqslant 47 \mathrm{~dB}$ nonweighted
(c) Low Quality Home Reception $T T / N \geqslant 36$ dB non-weighted
where the test tone (TT) level is the standard broadcast level of +8 dBm at 1 kHz . A 15 kHz bandwidth is considered for all quality levels and 5 kHz are considered for the low quality level only.

The four existing systems are:

| Feature | 12 Bit Companded PCM (32 kbps) |  |  | CVSDM (384 kbps) |  |  |  |  |  |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
|  | Low Quality | High Quality | CAIV Quality | Low Quality |  | High Quality |  | CATV Quality |  |
| Test tone frequency | DC to 15 kHz | DC to 15 kHz | DC to 15 kHz | 1 kHz | 15 kHz | 1 kHz | 15 kHz | 1 kHz | 15 kHz |
| 1. Non-weighted test tone to noise ratio (TINR) | $53 \mathrm{~dB} \text { al } 0^{-6}$ | $60 \mathrm{~dB} \mathrm{alO}^{-7}$ | $760 \mathrm{~dB} \mathrm{Cl}^{-7}$ | $\left\lvert\, \begin{aligned} & 58 \mathrm{~dB} \\ & \mathrm{al} 0^{-6} \end{aligned}\right.$ | $\left\|\begin{array}{rl} \leqslant 27 & d B \\ @ 10^{-6} \end{array}\right\|$ | $\left\lvert\, \begin{gathered} 62 \mathrm{~dB} \\ \mathrm{al0} 0^{-7} \end{gathered}\right.$ | $\left\|\begin{array}{rr} \leqslant 28 & \mathrm{~dB} \\ \text { @10 } \end{array}\right\|$ | $\left\|\begin{array}{l} 63 \mathrm{~dB} \\ @ 10^{-7} \end{array}\right\|$ | $\left\lvert\, \begin{gathered} \leqslant 28 \mathrm{~dB} \\ @ 10^{-7} \end{gathered}\right.$ |
| 2. Dynamic Range for tabulated values of BER and required TINR | 44 dB | 42 dB | $\sim 40 \mathrm{~dB}$ | 30 to 40 dB |  |  |  |  |  |
| 3. Error Control <br> Capabilities <br> (Also see Section 4.0) | A subjective improvement with error concealment using a single parity bit. Requires minor bit rate increase. Simple to incorporate |  |  | By Forward Error Correction (FEC) only. Requires greater bit rate increase. Significant additional circuitry required. |  |  |  |  |  |

NOTE: TINR requirements (nonweighted):' Low Quality, 36 dB ; High Quality, 47 dB ; CAIV Quality, 50 dB .

Table 2.3: PCM vs CVSDM For Radio Program Distribution by DBS (With The Same Transmission Rate)

| S SYSTEM | BBC | PB |  | SCIENTIFIC |  | PROPOSED CODEC |  |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
| P | NICAM3 | DATE | SONY | ATLANTA | CATV | HIGH QUALITY | LOW QUALITY |
| Baseband Nominal BW | 15 kHz | 15 kHz | 15 kHz | 15 kHz | 15 kHz | 15 kHz | $15 \mathrm{kHz}(5 \mathrm{kHz})$ |
| Sampling Frequency | 30 kHz | 34.42 kHz | 32 kHz | 32 kHz | 32 | 32 | $32 \mathrm{kHz}(11 \mathrm{kHz})$ |
| Transmission Rate (kbps) per Channel (no overhead) | 338 | 447 | 768 | 402 | 384 | 352 | $\begin{aligned} & 288 \\ & (99 \text { for } \\ & 5 \mathrm{kHz}) \end{aligned}$ |
| Digital Companding | 14 to 10 bits +1 overhead bit per sample | 14 to 12 <br> bits +1 <br> parity <br> bit per <br> sample | None. 13 +ll bits overhead per sample | $\begin{aligned} & 15 \text { to } 11 \\ & \text { bits } \end{aligned}$ | $\left\lvert\, \begin{aligned} & 15 \text { to } 12 \\ & \text { bits } \end{aligned}\right.$ | $\begin{aligned} & 14 \text { to } 11 \\ & \text { bits } \end{aligned}$ | $\left\lvert\, \begin{aligned} & 12 \text { to } 9 \\ & \text { bits } \end{aligned}\right.$ |
| BER <br> EC=Error Concealment <br> JPD=Just Percept- <br> ible Degradation | $\begin{aligned} & \leqslant 10^{-5} \\ & \text { for JPD; } \\ & \text { before } \\ & \text { EC } \end{aligned}$ | ```\leqslant10-5 for JPD; before EC.``` | $\begin{aligned} & \leqslant 10^{-4} \\ & \text { for JPD; } \\ & \text { before } \\ & \text { EC } \end{aligned}$ | $\begin{aligned} & \leqslant 10-5 \\ & \text { For JPD; } \\ & \text { before } \\ & \text { EC } \end{aligned}$ | $10^{-7}$ | $10^{-7}$ | $10^{-6}$ |
| Required non-weighted output TT/N | N.S. | N.S. | N.S. | N.S. | $\geqslant 50 \mathrm{~dB}$ | $\geqslant 47 \mathrm{~dB}$ | $\geqslant 36 \mathrm{~dB}$ |
| Dynamic Range for tabulated values of $T T / \mathrm{N}$ and BER | N.S. | N. S. | N.S. | N.S. | $\sim 40 \mathrm{~dB}$ | $\sim 42 \mathrm{~dB}$ | $\sim 44 \mathrm{~dB}$ |
| ( $\mathrm{S} / \mathrm{N}$ ) $\mathrm{PCM}-\mathrm{FB}$ <br> 25\% Quantizer <br> Loading; <br> Laplacian Signal PDF. | N.S. | N.S. | N.S. | N.S. | $\geqslant 60 \mathrm{~dB}$ | $\geqslant 55 \mathrm{~dB}$ | $\geqslant 43 \mathrm{~dB}$ |

Table 2.4: Comparison of Digitally Companded PCM Codecs For Digital Audio Program Distribution (N.S. - Not Specified)
(a) NICAM 3 [8] (Near Instantaneously Companded Audio Multiplex, Mark 3) - a BBC developed digital sound programme transmission system for use, in particular, on $2048 \mathrm{kbit} / \mathrm{s}$ digital bit streams. .The system offers high quality audio and provides six sound programme channels on a $2048 \mathrm{kbit} / \mathrm{s}$ bit stream by using 30 telephone chanels.
(b) DATE [9], (Digital Audio for Television), a joint project of Public Broadcasting Service (PBS) and Digital Communications Corp, for a multi-channel digital audio system, multiplexing up to four 15 kHz audio channels onto a single TV video channel, to provide a combined audio with video transmission system.
(c) A Sony Corporation prototype of a multi-channel radio program transmission system for distribution by satellite [10].
(d) Digital Audio 3 Meter Earth Station System manufactured by Scientific Atlanta [ll], designed for reception of media information broadcast via satellite using digital data techniques.

In selecting the number of bits per sample ( $n$ ) for the recommended codecs it was ensured that the required TTNR would be safely achievable even with some BER degradation. The nominal BER selected for each channel is a practical requirement for satellite links and may be achieved with or without error correction schemes. Moreover, the $n$ and BER selected allow for an input signal dynamic range of 40 to 44 dB over which the codecs performance would meet the required TTNR performance. The dynamic range is achievable by employing the $\mu=255$ digital companding technique.

The recommended BER is $10^{-7}$ for CATV and high quality receptions and $10^{-6}$ for low quality home reception. These BER values will be used with results in [12] and [13] to find the required $\mathrm{C} / \mathrm{N}_{\mathrm{o}}$ for the candidate modulation techniques (coherent and differential PSK, MSK) in Section 3.0. Furthermore, subjective studies indicate that even at these low bit error rates channel errors still cause perceptible (audible) performance degradation. Forward error correction (FEC) and/or error concealment (EC) techniques (such as previous sample substitution upon error detection in current sample) may be employed to mitigate the subjective degrading effect of channel errors. Error correction and concealment techniques will be discussed in Section 4.0.

## 2.2

2.2.1 Summary of Multiplexed Rates

Radio program distribution employing the $T D M$ technique applies to the following scenarios
(i) TiDM carriers in a dedicated transponder.
(ii) TDM subcarrier located above the highest TV baseband frequency in the shared transponder mode.

In both scénarios, each TDM carrier (or subcarrier) can carry up to 5 radio programs.

Table 2.5 provides the bit rate of the various TDM bit streams according to program quality level and number of channels multiplexed.

The following describes and evaluates the options for . multiplexing digitized radio program data streams.

|  | QUALITY LEVEL |  |  |
| :---: | :---: | :---: | :---: |
| NUMBER OF CHANNELS | CATV | HIGH | LOW |
| 1 | 384 | 352 | 288 |
| 2 | 768 | 704 | 576 |
| 3 | 1152 | 1056 | 864 |
| 4 | 1536 | 1408 | 1152 |
| 5 | 1920 |  | 1760 |

Table 2.5: Bit rate (kbps) for 1 to 5 channels TDM bit stream
2.2.2 Continuous and Packet Multiplexing Methods
2.2.2.1 Descriptions

Two basic multiplexing concepts that can be used are referred to as "continuous" and "packet" [14]:
(i) In continous multiplexing the basic structure is the frame in which fixed groups of bits play prescribed roles according to their position in the frame. Thus, particular bits are dedicated to convey the 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 convey each of the signals multiplexed. In continuous multiplexing the bit rate of each signal may be a precise sub-multiple of the final serial bit rate which is known as synchronous mutliplex. When this condition does not apply, a process of a synchronous mutliplexing is possible by arranging certain bits within the frame to carry either real or dummy information according to a control signal. Continuous multiplex.systems require only a small proportion of overhead bits for signalling such as framing. The NICAM-3 system [8] uses $7 \mathrm{kbit} / \mathrm{s}$.out of a total of $2048 \mathrm{kbit} / \mathrm{s}$ for framing.

A continuous multiplexing system is wellsuited for sound channels since sound channel reassignments are not likely to occur due to the continuous nature of the sound. With a low overhead required for secure synchronization a small capacity
can be used for signalling to indicate change of use of channel and thus provide flexibility. The capacity of any channel not used for sound can be assigned to data. This method, therefore, does not involve any increase in overhead to indicate the presence of data, it exploits to the utmost the channel capacity for the useful information, and allows for simplicity and stability in the receiver design.

In packet multiplexing, the final bit stream is composed of successive blocks, called packets, with two parts: heading and data parts. Each packet conveys data from only one input signal and the selection in the receiver is realized by detection in the heading of the address of the desired service; this method does not impose a predetermined content on the final bit stream. In systems designed for broadcasting applications, the length of each packet is constant. The parkets are transmitted 'synchronously' which means that there is continuity of phase of the binary sequence and of the modulated carrier, between packets or groups of packets. The packets are also transmitted with regular periodicity. The demúltiplexing process then becomes a synchronisation process similar to that used in a continuous multiplex structure. This synchronisation can be attained in a conventional manner by the use of a loop locked to a synchronisation flag in the heading of each packet. This restriction of fixed overall packet length does not imply a fixed length of the section containing useful data.

Within the constraints of periodic packet transmission, the rate at which packets are transmitted for a particular service is related directly to the bit rate of the input signal of the associated service. In the packet multiplexing method, it is possible to operate synchronously or asynchronously. In the latter case, the bit rates of the various input signals do not need to be related directly to the final serial bit rate. Thus, no special provisions are necessary for accommodating asynchronous signals. The content of the multiplex does not need to be pre-determined or fixed and can be changed at any time to be adapted to the needs. In the case of asynchronous operation for sound services (e.g. for insertion of sound signals in a video line-blanking interval), the packet multiplexing method inherently permits the asynchronous insertion into the final bit stream but the recovery of the sampling frequency needs resynchronisation processes.

A packet contains two parts:
(a) A heading, specific to the transmitter, which serves to synchronize the receiver and to identify the source of the data inserted in the packet and for the transmission of other information. The heading contains a synchronisation pattern followed by a prefix.
(b) A section containing useful data;

The synchronisation pattern enables the receiver to extract the bytes comprising the packet. It has a role equivalent to that of a locking loop in the case of a continous multiplex. In periodic transmission, it can also be used for the synchronization of the loop which serves for the
demultiplexing process. It should be noted also that the possibility of using this pattern to remove the ambiguity due to certain demodulation processes; such an arrangement which may be applied to continuous multiplexing too, is advantageous in that it avoids the coding of transitions and, hence, a certain propagation of transmission errors.

The role of the prefix is to characterize and identify the semantic content of the packet and, more specifically, the source of the data inserted in the useful data section. Several prefix configurations can be envisaged. In particular, it may be possible to use the prefix configurations considered for data broadcasting.

The increased flexibility to the data of the packet multiplex system is obtained at the price of increased channel overhead for the packet headers which includes, in normal operation, programme identifiers. A typical useful channel efficiency of $90-95 \%$ is possible.
2.2.2.2 Sensitivity to Errors

The continuous multiplex process is susceptible to two sorts of impairments due to bit errors:

- output $S / N$ degradation
- frame synchronization loss.

Errors introduced into multiplex signals can cause loss for a certain number of consecutive frames at the demultiplexer when the framing pattern is not recognised. Careful attention to the design of framing circuits can minimize the risk of losing synchronization; for example, the decoding equipment for the NICAM 3 system mentioned earlier remains synchronized until the bit-error rate approaches 1 in 10.

The packet broadcasting process is susceptible to two sorts of impairments due to bit errors as follows:

- output $S / N$ degradation
- packet loss.

For each source a digital channel identification is carried in the packet prefix. In the demultiplexer, the packets are selected by analysis of this identification. Errors in this information, can result in poor recognition of the address of the transmission source and hence, in the loss of the packet; a loss of a certain number of successive bytes results (the number depends on the format of the data section of the packet (s) lost).

Several tests of broadcasting digital sound signals with packet multiplexing and with a modulation system appropriate for satellite broadcasting (with television pictures) have been made in the laboratory as well as with the OTS satellite [15]. It has been shown that the packet losses appreciably degrade the sound quality only at levels of the carrier-to-noise ratio that are below the FM threshold, when the television picture is already severely impaired.
2.2.2.3 Comparison of Multiplexing Methods

Table $2.6^{\circ}$ compares the merits of the two types of multiplexing methods. The features in the table have been selected to cover various aspects of possible multiplexing systems adapted for broadcasting. It is obvious, however, that due to the state of the study and the development in some areas, the table entries are inevitably tentative.


Table 2.6: Comparison Between Continuous and Packet Multiplexing.

The packet multiplexing system inherently offers a greater flexibility with a receiver of medium complexity at the price of bit rate overhead. Both continuous and packet multiplexing systems require only simple synchronisation circuits at the receiver if the sampling frequencies of the signals to be combined are synchronous or can be synchronized. Each system is in principle vulnerable to the effects of channel errors, but frame loss and/or packet loss are not likely to occur in normal operation (see Section 2.2.3.2). Summarizing, continuous multiplexing offers a system which is most efficient for sound transmission and results in a simple receiver.

## 2.2 .3 <br> Framing, Synchronization and Bit Rate Adjustment in Continuous Multiplexing

Regardless of the specific multiplexer design, the following characteristics are common to all multiplexers:
(i) The basic mutliplexed structure is a frame, representing the smallest unit of time in which all bit streams to be mutliplexed are serviced at least once.
(ii) The frame is comprised of time slots allocations, which are dedicated to each data source. A timing discipline controls the transfer of a string of bits from each data source to its dedicated time slot in the frame on a periodic basis.
(iii) Framing and synchronization (or control) bits are appended to the frame; these enable the receiving system to synchronize to the frame structure.
(iv) The multiplexer should tolerate small rate variations in each of the incoming bit streams.

In the design of a multiplexer for use with up to 5 digital radio input program channels, the basic problems arising in the multiplexing of independent digital bit streams are to be addressed, namely, the problems of:
(i) framing,
(ii) synchronization and
(iii) rate adjustment to accomodate small variations in the incoming bit rates.
2.2.3.1 Bit Rate Adjustment

In a bit interleaving mutliplexer, the time slots dedicated to each bit stream within the frame structure are to be filled by a fixed number of bits of the particular bịt stream. In reality independent data sources may experience slight relative variations in their data rates. The multiplexer must accomodate for these variations, otherwise, frame synchronization may be lost and data misalignment may occur in the receiver.

The problem is handled by
(i) Operating the multiplexer output bit stream at a rate which is slightly higher than the sum of the maximum expected rates of the input channels.
(ii) To accomodate small reductions in the input rates as well as to handle the nominal rates, bit stuffing on a per channel basis is often used. Even when the data sources run at their nominal rates, the fact that the multiplexer operates at a rate slightly higher than the input rate implies that,
occasionally, non-information carrying bits should be stuffed into the bit stream. Within the frame structure there are dedicated slots, associated with each input, in which a stuff bit may be inserted. By means of additional overhead bits the receiver is advised, with respect to each channel, whether the bit in the associated stuff bit slot is a dummy bit to be ignored or an information carrying bit to be retained and output.

The multiplexer output rate $R_{o}$ required to accomodate the possible input clock rate increase is given by [2]:

$$
\begin{equation*}
R_{o}=m R_{1}(1+\delta)\left(\frac{I+X}{I}\right) \tag{2.17}
\end{equation*}
$$

where:
$R_{o} \triangleq$ the multiplexer output rate.
$\mathrm{R}_{1} \triangleq$ nominal input bit rate per channel.
m $\triangleq$ number of multiplexed channels.
I $\triangleq$ number of information bits per frame.
$\mathrm{X} \triangleq$ number of control bits per frame.
$\delta$. $\triangleq$ fractional bit rate increase per channel.

It is obvious that

$$
\begin{equation*}
\delta=\frac{S}{I / m} \tag{2.18}
\end{equation*}
$$

where $S$ is the number of stuffing bits per channel per frame.

If the control bits are about $2 \%$ of the total bit stream, $R_{1}=384 \mathrm{kbps}, m=5$ and $\delta=100 \mathrm{ppm}$ then $\mathrm{R}_{\mathrm{o}}=1959.38$ kbps. If the frame length is $L=1000$ bits then $I=980$ and $S=(I / m) \delta=1.96 \times 10^{-2}$ stuffed bits per channel per frame. In other words, there is a stuffed bit for each channel every 51 frames on average.

### 2.2.3.2 Frame Synchronization

Frame synchronization should be maintained by the receiver demultiplexer so that a specific radio program channel from all those multiplexed can be selected. Sychronization implies that the transmitter and receiver clocks are locked. to one another so that bit integrity is maintained.

Due to channel noise and other system impairments there is a probability that a frame will be lost owing to errors in the framing pattern. The longer the framing pattern is the less reliably synchronization is maintained. On the other hand longer frame patterns are easier to synchronize, therefore, there is a tradeoff in synchronization pattern length. In order to keep the frame length within practical limits for fast acquisition purposes, it is usually required that frame violation is detected $k$ times in succession before a decision is made that frame synchornization was lost.

The probability, $P_{\ell}$, that framing is lost is given by:

$$
\begin{equation*}
P_{\ell}=(n p)^{k} \tag{2.19}
\end{equation*}
$$

where
$n \triangleq$ the framing pattern length in bits in each frame.
$p \triangleq$ the probability of a channel bit error ( $p \leqslant \leqslant 1$ )
$\mathrm{k} \triangleq$ the number of successive frame loss detection before a frame loss is declared present.

If $F$ frames per second are transmitted then the mean time between frame loss is given by:

$$
\begin{equation*}
\mathrm{T}_{\ell}=\frac{1}{\mathrm{P}_{\ell} \mathrm{F}} \tag{2.20}
\end{equation*}
$$

Once a loss of frame synchronization occurs; some time is required to reacquire framing. It is desired that this time be as short as possible. If frame synchronization is lost there is the probability of $2^{-n}$ that $n$ bits of the data stream may look like the expected $n$ bits framing synchronization pattern. The probability that frame loss is detected is then

$$
\begin{equation*}
P_{D \ell}=\left(1-2^{-n}\right)^{k}=1-k 2^{-n}+\ldots \tag{2.21}
\end{equation*}
$$

Since we want $P_{D l} \simeq l$, it is required that

$$
k 2^{-n} \leqslant \leqslant 1
$$

The time required to re-establish frame synchronization is the time required to examine the incoming bit stream, $n$ bits at a time, until the frame synchronization pattern is identified as such.

Usually it is required that a framing pattern, once detected will appear consecutively at the same position in the frame $m \geqslant l$ times. Using $m \geqslant l$ reduces the probability of false synchronization. If in the search process the frame
pattern is erroneously detected, the probability that it will be detected again at the same location is $2^{-n}$. For practical values of $n$ only one frame interval is required to detect false synchronization, and resume the search. If $h$ such suspensions of search or "holds" are encountered during the $L$ bit search, then the worst case acquisition time is:

$$
\begin{equation*}
T_{a}=m+1+h \text { Frames } \tag{2.22}
\end{equation*}
$$

It is noted that ( $L+h$ ) pattern must be examined ( $L$ is the number of bits in a frame) before synchronization is achieved. The ratio $h /(L+h)$ is an estimate of the probability of detecting a framing pattern. Therefore,

$$
h /(h+L)=2^{-n}
$$

or

$$
h=\frac{L}{2^{n}-1}
$$

and

$$
T_{a}=m+1+\frac{L}{2^{n}-1} \text { Frames }
$$

The total time to detect frame loss and acquire is then

$$
\begin{equation*}
T=k+m+1+\frac{L}{2^{n}-1} \tag{2.23}
\end{equation*}
$$

For example, if $L=1000, \mathrm{n}=8, \mathrm{~m}=2$ and $\mathrm{k}=3$ then $\mathrm{T}=10$ frames ( 10,000 bits). At 2 Mbps it is $5 \mathrm{milliseconds}$. $\mathrm{p}=10^{-7}$ then the chance of losing synchronization is given by $P_{\ell}=\left(8.10^{-7}\right)^{3}=5 \times 10^{-19}$. In this case, where the
frame rate is $F=2 \times 16^{6} / 1000=2000$ frames per second, frame loss will occure every $T_{\ell}=1 / P_{\ell} F=1015$ seconds, a result which indicates that once frame synchronization was established interruptions due to random noise are vitrually non existent.

An elaborate computer aided study of framing codes with minimum associated probability of false synchronization (and therefore low out-of-phase partial cross-correlation property) is presented in [16]. Table 2.7 lists the resulting optimum codes of length 7 to 30 with their associated agreement (partial crosscorrelation) vector and probability of false synchronization (PFS). The PFS was calculated assuming that up to $\varepsilon=2$ errors may be included in the n bit string examined by the receiver. The string considered are those that have 1 to $n-1$ overlapping code bits with the full code pattern. Figure 2.11 is a graphic representation of the PFS as function of $n$ for several values of $\varepsilon(\varepsilon=0,1,2,3)$. The probability of an error was chosen as .l0. It is observed that for a 15 bit code, allowing $\varepsilon=2$, the PFS is $10^{-2}$.

### 2.2.3.3 Examples of Continuous Multiplexing Schemes

The following is a comparative description of multiplexing techniques which have been implemented for digital program distribution.

NICAM 3 [8] - Three blocks of 32 samples plus various housekeeping bits form a single channel sub-frame in a total time of 3 ms . The bit rate per channel is 338 kbps and, therefore, there are $\left(3 \times 10^{-3}\right) \times\left(338 \times 10^{3}\right)=1014$ bits per frame allocated as shown in Table 2.8. With 6 multiplexed channels, overall framing ( $7 \mathrm{kbits} / \mathrm{s}$ ), signalling (l kbits/s) and justification (l2 kbits/s, the

PROBABILITY OF FALSE SYNC $\quad 5.723 \times 10^{-1}$
AGREEMCNT VECTOR (03 02 O2 0101001
CODE $1011: 000$
PROBABILITY OF FALSE SYNC $\quad 4.235 \times 10^{-1}$ LENGTH O8

PROBABILITY OF FALSE SYNC $\quad 2.950 \times 10^{-1}$

PROBABILITY OF FALSE $5 Y N C \quad 1.783 \times 10^{-1}$

PROBABILITY OF FALSE SYNC $\quad 9.065 \times 10^{-2}$

PROBABILITY OF FALSE SYNC $\quad 5.142 \times 10^{-2}$

PROBABILITY OF FALSE SYNiC $\quad 2.821 \times 10^{-2}$

PROBABILITY OF FALSE SYNC $\quad 1.514 \times 10^{-2}$

PROBABILITY OF FALSE SYNC $6.611 \times 10^{-3}$

PR:OBABILITY OF FALSE SYNC $\quad 3.460 \times 10^{-3}$

PROBABILITY OF FALSE SYNC $\quad 1.657 \times 10^{-3}$

PROBABILITY OF FALSE SYNC $8.222 \times 10^{-4}$
CODE 111100110101000000

CODE 1111100110010100000
PROBABILITY OF FALSE SYNC $\quad 3.837 \times 10^{-4}$


CODE $11101101111000100000 \quad$ PROBABILITY OF FALSE $5 Y N C \quad 2.175 \times 10^{-4}$


CODE 111011.101001011000000
PROBABILITY OF FALSE SYNC $\quad 1.051 \times 10^{-4}$


CODE 11110011101101010000000
PROBABILITY OF FALSE SYNC $\quad 4.905 \times 10^{-5}$


CODE 11110101110011010000000
PROBABILITY OF FALSE SYNC $\quad 2.533 \times 10^{-5}$

CODE $111110101.11001100100000 \quad$ PROBABILITY OF FALSE SYNC $1.255 \times 10^{-5}$

CODE 1111100101101110001000000
PROBABILITY OF FALSE SYNC $\quad 6.449 \times 10^{-6}$

CODE 1 i 11101001101011000.1000000
PROBABILITY OFFALSE SYNC $3.144 \times 10^{-6}$


COOE 111110101101001100110000000
PROBABILITY OF FALSE $5 Y N C \quad 1.583 \times 10^{-6}$
 COUE 1 11 i 010111100101100110000000 PROBABILITYOFFALSE $5 Y N C$ E.OSS $\% 0^{-7}$

CODE 1111010111110011001101100000000 PROBABILITYOFFALSE SYNC $4.093 \times 10^{-7}$
 CODE : $1: 110101111001100110: 90000000$ PROSAELLITY OF FALSE SYNC $2.0: C \times 10^{-7}$


LENGTH 11

LENGTH 22 LENGTH 25 LENGTH 26
LENGTH 07 LENGTH OP LENGTH 10 LENGTH 12 LENGTH 13 LENGTH 14 LENGTH 15 LEINGTH 16 LENGTH 17 LENGTH 18 LENGTH 19 LENGTH 20 LENGTH 21 LENGTH 23 LENGTH 24 LENGIH 27 LENGTH 28 LENGTH 29 LENGIH 30


Figure 2.11 - Graphic Representation of $F_{\phi} \quad$ [16]

| Frame Element | Bits/ <br> Frame | Bit Rate/ <br> Channel <br> (kbps/channel) | Remarks |
| :---: | :---: | :---: | :---: |
| a. Sample words | 960 | 320 | $\begin{aligned} & \text { l0 bits/sample x } 32 \\ & \text { samples/block x } 3 \text { blocks } \\ & =960 \end{aligned}$ |
| b. Companding Range | 11 | 3.6 | 5 possible companding ranges per block. The number of bits required to identify the range for each block is $\log _{2}(5$ $=7$ bits. These 7 bits are error protected by additional 4 bits. |
| c. Sample Word Error Protection | 32 | 10.6 | 32 bits have been allocated for sample word error protection, together with a simple error concealment system. |
| a. Sub Framing | 7 | 2.3 | 7 bit framing pattern; different patterns for alternate frames. |
| e. Signalling | 4 | 1.3 | Facilitates an auxiliary low rate channel. |
| TOTAL | 1014 | 338 | Overhead bits 4\% |

Table 2.8: NICAM-3 Channel Sub-Frame
total bit rate is 2048 kbits/s. The justification bit stream is the overhead required to synchronize the 6 asynchronous channel bit streams. An overall framing time of about 2 ms is expected after frame loss. As indicated by Table 2.8 the overhead bits constitute $4 \%$ of the bit stream.

Figure 2.12 and Figure 2.13 describe the single channel frame and the master frame for the NICAM-3 system. Figure 2.14 shows some NICAM-3 possible multiplexing configurations.

Sony [10] - Formats for one stereo channel and 12 stereo channels are shown in Figure 2.15 and Figure 2.16, respectively.

In this prototype system, the transmitted TDM stream is arranged such that bits from each channel are grouped according to their sample time slot. With this format, even if a burst error is generated in the transmission system, it is converted into random errors when the signal of an individual channel is selected. "SB" in the figure is an abbreviation for "service bit". Identification of the broadcasted program contents and automatic ON/OFF loudness control is made with the SB bits.

The Sony-developed error-correction code applied here is one in which BCH and parity are combined to yield an extremely low error rate after correction. Even if the bit error rate is around $10^{-3}$ miss-correction is generated only once in several hours.

| R1 | $15 \times 10 \mathrm{Bits}$ | R3 | $16 \times 10$ Bits | R5 | $16 \times 10$ Bits | R7 | $16 \times 10 \mathrm{Bits}$ | R9 | 6× 10 Biss | R11 | $16 \times 10$ Bits |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
| $P_{1}$ |  | P2 |  | P3 |  | P4 |  | $P 5$ |  | F |  |
| P6 |  | P7 |  | PB |  | P9 |  | P10 |  | F |  |
| P11 |  | P12 |  | P13 |  | P14 |  | P15 |  | $F$ |  |
| P16 |  | P17 |  | P18 |  | P19 |  | P20 |  | $F$ |  |
| P21 |  | P22 |  | P23 |  | P24 |  | P25 |  | F |  |
| P26 |  | 927 |  | P28 |  | P29 |  | P30 |  | F |  |
| P31 |  | P32 |  | 51 |  | 52 |  | 53 |  | $F$ |  |
| R2 |  | R4 |  | R6 |  | R8 |  | R10 |  | 54 |  |

The sample bits are shown as groups of $16 \times 10$ bits.
F = Froming Bit.
$P=$ Somple Parity Bit.
$R=$ Ronge Code Bit.
$\mathrm{S}=$ Signalling Bit.

## NOTE:

Single channel framing is confained in a mulliframe - Frame 0 and 1. Framing bits are different for eoch frame although their positions in the bitstream sequence are identical.

Figure 2. 12: NICAM.3. SINGLE CHANNEL FORMAT.

## NOTE:

Six channel multiplex framing is contained in a multiframe -Frame $A$ and $B$. Framing bits are dilferent for each frame although their positions in the bitstream sequence are identical.
$F=$ Framing sit.
$S=$ Signal Bit.
$\int_{C}^{X(Y, Z)}=$ Justilication Conlrol Bit For Channel Pair $X(Y, Z)$.
Bit Ratès:
$C_{X(Y, Z)}=$ Channel Bit From Coder Pair $X(Y, Z)$.
3 Coder Pairs $=3 \times 676=2028$ kbit $/ \mathrm{s}$
Framing $=7 \mathrm{kbit} / \mathrm{s}$
$\int_{D}^{X(Y, Z)}=\begin{aligned} & \text { Negalive Justificalion Service Bit For } \\ & \text { Channel Pair } X(Y, Z) \text {. }\end{aligned}$
$C_{X(Y Z)}=$ Posilive Justification Service Bit For
Juslificalion $=3 \times 4=12 \mathrm{kbit} / \mathrm{s}$

Signalling $=\frac{1 \mathrm{kbit} / \mathrm{s}}{2048 \mathrm{kbits}}$

Figure 2.I3: : NICAM.3. SIX CHANNEL MULTIPLEX FORMAT.


Fígure 2:.14: NICAM 3. SOME POSSIBLE ARRANGEMENTS.


Figure 2.15: Transmission Word -. Sony System


Figure 2.16: All Chamel Tranmission Format - Sony System

By using this code in this prototype system, sufficient reproduction signals are obtained even when the receiver $\mathrm{C} / \mathrm{N}$ is as low as 7 dB .

Scientific Atlanta [17] - The block diagram of Figure 2.17 shows the basic system which will accept audio programs and/or data inputs at the network studio and distribute them to the network affiliates.

The network studio contains the studio link multiplexer, the channel units and a stable clock source. The studio link multiplexer accepts four 384 kbps inputs and combines them with synchronization information to form a Tl (1.544 Mbps) output (1.536 Mbps information bits and 8 kbps overhead). The Tl data stream is routed to a studio link demultiplexer at the transmit satellite earth station, where the Tl stream is separated into its component channels. The demultiplexed data is sent to the TDM multiplexer where it is combined with the outputs of other studio link demultiplexers to form one 7.68 Mbps data stream. The TDM multiplexer has the capacity to handle 19 complete 384 kbps channels and one additional 384 channel in which a 32 kbps slot is reserved for system synchronization. It is noted that the overhead bits constitute less than $1 \%$ of the total bit stream. In the receiver, the 7.86 Mbps data stream is applied to the TDM demultiplexer, where it is separated into the individual data streams. The actual sub and master frame detailed structure for the Scientific Atlanta system was not available.


Figure 2.17: Digital Audio System Concept Simplified Block Diagram

### 3.0 DIGITAL MODULATION TECHNIQUES FOR RADIO PROGRAM DISTRIBUTION

Properties of PSK and MSK Modulation Techniques

In this section a number of the key aspects of the candidate modulation techniques are examined with respect to their impact on a DBS-type radio program distribution system. The primary issues relate to power and bandwidth requirements, modem implementation and. system requirements imposed by the use of these modems. We begin with a tabulated comparison of these modulation technigues and a survey of those that have been used in prototype radio program distribution

Two- and 4-phase PSK and MSK (FFSK) with coherent and differential detection are compared in Tables 3.la and 3.lb with respect to performance and implementation aspects. This information will be used to assess the modems in subsequent sections.

A number of prototype systems have been developed for radio program distribution. Some background information is given below for systems using a TDM subcarrier located above the TV video baseband signal.

In order to minimize video and audio mutual interference it would seem that the modulation techniques characterized by a narrow power spectrum, namely OPSK, OQPSK and MSK would be more suitable [18] than BPSK (with either coherent or differential detection). A number of tests using the OTS system have been performed. As reported in [19], a 27 MHz bandwidth transponder was used for FMTV* with one digitally modulated TV subcarrier. Several high quality sound channels were multiplexed and the performance with QPSK and MSK digital subcarriers was evaluated. These tests were

[^1]- 66 -

| CHARACTERISTIC | MODULATTION TECHIJIQUE |  |  |  | REMARKS |
| :---: | :---: | :---: | :---: | :---: | :---: |
|  | BPSK | QPSK | OOPSK | MSK (FFSK) |  |
| (1) Unfiltered Signal Format <br> 2) Duration of transmitted symbols ( $\mathrm{T}_{\mathrm{s}}$ ) <br> 3) phase transitions possible from symbol to symbol | $s(t)=a_{n} \cos \left[\omega_{c} t+\theta\right]$ $T_{S}=T$ $0^{\circ}, 180^{\circ}$ | $\begin{aligned} & s(t)=a_{I} \cos \left[\omega_{C} t+\theta\right] \\ & +a_{Q} \sin \left[\omega_{c} t+\theta\right] \\ & T_{s}=2 T \end{aligned}$ | The same as QPSK with I and Q channel pulses offset by T $T_{s}=2 T$ $10^{\circ}, \pm 90^{\circ}$ | $\left\{\begin{array}{l} s(t)=C_{I} \cos \left[\omega_{C} t+\theta\right] \\ \quad+C_{Q} \sin \left[\omega_{C} t+\theta\right] \\ \text { pulses of I and } Q \\ \text { channels are offset } \\ \text { by } T, \text { or } \\ s(t)=\cos \left[\omega_{C} t+b_{k} \frac{\pi t}{2 T}+t_{k}\right] \end{array}\right]$ <br> where $b_{j}=+1$ when $\begin{aligned} & C_{I_{k}}=C_{Q_{k}}^{\prime} \quad b_{k}^{\prime}=-1 \\ & \text { when } C_{I_{k}}^{\prime}=-C_{Q_{k}} \end{aligned}$ $\mathrm{T}_{\mathrm{S}}=2 \mathrm{~T}$ $\theta_{k}=\theta_{k-1}+b_{k} \frac{\pi k}{2}$ <br> i.e. continuous phase at transition; $\pm \pi / 2$ phase shift during bit interval | $T$ is a source bit period. $\left\{a_{n}\right\},\left\{a_{I}\right\},\left\{a_{Q}\right\}$ are rectangular pulses of +1 or -1 polarity and duration 2 T , except $\left\{a_{\mathrm{n}}\right\}$. which are $T$ long. $\left\{C_{T}\right\},\left\{C_{Q}\right\}$ are half sine shaped pulses of +1 or ${ }^{-1}$ polarity and duration 2T. <br> Data can be differentially encoded encoded prior to modulation. |

Table 3.1(a): Properties of Modulation Techniques for RPD (continued next page)

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Table 3.1(a): Properties of Modulation Techniques for RPD (continued)

## MODULATION TECHNIQUES

|  | MODULATION TECHNIOUES |  |  |  |  |  |  |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
| Criterion | BPSik | QPSK | DPSK | DQPSK | OQPSK | MSK | DMSK |
| Complexity of modulator (excluding filters) | Simple rectangular base-band pulses amplitude modulate one carrier | More complicated than 2 phase PSK; both I and Q channels are required | Same as BPSK DPSK refers to detection; requires differential encoding | Same as QPSK. Requires differential encoding | Same as QPSK; relative delay of 1 bit between I\&Q channels required. | Most complicated Requires sinusoidal pulses in OK-QPSK modulator | Same as MSK - DMSK refers to differential detection |
| Complexity of demodulator | Fairly simple. Signal is detected in one channel. Except for possible differential decoding no further logical operation is needed after detection | More complex than 2 CPSK since both I and $Q$ channels are required | Simplest. Detection is based on phase comparison between adjacent symbols | Conceptually the same as DPSK requires twice as much circuitry | About the same order as QPSK (slightly more complex timing) | More complex carrier recovery circuits and detection circuits than OOPSK | essentially the same as DPSK |
| Carrier <br> Recovery <br> loops | Requires 1 coherent carrier recovery loop. Requires doubling of the carrier frequency and a PLL or a Costas loop. | Requires one coherent carrier recovery loop Requires quadrupling the carrier frequency and a PLL or a Costas loop | No coherent carrier recovery loop required. Use delay line and multiply to accomplish differential detection | Conceptually the same as DPSK. Requires quadrature delay and multiply circuit | Requires a coherent recovery loop, same as QPSK | Requires doubler and two coherent carrier recovery loops or Costas loop equivalent. | essentially the same as DPSK |

[^2]|  | modulation techinioue |  |  |  |  |  |  |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
| Modem <br> Criterion | BPSK | QPSK | DPSK | DQPSK | OQPSK | MSK | DMSK |
| bit timing recovery circuits | IF or baseband circuit required to recover timing based on data transitions | same as BPSK | same as BPSK | Same as BPSK | same as BPSK | same as BPSK | same as BPSK |
| ideal detector <br> (matched <br> filter) <br> requirements | One integrate and dump filter One sampler | Two integrate and dump filters and two samplers | IF filter required and post-multiplier lowpass filter and sampler | Requires two post multiplier lowpass filters and samplers | Two integrate and dump filters and samplers. Integrate and dump and sampling actions in I and Q channel are staggered by one bit | Two sinuisoidalweighted integrate and dump filters; operation similar to OK-QPSK | same as 2 DPSK |
| differential coding to resolve data ambiguities* | Differential coding needed to resolve ambiguity | Differential coding of $I$ and $Q$ channels needed to resolve ambiguities. Logic more complicated | Data is already coded in differertial form: See comment for DMSK | Differential coding of I and Q channels needed. | same as QPSK | same as BPSK | See MSK. Also, Non-redundant errorcorrection can be used to reduce $\mathrm{E}_{\mathrm{b}} / \mathrm{N}_{\mathrm{o}}$ degradation from differential detection of MSK [22] |

Table 3.1(b): Issues of Modem Implementation (continued)
*Can be avoided for coherent detection if use sync word detection to indicate reference phase
conducted by IRT-West Germany, TDF-France and the BBC in the U.K. Also reported in [19] by RAI, TDF and the BBC, was a QPSK system.

The QPSK system used a 2 Mbps subcarrier at 7.5 MHz accompanying the video signal (625 line system). The video FM deviation was $13.5 \mathrm{MHz} / v o l t$ and the maximum subcarrier level was limited to 320 mv to avoid visible interference to the picture and at the same time to obtain an acceptable BER. Under these conditions the achievable BER was $10^{-7}$ with $C / N=14 \mathrm{~dB}$. However, the $B E R$ increased rapidly with a slight misalignment of the IF filter. With error control, a good sound quality was maintained even with $\operatorname{BER}=10^{-3}$, with no filter phase correction. The 2 Mbps bit rate supported 6 independent high quality sound channels.

The MSK modulation technique studied also used a 2 Mbps bit stream, but modulated a 6.656 MHz subcarrier accompanying the video signal. With the same video deviation (13.5 $\mathrm{MHz} / \mathrm{volt}$ ) and a subcarrier level ( 300 mv ), the BER was below $6 \times 10^{-5}$ at $C / N=14 \mathrm{~dB}$. Subjective tests under these conditions showed that a good quality of sound was achievable even at $C / N=10$ dB (a simple error control technique was employed).

A third experiment was performed by the BBC using a $4 \phi$-DPSK (DQPSK) 704 kbps subcarrier located between 6 and 7.5 MHz , to provide 2 sound channels. Under these conditions the BER was below $5 \times 10^{-5}$ for $C / N=14 \mathrm{~dB}$, providing a basis for a good sound quality.

The conclusion derived from these three tests was that in order to provide good sound quality the subcarrier frequency should be about . 5 MHz above the video between 6.5 to 7 MHz and a peak deviation of 1.4 MHz (at least) should be used. Depending upon the sound quality and
modulation technique used, 2 to 6 programs could be transmitted. It is indicated in other studies that an upper limit exists and is determined by the second harmonic of the colour sub-carrier.
3.2 Quantitative Performance Comparison of PSK and MSK Modulation Techniques

To evaluate the effect of a modulation technique on system performance it is necessary to establish the power and bandwidth requirements of the modulation technique. These parameters are determined by ideal or theoretical values as well as by practical implementation margins.

The ideal error rate performance of a modulation technique refers to the theoretical value of $E_{b} / N_{o}$ required to attain a specified BER with additive white Gaussian noise being the only source of degradation. Table 3.2 provides the minimal values of $E_{b} / N_{o}$ required to achieve the $B E R$ requirements for the three quality levels by the various candidate modulation techniques. Also shown are the corresponding values of $C / N_{o}$ for the various TDM serial rates.

With respect to bandwidth occupancy, some care is required. In the case of PSK modulations, raised-cosine filtering may be used to improve bandwidth occupancy over the unfiltered rectangular data pulse case. For example, $50 \%$ roll-off filtering is often used and will limit the occupied bandwidth to $1.5 \mathrm{R}_{\mathrm{S}}$ ( $\mathrm{R}_{\mathrm{S}}=$ symbol rate). With ideal filters no error rate degradation will be incurred.

In the case of MSK the bandwidth can be limited by filtering. However, as this bandwidth is reduced below 1.5R an error rate degradation will result.

| QUALITY LEVEL @ NO. OF CHANNELS PARMETER |  | CAIV |  |  |  |  | HIGH QUALITY |  |  |  |  | LOW QUALITIY |  |  |  |  |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
|  |  | 1 | 2 | 3 | 4 | 5 | 1 | 2 | 3 | 4 | 5 | 1 | 2 | 3 | 4 | 5 |
| R BIT RATE (kbps, no overhead bits) |  | 384 | 768 | 1152 | 1536 | 1920 | 352 | 704 | 1056 | 1408 | 1760 | 288 | 576 | 864 | 1152 | 1440 |
| BER-BIT ERROR RATE |  | $10^{-7}$ |  |  |  |  | $10^{-7}$ |  |  |  |  | $10^{-6}$ |  |  |  |  |
| $\mathrm{E}_{\mathrm{b}} / \mathrm{N}_{\mathrm{O}}$ ( dB , for specified BER with no error control, with optimum filtering. | BPSK | 11.3dB (CD); 11.9 dB (DD) |  |  |  |  | 11.3dB (CD); 11.9 dB (DD) |  |  |  |  | 10.5 dB (CD) ; 11.2 dB (DD) |  |  |  |  |
|  | $\left\|\begin{array}{l} \text { OPSKK } \\ \text { OQPSK } \end{array}\right\|$ | 11.3 dB (CD); 13.6 dB (DD) |  |  |  |  | 11.3 dB (CD); 13.6 dB (DD) |  |  |  |  | 10.5 dB (CD); 12.8 dB (DD) |  |  |  |  |
|  | MSK | 11.3 dB (CD) ; 11.9 dB (DD) |  |  |  |  | 11.3dB (CD); 11.9 dB (DD) |  |  |  |  | 10.5 dB (CD) ; 11.2 dB |  |  |  | (DD) |
| $\mathrm{C} / \mathrm{N}_{\mathrm{O}}$ ( dB , for $\mathrm{co-}$ herent detection of BPSK,(O) QPSK, MSK) |  | 67.1 | 70.1 | 71.9 | 73.2 | 74.1 | 66.7 | 69.8 | 71.5 | 72.8 | 73.8 | 65.1 | 68.1 | 69.9 | 71.1 | 72.1 |
| $C / \mathbb{N}(\mathrm{dB})$ for differential detection | $\begin{array}{\|l\|} \hline \text { BPSKK } \\ \text { DBPSK } \\ \hline \end{array}$ | 67.7 | 70.7 | 72.5 | 73.8 | 74.1 | 67.3 | 70.3 | 72.1 | 73.4 | 74.4 | 65.8 | 68.8 | 70.6 | 71.8 | 72.8 |
|  | DQPSK | 69.4 | 72.4 | 74.2 | 76.1 | 76.4 | 69.0 | 72.0 | 73.8 | 75.1 | 76.1 | 67.4 | 70.4 | 72.2 | 73.4 | 74.4 |
| $\mathrm{E}_{\mathrm{b}} / \mathrm{N}_{\mathrm{O}}$ loss due to frequency offset or delay error at tabulated BER values | DBPSK | $.3 \mathrm{~dB} \Delta \omega_{\mathrm{o}} T=.1$ |  |  |  |  | $.3 \mathrm{~dB} \Delta \omega_{\mathrm{o}} \mathrm{T}^{\text {P }}=.1$ |  |  |  |  | $.2 \mathrm{~dB} \Delta \omega_{\mathrm{o}}{ }^{T}=.1$ |  |  |  |  |
|  | DQPSK | $.5 \mathrm{~dB} \omega_{0} \Delta T=.04$ |  |  |  |  | $.5 \mathrm{~dB} \omega_{\mathrm{o}} \Delta \mathrm{T}=.04$ |  |  |  |  | $.4 \mathrm{~dB} \omega_{0} \Delta T=.04$ |  |  |  |  |
|  | DMSK | $.1 \mathrm{~dB} \Delta \omega_{\mathrm{o}} \mathrm{O}^{T}=.06$ |  |  |  |  | $.1 \mathrm{~dB} \Delta \omega_{\mathrm{o}} \mathrm{T}=.06$ |  |  |  |  | $.1 \mathrm{~dB} \Delta \omega_{0} T=.06$ |  |  |  |  |

Table 3.2: 1 to 5 Channels TDM Bit Stream Characteristics For Three Quality Levels ( 15 kHz channels) and For the Modulation Techniques Considered. Differential detection of MSK assumed to be ISI-free.
CD - Coherent Detection
DD - Differential Detection

Subsequent sections address the impact of phase noise, channel filter and amplifier (linear or nonlinear) characteristics, presence of adjacent signals and the effects of synchronizations circuits on ideal performance. The modulation section concludes with realistic bandwidth and power parameters for the candidate modulation techniques.
3.2.1 Specification of Cumulative Oscillator Phase Noise

In this section an approach to specifying cumulative oscillator phase noise for differential and coherent detection of BPSK, QPSK and. MSK is presented. It should be recognized that there exists a tradeoff between various receiver parameters (e.g. noise versus acquisition performance). Nevertheless, representative values are provided which apply to the three radio program quality levels.

While the derivation is general, results will be presented for single radio program situations because the SCPC cases represent the lowest data rates and are therefore most sensitive to phase noise.

### 3.2.1.1 Phase Noise in Differential Detection Receivers

Figure 3.1 depicts demodulator block diagrams for differential detection of BPSK, MSK and QPSK. In each case a bandpass filter followed by a hard limiter appears at the demodulator input. The hard limiter serves to convert additive Gaussian noise (at high SNR) to low-deviation phase modulation which is additive to the phase noise of the carrier. This allows the computation of the degradation due to phase noise relative to that associated with thermal noise only.

(a) DBPSK

(b) DMSK

(c) DQPSK

Figure 3.1: Differential Detectors for BPSK., MSK and QPSK

Case 1: DBPSK

The equivalent phase processing model is shown in Figure 3.2a [23]. The lowpass filter is assumed to be ideal with the cutoff frequency equal to half the noise bandwidth of the IF filter. Note that for BPSK and Nyquist filtering the. IF noise bandwidth is $R$ where $R$ is the data rate.

The phase passband is determined by the cascade of the lowpass filter and the delay line cancellor. The transfer function is then the product of the filter and cancellor transforms.

For the delay line cancellor, the impulse response is

$$
\begin{align*}
h_{c}(t) & =\delta(t)-\delta(t-T) \\
\therefore H_{C}(f) & =1-e^{-j 2 \pi f T} \tag{3.1}
\end{align*}
$$

The cascaded response is then

$$
\begin{align*}
H(f) & =K\left(1-e^{-j 2 \pi f T}\right) \\
|H(f)|^{2} & =2 K(1-\cos 2 \pi f T) \\
& =\sin ^{2} \pi f T \\
& =\left[\begin{array}{cl}
\sin ^{2} \pi f / R & \\
0 & \\
0 & \text { elsewhere }
\end{array}\right. \tag{3.2}
\end{align*}
$$

where $T=1 / R$ and the constant $K=\frac{1}{4}$. The magnitude of the transfer function is shown in Figure 3.3.

(a) DBPSK

(b) DMSK

(c) DQPSK

Figure 3.2: Equivalent phase processing models


Thermal and phase noise-to-carrier power ratios at the output of the cascaded network are then determined as follows

$$
\begin{aligned}
& P_{n}=K_{1} \int_{-\infty}^{\infty}|H(f)|^{2}\left(\frac{C}{N_{0}}\right)^{-1} d f \\
& P_{\phi}=2 K_{1} \int_{-\infty}^{\infty}|H(f)|^{2}\left(\frac{C}{\phi_{0}}\right)^{-1} d f
\end{aligned}
$$

where $C / N_{0}$ and $C / \phi_{o}$ are the input carrier-to-noise spectral density and carrier-to-phase noise spectral density ratios respectively. The factor of 2 in the expression for $P_{\phi}$ accounts for the fact that upper and lower sidebands due to pure phase modulation combine on a voltage basis whereas corresponding bands of Gaussian noise are added noncoherently on a power basis.

The relative degradation due to phase noise is then given by

$$
\begin{aligned}
L & =1+\frac{P_{\phi}}{P_{n}} \\
& =1+\frac{0}{\infty}|H(f)|^{2}\left(C / \phi_{O}\right)^{-1} d f \\
& \int_{0}^{\infty}|H(f)|^{2}\left(C / N_{O}\right)^{-1} d f
\end{aligned}
$$

Now

$$
\begin{align*}
\int_{0}^{\infty}|H(f)|^{2}\left(\frac{C}{N_{0}}\right)-1 d f & =\left(\frac{C}{N_{0}}\right)^{-1} \int_{0}^{R / 2} \sin ^{2} \frac{\pi f}{R} d f \\
& =\frac{R}{4\left(C / N_{0}\right)} \tag{3.4}
\end{align*}
$$

Figure 3.4 depicts phase spectral densities for a number of oscillators. A common characteristic shown is that the spectral density is essentially a decreasing function of frequency. The transfer function $|H(f)|^{2}$ is an increasing function of frequency to almost $f / R=0.5$. The product of the spectra can then be expected to give a maximum. It will be assumed here that this occurs at $f / R=0.3$ for which the slope of $|H(f)|^{2}$ is nearly $20 \mathrm{~dB} /$ decade. Note that $|H(f)|^{2} \simeq 0.65$ for $f / R=0.3$.

The numerator in the expression for $L$ is then approximated by

$$
\begin{equation*}
\int_{0}^{\infty}|H(f)|\left(\frac{C}{\phi_{0}}\right)^{-1} d f \simeq \frac{\left(\frac{R}{2}-f_{0}\right)\left|H\left(f_{m}\right)\right|^{2}}{\left(\frac{C}{\phi_{0}}\right)} \tag{3.5}
\end{equation*}
$$

where

$$
\begin{aligned}
f_{0}= & \text { the lowest frequency of significant weighted } \\
& \text { phase noise }
\end{aligned}
$$

$\left(C / \phi_{O}\right)_{m}=C / \phi_{O}$ at the frequency of maximum density of

$$
|H(f)|^{2} \quad\left(C / \phi_{O}\right)^{-1}
$$



FIGURE 3.4 PHASE-NOISE PLOTS FOR COMMERCIALLY AVAILABLE REFERENCE STANDARDS.
MEASURED DATA OR DATA SHEET TYPICALS FOR:
Curve 1: H.P. 5105 A 5110 Synthesizer at 400 MHz
Curve IV: Frequency-West Crysial Oscillator al 100 MHt
Curve II: Fluke. 6160 A at 160 MHz
Curve III: Frequency-West VCXO at 100 MHz
Curve V; H.P. 105 A at 5 MHz

$$
f_{m}=\text { frequency at maximum density of }|H(f)|^{2}\left(\frac{C}{\phi_{O}}\right)^{-1}
$$

Letting $f_{o}=0$,

$$
\begin{equation*}
\int_{0}^{\infty}|H(f)|^{2}\left(\frac{C}{\phi_{O}}\right)^{-1} d f \simeq \frac{.65 R}{2\left(\frac{C}{\phi_{O}}\right) \cdot 3 R} \tag{3.6}
\end{equation*}
$$

Thus,

$$
\begin{align*}
L & =I+\frac{2\left(\frac{.65 \mathrm{R}}{2}\right)\left(\mathrm{C} / \mathrm{N}_{\mathrm{O}}\right)}{\left(\frac{\mathrm{R}}{4}\right)\left(\mathrm{C} / \phi_{O}\right) \cdot 3 \mathrm{R}} \\
& =I+\frac{2.6\left(\mathrm{C} / \mathrm{N}_{\mathrm{O}}\right)}{\left(\mathrm{C} / \phi_{\mathrm{O}}\right)_{.3 R}} \tag{3.7}
\end{align*}
$$

Allowing . 2 dB degradation due to phase noise gives

$$
L=1.05
$$

for which

$$
\begin{align*}
\left(\frac{C}{\phi_{0}}\right)^{2 R} & =52\left(\frac{C}{N_{0}}\right), \quad \text { absolute units } \\
& =\frac{C}{N_{0}}+17.2, d B \tag{3.8}
\end{align*}
$$

Example

$$
\begin{aligned}
& \text { BER required } \\
& \begin{array}{ll}
\mathrm{R}(\mathrm{kbit} / \mathrm{s}) & : 10^{-6} \\
\text { Modulation } & : 288 \\
\mathrm{E}_{\mathrm{b}} / \mathrm{N}_{\mathrm{O}} \text { for } 0.2 \mathrm{~dB} \text { degradation: } 11.4 \mathrm{~dB} \\
& \begin{aligned}
\therefore\left(\frac{\mathrm{C}}{\phi_{O}}\right) .3 \mathrm{R} & =\frac{C}{N_{O}}+17.2, \mathrm{~dB} \\
& =11.4+10 \log _{10}\left(2.88 \mathrm{x} 10^{3}\right)+17.2
\end{aligned} \\
& =83.2 \mathrm{~dB}-\mathrm{Hz}
\end{array}
\end{aligned}
$$

The specification frequency is

$$
\begin{aligned}
.3 \mathrm{R} & =.3\left(288 \times 10^{3}\right) \\
& =86.4 \mathrm{kHz}
\end{aligned}
$$

## Case 2: DMSK

The equivalent phase processing model is shown in Figure 3.2b. There are two aspects that differentiate this model from the one for DBPSK:
(1) detection is based on sampling the sine rather than the cosine of the phase difference output
(2) the transmission bandwidths of MSK and BPSK are different

Point (l) has no impact on the analysis. With respect to the impact of bandwidth what is pertinent is the bandwidth
of the receive filter. As previously mentioned the lowpass equivalent noise bandwidth for Nyquist-shaped BPSK is R/2. It has been determined in [13] that a 4-pole Butterworth filter having an IF 3 dB bandwidth of 1.1 R should be used for DMSK. For a 4-pole Butterworth filter the noise bandwidth is 1.03 x ( 3 dB bandwidth). For the purposes of this analysis we shall assume that the lowpass noise bandwidth required for DMSK is essentially the same as that for DBPSK. Thus, the results for DBPSK also apply to DMSK.

Case 3: DQPSK

The equivalent phase processing model for DQPSK is shown in Figure 3.2c. Note that for the same serial data rate, QPSK will require half the bandwidth of $B P S K$ and the delay in the differential detector will be twice as long because the symbol duration is twice as long.

The analysis presented for DBPSK thus applies except $R$ and $T$ are now replaced by $R_{S}=R / 2$ and $T_{S}=2 T$.

It follows that

$$
\begin{equation*}
L_{\text {DQPSK }}=1+\frac{2.6\left(\mathrm{C} / \mathrm{N}_{\mathrm{O}}\right)}{\left(\mathrm{C} / \phi_{\mathrm{O}}\right) \cdot 3 \mathrm{R}_{\mathrm{S}}} \tag{3.9}
\end{equation*}
$$

and for 0.2 aB degradation

$$
\begin{equation*}
\left(\frac{\mathrm{C}}{\phi_{0}}\right) \cdot 3 \mathrm{R}_{\mathrm{s}}=\frac{\mathrm{C}}{\mathrm{~N}_{\mathrm{O}}}+17.2 \mathrm{~dB} \tag{3.10}
\end{equation*}
$$

3.2.1.2 Phase Noise in Coherent Detection Receivers

Case 1: Coherent PSK

For the PSK signals the carrier phase is an integer multiple of $2 \pi / N$ where $N=2,4$ for $B P S K$ and QPSK respectively. The simplest class of suppressed carrier tracking loop is the Nth power loop shown in Figure 3.5.

The noisy modulated input signal is filtered and then passed through a (xN) multiplier which removes the phase modulation and produces a discrete component at $N f_{0}\left(f_{0}=\right.$ input centre frequency). The phase locked loop (PLL) tracks the signal component at $\mathrm{Nf}_{\mathrm{O}}$ and the reconstructed carrier references are obtained from the VCO output at $\mathrm{f}_{\mathrm{o}}$.

It is shown in [12] that the variance ( $\sigma_{\phi}^{2}$ ) of the. phase jitter associated with the noisy reference at $f_{0}$ is related to that associated with the reference ( $\sigma^{2}{ }_{N \phi}$ ) at Nfoby

$$
\begin{equation*}
\sigma_{\phi}^{2}=\frac{1}{N^{2}} \sigma_{N \phi}^{2}=\frac{S_{L^{-1}}}{\rho} \tag{3.11}
\end{equation*}
$$

where

$$
\begin{aligned}
\rho= & \text { equivalent signal-to-noise ratio in the loop } \\
& \text { bandwidth } B_{L} \text { of a second-order PLL (assumed for } \\
& \text { this case) }
\end{aligned}
$$

The intermodulation loss depends on $N$, the SNR in the tracking loop input filter bandwidth and the type of nonlinearity. The following assumptions apply to the tracking loop SNR:


Figure 3.5:
The Nth power loop

- the PLL operates in its linear region
- the I- and Q- data streams (for QPSK) and the I data stream for BPSK are binary, zero mean and independent
- the signalling pulses all have the same waveshape and the input filter is symmetrical about the carrier frequency
- the input noise is Gaussian
$S_{L}{ }^{-1}$ is greater than 1 and therefore the loop SNR is degraded, leading to greater phase jitter than for a PLL not preceded by a nonlinearity.

Figure 3.6 shows the relationship between $S_{L^{-1}}$ and $\rho \gamma=2 \rho_{i}$ where $\gamma$ is the ratio of the two-sided loop bandwidth to that of the input bandpass filter bandwidth and $\rho_{i}$ is the SNR in the tracking loop input bandwidth. Note that for an ideal input bandpass filter a (x4) loop (for QPSK) exhibits greater loss than a doubler (for BPSK).

This presentation has so far accounted for the effects of additive noise only. If phase noise is present, the $N$-fold nonlinearity will multiply the phase by $N$ so that the phase noise variance presented to the PLL will be increased by $\mathrm{N}^{2}$. However, the PLL output is effectively divided by $N$ so that the loop phase noise variance is reduced by $\mathbb{N}^{2}$.

The total mean square phase error of the tracking loop output is then given by [24]

$$
\sigma_{T}{ }^{2}=\sigma_{\phi}{ }^{2}+\sigma_{P}{ }^{2}
$$



Figure 3.6: Reciprocal of the $N$-phase loss as a function of $N$ and input signal-tonoise ratio for $R C$ and ideal band-pass filters


Figure 3.7: Error response of high-gain loop, $\zeta=0.707$.

$$
\begin{equation*}
=\frac{S_{L}^{-1}}{\frac{C}{N_{O} B_{L}}}+2 \int_{0}^{\infty}|1-H(f)|^{2} S_{P}(f) d f \tag{3.12}
\end{equation*}
$$

where $\sigma_{P}{ }^{2}$ is the contribution due to phase noise, $H(f)$ is the open loop transfer function of the $P L L$ and $S_{P}(f)$ is the phase noise spectral density measured in $\mathrm{rad}^{2} / \mathrm{Hz}$.

Figure 3.7 depicts $|i-H(\omega)|^{2}$ as a function of $\omega / \omega_{n}$ where for a high-gain second-order loop

$$
\begin{aligned}
B_{L} & =\frac{\omega_{n}}{2}\left(\zeta+\frac{1}{4 \zeta}\right) \\
& =.53 \omega_{n}
\end{aligned}
$$

with $\zeta=.707$
Note that the transfer function $|1-H(\omega)|^{2}$ is highpass. If the signal at the input to the tracking loop is filtered by an ideal filter that limits the input bandwidth then the upper limit in the integration in (3.12) is finite. The lower limit is effectively determined by $B_{L}$. The larger the value of $B_{L}$ the less will be $\sigma_{P}{ }^{2}$ for a given $S_{P}(f)$. However, increasing $B_{L}$ increases $\sigma_{P}{ }^{2}$. Thus there will exist an optimum value of $B_{L}$ which minimizes $\sigma_{T}{ }^{2}$. While $B_{L}$ will determine the output phase error characteristic its selection may be affected by the required acquisition performance of the PLL. In what follows only the phase error aspect will be considered.

Now as in the differential detection case

$$
\begin{align*}
\sigma_{P}^{2} & =4 \int_{0}^{\infty}|1-H(f)|^{2}\left(\frac{C}{\phi_{O}}\right)-1 d f \\
& \simeq \frac{4\left(R_{S}-f_{O}\right)|1-H(f m)|^{2}}{\left(C / \phi_{O}\right)_{m}}
\end{align*}
$$

where $R_{S}$ is the symbol rate. For BASK $R_{s}=R$ for QPSK $R_{s}=$ R/2.

Note that the slope of $|1-H(f)|^{2}$ is greater than that of $|H(f)|^{2}$ for the differential detection case. In any event there will be a frequency at which the weighted phase noise will be a maximum. For the purposes of this discussion assume that this occurs at $\omega / \omega_{n}=1$ or $\omega=1.9 \mathrm{~B}_{\mathrm{L}} \mathrm{rad} / \mathrm{sec}$ at which frequency $|1-H(f)|^{2}=0.5$. Also assume $f_{0}=0$.

$$
\begin{equation*}
\therefore \sigma_{P}^{2} \simeq \frac{2 R_{S}}{\left(\frac{\mathrm{C}}{\phi_{O}}\right)_{1.9 B_{L}}} \tag{3.14}
\end{equation*}
$$

Assume that the $N$-ary PSK system is to operate at a nominal $\mathrm{E}_{\mathrm{b}} / \mathrm{N}_{0}=r_{0}$. The SNR at the input to the tracking loop* is

$$
\rho_{i}=\frac{r_{0} R}{W_{i}}=\frac{C}{N_{0} W_{i}}=\frac{C}{N_{0}\left(2 R_{s}\right)}=\frac{r_{0} R}{2 R_{s}}
$$

For the output phase jitter due to noise
*The IF bandwidth of $2 \mathrm{R}_{\mathrm{S}}$ is required so that the effects of modulation envelope ripple may be ignored.

$$
\begin{equation*}
\sigma_{\phi}^{2}=\frac{S_{L}^{-1}}{\rho}=\frac{S_{L}^{-1}}{\frac{C}{N_{o} B_{L}}}=\frac{S_{L^{-1}}^{r_{O}}}{\frac{r_{0}}{B_{L}}}=\frac{B_{L}}{R}\left[\frac{S_{L}-1}{r_{O}}\right] \tag{3.15}
\end{equation*}
$$

Figure 3.8 [25, 26] provides the phase reference SNR required by PSK and MSK systems as a function of tolerated detection loss*. For a desired error rate and detection loss these graphs provide the required phase reference SNR which in turn is inversely proportional to the phase jitter variance.

## Example:

| Radio Program Quality Level | Low |  |
| :--- | :--- | :--- |
| BER required | $: 10^{-6}$ |  |
| $R(k b i t / s)$ | $:$ | 288 |
| Modulation | $:$ | BPSK |
| $\mathrm{E}_{\mathrm{b}} / \mathrm{N}_{\mathrm{o}}$ for .2 dB detection loss: | 10.7 dB |  |

Figure 3.8 shows that the $\operatorname{loop} \operatorname{SNR}\left(A=\log _{10} \alpha=\log _{10} \rho\right)$ should be 15.4 dB .

$$
\therefore \sigma_{\mathrm{T}}^{2}=\frac{1}{10^{1.54}}=2.88 \times 10^{-2} \quad \mathrm{rad}^{2}
$$

Suppose that the jitter is divided equally between phase. noise associated with oscillators and that due to input noise

$$
\begin{align*}
& \therefore \sigma_{\phi}^{2}=1.44 \times 10^{-2}  \tag{3.16}\\
& \therefore B_{L}=\frac{R r_{O} \sigma_{\phi}^{2}}{S_{L}^{-1}} \tag{3.17}
\end{align*}
$$

* Note that the PSK results assume non-return-to-zero (NRZ) baseband pulses.


Figure 3.8 Phase Reference SNR as Function of Error Rate for Given Detection Loss $L$ (dB). (a) $L=.2 \mathrm{~dB}$. (b) $L=.1 \mathrm{~dB}$. (c) $L=$ .05 dB .

For BPSK

$$
\begin{equation*}
\rho_{i}=\frac{r_{o}}{2}=\frac{10^{1.07}}{2}=5.9 \tag{3.18}
\end{equation*}
$$

Extrapolating the results in Figure 3.6 shows that $S_{L^{-1}} \simeq$ 1.l for an ideal BPF at the tracking loop input

$$
\begin{align*}
\therefore B_{\mathrm{L}} & =\frac{(288 \mathrm{kbps})\left(10^{1.07}\right)\left(1.44 \times 10^{-2}\right)}{1.1} \\
& =44,296 \mathrm{~Hz} \tag{3.19}
\end{align*}
$$

From (3.14)

$$
\begin{align*}
\left(\frac{\mathrm{C}}{\phi_{O}}\right)_{1.9 B_{L}} & =\frac{2 R_{S}}{\sigma_{P}{ }^{2}} \\
& =\frac{2 \times 288 \times 10^{3}}{\left(1.44 \times 10^{-2}\right)} \\
& =76 \mathrm{~dB} \cdot \mathrm{~Hz} \tag{3.20}
\end{align*}
$$

## Case 2: Coherent MSK

The regeneration of a local carrier for MSK is accomplished with the circuit of Figure 3.9. The MSK signal with frequency deviation $h=0.5$ (i.e. signalling frequencies separated by half the data rate) is passed through a frequency doubler producing Sunde's FSK with h = 0.1 [25]. Whereas the MSK spectrum is continous, the Sunde FSK contains both continuous and line spectral components. Two phase locked loops track the line components. The PLL outputs are combined and the sum frequency divided by 4 to provide a local carrier.


Figure 3.9: Carrier Regeneration for MSK

It is shown in [27] that the SNR in each PLL bandwidth is given by

$$
\begin{equation*}
\rho=\frac{1}{T B_{L}\left[1+\frac{15.9}{E_{b} / N_{o}}+\frac{.636}{\left(E_{b} / N_{o}\right)^{2}}\right]} \tag{3.21}
\end{equation*}
$$

The phase jitter on the output carrier due to the noise will then be given

$$
\begin{align*}
\sigma_{\phi}^{2} & =\frac{1}{16}\left[\frac{2}{\rho}\right] \\
& =\frac{1}{8 \rho} \tag{3.22}
\end{align*}
$$

With respect to oscillator phase noise the doubling, PLL output multiplication, division by 4 operations produce a result equivalent to filtering the input phase noise spectrum by the PLL inverse transfer characteristic* i.e.

$$
\begin{align*}
\sigma_{P}^{2} & =4 \int_{0}^{\infty}|1-H(f)|^{2}\left(\frac{C}{\phi_{O}}\right)^{-1} d f \\
& \simeq \frac{4 \mathrm{R}}{2\left(\mathrm{C} / \phi_{O}\right)_{\mathrm{m}}} \\
& =\frac{2 \mathrm{R}}{\left(\mathrm{C} / \phi_{O}\right)^{1} 1 \cdot 9 \mathrm{~B}_{\mathrm{L}}}
\end{align*}
$$

Example:

| Radio Program Quality Level | Low |
| :--- | :--- |
| BER required | $: 10^{-6}$ |
| R(kbit/s) | $: 188$ |
| Modulation | : MSK |
| $E_{b} / N_{o}$ for .2 dB detection loss: | 10.7 dB |

Figure (3.8) shows that the loop SNR should be 21.9 dB

$$
\therefore \sigma_{T}^{2}=\frac{1}{10^{2} \cdot 19}=6.46 \times 10^{-3} \mathrm{rad}^{2}
$$

Suppose that the jitter is divided equally between phase noise associated with oscillators and that due to input thermal noise

$$
\begin{align*}
\therefore \sigma_{\phi}^{2} & =3.23 \times 10^{-3} \mathrm{rad}^{2} \\
& =\frac{1}{8 \rho} \tag{3.24}
\end{align*}
$$

$$
\begin{aligned}
\therefore B_{L} & =\frac{8 \sigma_{\phi}{ }^{2}}{T\left[1+\frac{15.9}{\left(E_{b} / N_{o}\right)}+\frac{.636}{\left(E_{b} / N_{o}\right)^{2}}\right]} \\
& \simeq \frac{8 \sigma_{\phi}{ }^{2} \mathrm{R}}{1+\frac{15.9}{E_{b} / N_{o}}} \\
& =\frac{8 \times\left(3.23 \times 10^{-3}\right)\left(288 \times 10^{3}\right)}{1+\frac{15.9}{10^{1.07}}}
\end{aligned}
$$

$$
\begin{equation*}
=3163 \mathrm{~Hz} \tag{3.25}
\end{equation*}
$$

$$
\begin{align*}
\left(\frac{\mathrm{C}}{\phi_{O}}\right)_{1.9 \mathrm{~B}_{\mathrm{L}}} & =\frac{2 \mathrm{R}}{\sigma_{\mathrm{P}}^{2}} \\
& =\frac{2 \times 288 \times 10^{3}}{3.23 \times 10^{-3}} \\
& =82.5 \mathrm{~dB} \cdot \mathrm{~Hz} \tag{3.26}
\end{align*}
$$

3.2.1.3 Phase Noise Results

Table 3.3 provides phase noise specifications for differential and coherent detection of BPSK, QPSK and MSK. The results apply to single radio program per digital carrier. For differential detection the $E_{b} / N_{o}$ degradation due to oscillator phase noise is taken to be 0.2 dB . For coherent detection the locally derived phase reference will suffer jitter due to thermal noise as well as oscillator phase noise. In these cases the degradation is taken to be equal. with a combined degradation of 0.2 dB .

For the three quality levels, the least stringent phase noise specification is for coherent BPSK while the most stringent is coherent QPSK. Coherent BPSK does not suffer from I-/Q- channel crosstalk as do coherent QPSK and MSK and therefore is less sensitive to phase noise.

Examination of the oscillator phase noise characteristics given in Figure 3.4 show that all easily meet the differential detection requirement. The coherent detection requirements are also met. However, coherent MSK and QPSK require substantially better oscillators.

Low Quality radio program distribution places the most stringent phase noise specification primarily because of its

|  | Quality Level |  |  |  |  |  |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: |
|  | CATV |  | High Quality |  | Low Quality |  |
| Modulation | $\begin{gathered} \frac{\mathrm{C}}{\phi_{O}} \\ (\mathrm{~dB} \cdot \mathrm{~Hz}) \end{gathered}$ | $\Delta \mathrm{f}$ <br> ( kHz ) | $\begin{gathered} \frac{C}{\phi_{O}} \\ (\mathrm{~dB} \cdot \mathrm{~Hz}) \end{gathered}$ | $\Delta \mathrm{f}$ (kHz) | $\begin{gathered} \frac{\mathrm{C}}{\phi_{O}} \\ (\mathrm{~dB} \cdot \mathrm{~Hz}) \end{gathered}$ | $\Delta \mathrm{f}$ $(\mathrm{kHz})$ |
| DIFFERENTIAL DETECTION: |  |  |  |  |  |  |
|  |  |  |  |  |  |  |
| DBPSK | 85.1 | 115.2 | 84.8 | 105.6 | 83.2 | 86.4 |
| DQPSK | 86.8 | 57.6 | 86.5 | 52.8 | 84.8 | 43.2 |
| DMSK | 85.1 | 115.2 | 84.8 | 105.6 | 83.2 | 86.4 |
| COHERENT DETECTION: CBPSK |  |  |  |  |  |  |
|  |  |  |  |  |  |  |
|  | 77.4 | 129.2 | 77.1 | 118.5 | 76 | 84.1 |
| CQPSK | 87.0 | 3.6 | 86.6 | 3.3 | 85.1 | 2.6 |
| CMSK | 84.4 | 7.7 | 84.0 | 7.0 | 82.5 | 6.0 |

Table 3.3: ( $C / \phi_{o}$ ) required for 0.2 dB degradation with differential detection and coherent detection
lower data rate. For all the differential schemes as well as coherent BPSK there is at least 20 dB margin between the required $C / \phi_{o}$ and the measured performance of the poorest oscillator in Figure 3.4. Using these modulations would then allow for minimizing the oscillator cost component of the radio program system. This cost savings would have to be compared with the transponder utilization efficiencies in order to arrive at a cost-effective selection of a modulation technique.
3.2.2 Sensitivity to Frequency and Delay Errors In Differential Detection Systems

Coherent detection (CD) requires frequency lock and phase coherence. Differential detection (DD) is possible with an IF frequency offset ( $\left.\Delta \omega_{o} r a d / s\right)$ but with an $E_{b} / N_{o}$ penalty. This penalty arises from $\Delta \omega_{o} T \neq 2 k \pi$. In bandlimited systems, additional degradation is expected due to IF frequency offset since the receive IF filter characteristic is assymetrical with respect to the received signal IF frequency.

Figure 3.10 [4] shows the bit error probability for differentially coherent detection of a non-bandlimited DBPSK signal for various values of frequency offset normalized to the bit duration ( $\Delta \omega_{0} T$ ). As can be seen the normalized frequency offset should be $\leqslant .2$ if small degradation in performance is to be obtained. The DBPSK degradation in in $E_{b} / N_{o}$ with $\Delta \omega_{o} T=.1\left(5.7^{\circ}\right)$ is indicated in Table 3.2 for the tabulated BER's (At $P_{e}=10^{-4}$ the degradation for $\Delta \omega_{O} T=.1$ is about 0.2 dB$)$.


Fig 3.10 Bit error probability for differentially coherent detection of PSK(DCPSK) for various values of frequency offset $\Delta \omega_{0} T$ normalized to bit duration $T \mathrm{sec}$ !

A study performed by FACC [29] provides simulation results for the degradation in performance of a DQPSK system as a function of the delay line error ( $\Delta T$ )*. Table 3.4 summarizes some of the main results. It may be concluded that with differential demodulation of OPSK the phase error has to be less than $2.5^{\circ}$ (. 04 rad ) in order to limit the degradation in $E_{b} / N_{o}$ to .2 dB at $P_{e}=10^{-4}$. The most obvious source of phase error is the delay line. If the delay changes by $\Delta T$ then the phase changes by $\omega_{0} \Delta T$ radians where $\omega_{o}=2 \pi f_{0}$ and $f_{o}$ is the carrier frequency. Table 3.2 shows the estimated delay or frequency error degradation for DOPSK for the tabulated BER values, by linearly extrapolating the results of Table 3.4.

A study performed by MCS [13] provides some estimates of DSMK degradation from CMSK performance with carrier frequency offset. It is indicated that for frequency offset of $\Delta \omega_{o} T=.06\left(3.4^{\circ}\right)$ the degradation is on the order of 0.1 dB at $\mathrm{P}_{\mathrm{e}}=10^{-4}$.

Based on the comparison at $P_{e}=10^{-4}$ it seems that the phase and delay control requirements of DMSK and DBPSK are similar. Somewhat tighter control is required with DQPSK.

It may be concluded that in all differentially coherent systems a tight control of transmit frequency and/or

[^3]| CAUSES OF DEGRADATION | AMOUNT OF DEGRADATION |  |
| :---: | :---: | :---: |
|  | $\mathrm{P}_{\mathrm{e}}=10^{-4}$ | $\mathrm{P}_{\mathrm{e}}=10^{-5}$ |
| Intersymbol Interference Only | 0.4 | 0.6 |
| Intersymbol Interference plus <br> Phase Shift Error (2.5 rms) | 0.6 | 0.9 |
| Intersymbol Interference plus <br> Symbol Timing Jitter <br> (0.015 T rms) | 0.5 | 0.8 |
| Intersymbol Interference plus <br> Symbol Timing Jitter plus <br> Phase Shift Error | 0.7 |  |

Table 3.4: Degradations in DQPSK System
receiver delay should be maintained in order to keep the associated degradation lower than a few tenths of a dB. This may require AFC circuits.

Performance Over a Satellite Channel Without Adjacent Channel Interference

A comparative evaluation of all the modulation and detection schemes for a satellite channel but excluding DMSK was performed in [30]. It should be noted that there was no attempt in this study to optimize performance with respect to the type of filtering used. Rather, filters were selected somewhat arbitrarily and the optimization was performed with respect to the bandwidth $x$ bit duration (BT) product. In fact, with respect to PSK systems the results would appear to be pessimistic because Nyquist filtering was not used to reduce intersymbol interference (ISI). Nevertheless, the results represent one of the few comprehensive comparisons available.

In [30] the simulated channel consists of identical transmitter and receiver filters. A 4-pole $0.5-\mathrm{dB}$ ripple Chebyshev characteristic was used in each case. Nyquist filtering was not used. The satellite effects were represented by an INTELSAT IV TWTA model. Computer simulation results were generated as a function of the IF $3-\mathrm{dB}$ bandwidth-bit period product ( $B T$ ) for satellite input back-off values of 12 dB (linear region) and 1 dB (nonlinear region). Table 3.5 presents the $B T$ required for each case to yield an $E_{b} / N_{o}$ degradation of no more than 1 dB at $\mathrm{BER}=10^{-5}$.

| $B P S K$ | QPSK | $0 Q P S K$ | $M S K(F F S K)$ |
| :---: | :---: | :---: | :---: |
| $(B T)_{\ell}=1.9(C D, D D)$ | $(B T)_{\ell}=1.0(C D) ; 1.3(D D)$ | $(B T)_{\ell}=1.5(C D)$ | $(B T)_{\ell}=1.2(C D)$ |
| $(B T)_{n \ell}=2.1(C D, D D)$ | $(B T)_{n \ell}=1.0(C D) ; 2.1(D D)$ | $(B T)_{n \ell}=1.4(C D)$ | $(B T)_{n \ell}=1.1(C D)$ |

3.2.4 Performance Over a Satellite Channel With Adjacent Channel Interference

An indication of the sensitivity to adjacent channel interference (ACI) is presented in the following. Note that interferers are of the same modulation type.

The performance of MSK and OQPSK systems in the presence of ACI is studied in [31] for a bandpass, hardlimited channel. Table 3.6 shows the performance degradation as a function of the normalized channel spacing for $P_{e}=10^{-4}$ and $S / A C I=25 d B$ (with a single adjacent channel of the same type) assuming a 7 -pole Chebyshev receiver filter. It is observed that an MSK and OQPSK degradation crossover occurs at a channel spacing of $1.1 R(R=1 / T=$ bit rate), moderate signal bandwidth restrictions (1.05R for MSK and 1.1R for OQPSK). It is also noted that for 1.4 R channel spacing MSK out-performs OQPSK.

A simulation study [32] of modulation schemes proposed for nonlinear satellite channels compares the performance of MSK, QPSK and OQPSK over wideband and narrowband channels, where the channel spacing is 1.15 and .65 times the symbol rate ( $1 / T$ ) respectively. Performance results were obtained for single carriers as well as for carriers with upper and lower adjacent channels. In the study, the channel filter is assumed to be a $15 \%$ roll-off raised cosine filter. In

| Channel <br> Frequency <br> Spacing | Classification <br> of <br> Bandwidth <br> Restrictions | Required Bandwidth* for $25 d B$ S/ACI |  | $E_{b} / N_{o}$ Performance Degradation (dB) at $10^{-4}$ |  |
| :---: | :---: | :---: | :---: | :---: | :---: |
|  |  | MSK | O-QPSK | MSK | O-QPSK |
| 1. 4 R | Very Slight | 1.45R ${ }^{\text {a }}$ | 1.30R | 0.15 | 0.50 |
| 1. 2 R | Slight | 1.15R | 1.15R | 0.50 | 0.65 |
| 1.1R | Moderate | 1.05R | 1.10R | 0.75 | 0.75 |
| 1.0R | Tight ${ }^{\text {b }}$ | 0.93 R | 1.02R | 1.20 | 0.85 |
| 0.8 R | Very Tight | 0.82 R | 0.90 R | 2.40 | 1.10 |

a Note that the filter bandwidth of $B=1.45 \mathrm{R}$ for each of two adjacent channels with a frequency separation of $1.4 R$ between band centres results in some overlap in passbands for adjacent channels.
b Tight filtering for sharp-cutoff case without delay equalization. With equalization, a bandwidth of 0.625 R and comparable spacing is allowable for offset QPSK.
*3 dB bandwidth

Table 3.6: Performance Degradation as a Function of Normalized Channel Spacing For $10^{-4} \mathrm{BER}$ and 25 dB S/ACI
the simulation process, hard-wired carrier and bit timing clock are used. Results indicate that, for all modulation methods, the best filtering distribution had a sharp cutoff filter at the transmit side and an equalized Nyquist shaping filter at the receive side. For nonlinearities, typical measured $A M / A M$ and $A M / P M$ data for an HPA and a TWTA are used. Figures 3.11a, b, c [32]. show some of the simulation results. (In the figures the notation (14/4) refers to 14 dB earth station HPA input backoff and 4 dB satellite TWTA input backoff, respectively).

With optimized transmit and receive filters, OQPSK and MSK show similar BER performances. . In the wideband model QPSK is inferior to the others by an amount of 0.5 dB in $\mathrm{E}_{\mathrm{b}} / \mathrm{N}_{0}$ at BER of $10^{-4}$ at $(6 / 4)$ dB back-off. However QPSK is superior to the others in the narrowband model.

It is also shown in this paper that OQPSK inherently has a large carrier jitter even when noise and ISI are absent (i.e. pattern jitter is significant); therefore, carrier recovery in OQPSK systems requires longer filtering time constants (to obtain the same carrier phase jitter) than in a QPSK system.

The conclusions of this study were:
(i) QPSK is the most preferable modulation technique for tightly packed nonlinear satellite channels,

MSK and OQPSK are preferred for wideband nonlinear channels.


Figure 3.11a. Equivalent power loss for various input back-off values for HPA and TWTA (BER: $10^{-4}$ )


TWTA input attenuation of center channel (dB)
Figure 3.11b Equivalent power loss due to TWTA input attenuation of center channel. (BER: $10^{-4}$ )


Carrier phase error (deg)
Figure 3.1 c
Equivalent power loss due to carrier phase error. (BER: $10^{-4}$ )

A comparative study of modem techniques for nonlinear channels was conducted by COMSAT Laboratories [33]. The study addressed the sources of performance degradation in high speed 4-phase PSK systems operating over band-limited nonlinear channels. Optimization of modem design included the following strategies:
(i) Selection of an optimum modulation format from among QPSK, OQPSK and MSK.
(ii) Selection of a channel filter for the optimum modulation format.
(iii) Sensitivity of synchronization loops.
(iv) Optimization of the power bandwidth tradeoff and operating points of the nonlinear elements for a given satellite link.
(v) Feasibility of adaptive equalization in a nonlinear environment.

The conclusions of this study were as follows:
(a) QPSK is the optimum modulation format for bandrestricted channels. For relaxed bandwidth channels in which adjacent channel interference is of primary importance, the OQPSK or MSK formats may be more appropriate.
(b) For mild nonlinearities in tandem, a Nyquist transmit filter with 45 -percent roll-off and a maximally flat receive filter with $\mathrm{BT}_{\mathrm{s}}=1.1$ offer the optimum channel performance. For strong nonlinearities in tandem, the reciprocal filter combination is a better solution.
(c) Synchronization error in a nonlinear environment is a critical issue and needs careful circuit design for optimum modem performance.
(d). For a band restricted channel, the optimum HPA and satellite IWTA operating points are at about 6- and 4- dB input backoff, respectively.*
(e) Adaptive equalization may improve the channel performance when the nonlinear effect is relatively mild and the channel is heavily distorted as indicated by Figure 3.12.

Extensive computer aided QPSK modem performance analysis over a linear channel is included in [34]. In this report, a degradation allocation analysis is presented for 3 types of channel filters, and with two identical adjacent interfering channels (each with 6 dB more power than in the main channel and with channel spacing equal to the bit rate) at $\mathrm{P}_{\mathrm{e}}=10^{-7}$. Table 3.7 summarizes some of the results obtained in [34]. It is observed that with a

[^4]

Figure 3.12(a): : Linear Channel Performance with 5-Tap Adaptive Equalization


Tigure 3.1.2(b) Nonlinear Charnel Performance with S-Tap Adantive Equalization

|  | $\begin{aligned} & \mathrm{E}_{\mathrm{b}} / \mathrm{N} \text { O Degr } \\ & 10^{-7} \text { For } \\ & \text { Eilters } \end{aligned}$ | radation At The Following | $3 E R=$ <br> g IF Rx |  |
| :---: | :---: | :---: | :---: | :---: |
| Source of Degradation | Square Root <br> Raised <br> Cosine $\alpha=0.8$ | 3rd-Order Butterworth Filter | $\begin{aligned} & \text { 4th-Order } \\ & \text { TBT } \\ & \text { Filter } \end{aligned}$ | Remarks |
| Filter Distortion | . 38 (.47) | . 58 (.73) | . 40 (.56) | Due to filter mismatch, BT\#l.0 and parabolic group delay $D_{3 d B}=.1 \pi$ |
| Carrier Frequency Offset | . 16 (.09) | . 15 (.05) | . 15 (.08) | $\Delta f=-8.3 \%$ of the $R x$ dB bandwidth |
| Carrier Phase Error | . 29 (.29) | . 31 (.31) | . 29 (.29) | $\varepsilon=3^{\circ}$ |
| Timing Error | . 33 (.03) | . 33 (.02) | .31 (.03) | $\delta=5 \% \mathrm{~T}$; timing error is timing offset from mid-bit instant |
| Adjacent Channel Interference | $0 . \quad(0 .)$ | .14 (.72) | . 05 (.42) | Approximation.Interference power is converted to equivalent thermal noise |
| Unailocated Sources | 0.5 (0.5) | 0.5 (0.5) | 0.5 (0.5) | Carrier phase jitter temperature variation, aging effects etc. |
| TOTAL | 1.66 (1.38) | 2.01 (2.33) | 1.70 (1.88) | The 4 th order TBT filter meets NATO Specifications |

Table 3.7
$E_{b} / N_{o}$ degradation allocation for $B T=0.9$ ( $B T=1.2$ ). Tx filter is square-root raised-cosine with $\alpha=0.8$.

4th order Transitional Butterworth Thompson (TBT) IF filter, a total $E_{b} / N_{o}$ degradation of less than 2 dB is expected over a $\mathrm{BT}_{s}$ (3 dB bandwidth-symbol period product) range of . 9 to 1.2 , allowing for $8.3 \%$ carrier offset, $3^{\circ}$ carrier reference phase error, 5\% timing error and ACI as indicated above.

Table 3.8 [34] shows the signal to adjacent channel interference ( $A C I$ ) ratio ( $A B$ ) for different $T x$ (squareroot raised cosine) filters and Rx filters. Based on this table, it is possible to estimate the loss in $E_{b} / N_{o}{ }^{\prime}$ assuming that the ACI power adds to the channel noise power as given by the following expression:

$$
\begin{equation*}
\left(\frac{E_{b}}{N_{0}}\right)_{A C I}=\left[\left(\frac{E_{b}}{\mathbb{N}_{0}}\right)^{-1}+\left(\frac{S}{A C I}\right)^{-1}\right]^{-1} \tag{3.27}
\end{equation*}
$$

where

$$
\begin{aligned}
& \left(\frac{E_{b}}{N_{O}}\right)_{A C I} \triangleq \text { the required } E_{b} / N_{o} \text { to achieve a desired } \\
& \text { BER in the presence of ACI. } \\
& \frac{S}{A C I} \quad=\text { signal to ACI power ratio for given } T x \\
& \text { filter, Rx filter and } B T \text { (3 } \mathrm{dB} \text { channel } \\
& \text { bandwidth } \mathrm{x} \text { symbol period) product. }
\end{aligned}
$$



Table 3.8: Signal-to-adjacent channel interference ratio (dB) for different Tx and Rx filters.' The two adjacent channel signals each have 6 'dB" more power than the wanted signal.

The preceding sections have described modulation performance as a function of various distortion mechanisms. To evaluate the incremental requirements of the RPD system, practical values are required for the bandwidth and power requirements of the various modulation techniques. The values and the rationale for their selection is based on the following:
(1) For the modulation techniques, transmission bandwidths will be taken to be moderate rather than overly tight to avoid excessive degradation. Furthermore, guard bands will be used to reduce the effects of adjacent channel interference.
(2) The channel as seen by the digital modulation will be linear or quasi-linear, rather than equivalent to saturated operation. This assumption will apply to the sub-carrier and SCPC cases. For TDM operation in a dedicated transponder this assumption will also hold if more than $l$ carrier is present (multicarrier operation).
(3) The data presented has shown that the relative performance of modulation techniques depends very much on the channel characteristics (e.g. wideband versus narrowband comparisons of MSK, OQPSK and QPSK). Within the context of this study, firstorder estimates of bandwidth and $E_{b} / N_{o}$ requirements will be presented. It is expected that these estimates are accurate to within 0.5 dB . It should be noted that a revised computer program has been produced for D.O.C. in [35] that allows a detailed
comparison of all modulation techniques considered here except for DMSK. This program includes the effects of cascaded nonlinearities, adjacent and cochannel interference. Furthermore, approximations to raised-cosine pulse shaping filters are included. This program could possibly be used to refine the degradation estimates presented here.
(4) For PSK systems raised cosine filtering will be used to yield a bandwidth requirement of $1.5 \mathrm{R}_{\mathrm{S}}\left(\mathrm{R}_{\mathrm{S}}=\right.$ symbol rate; $R_{S}=R=$ bit rate for $B P S K$ and $R_{S}=R / 2$ for QPSK). To reduce the effects of adjacent channel interference a $20 \%$ guardband will be used to yield a transmission bandwidth of $1.8 \mathrm{R}_{\mathrm{s}}$.
(5) For MSK systems, a filter characteristic is not specified. However, a bandwidth of l. 25 R plus a $20 \%$ guard band yielding a transmission bandwidth of 1.5 R is reasonable.
(6) For the various modulations and the bandwidths as given in paragraphs (4) and (5), the BER performance should be within $l d B$ of the theoretical $E_{b} / N_{o}$ values given in Table 3.2. For all modulations but DMSK the $E_{b} / N_{o}$ required will be taken to be the theoretical value plus a 2 dB implementation margin. DMSK will be assumed to require a 2.5 dB margin to account for increased ISI degradation relative to DPSK. This margin accounts for the up to $1 d B$ degradation (or less) due to the channel plus an additional allowance for synchronization errors and misalignment.

The values for transmission bandwidth and ( $E_{b} / N_{o}, C / N_{o}$ ) for each modulation technique for each RPD PCM system are given in Table 3.2. These will be used in Section 5.0 to calculate the incremental DBS bandwidth and power requirements.
?

This section considers two error control techniques which can be used to improve the quality of the received radio signal. The first technique is forward acting error correction (FEC) which is effective in power limited digital communication links, but at the expense of bandwidth efficiency and system complexity.* There are two main types of FEC codes of interest, namely block and convolutional codes. In this section, the properties of both categories including coding gain (i.e. $E_{b} / N_{0}$ gain), code rate, error-correcting capability and complexity (or cost) are compared. The second technique known as error concealment (EC) is a very inexpensive method to further enhance the subjective performance of the digital radio receiver using PCM. In systems employing the EC technique, the subjective effects of bit error rate as high as $100^{-4}$ on PCM encoded audio programming has been shown to be just perceptible $[9,10]$ while no noticeable impairment due to bit error rate of $10^{-7}$ or lower was found [17].

Forward Acting Error Correction
Forward error correction (FEC.) [36] is used to combat the power limitations of digital communication links. With FEC coding redundant parity bits are added to the data bits resulting in an increase in the actual transmitted bit rate and also the system complexity, but at the same time achieving a net reduction in the amount of power required to obtain the same error rate performance as in the uncoded case. The choice of a suitable code for the radio program application will be based on the following properties of the code:

[^5](i) coding gain, i.e. $E_{b} / N_{o}$ gain;
(ii) code rate;
(iii) error-correcting capability; and
(iv) complexity (cost) of the codec.
4.1.1 Block and Convolutional Codes

For a given bit error rate (BER) performance and modulation technique, the required $\mathrm{E}_{\mathrm{b}} / \mathrm{N}_{\mathrm{o}}$ with a coded signal is smaller than that with an uncoded signal. The $E_{b} / N_{o}$ difference between the two is a function of the characteristics of the coding and decoding methods as well as the modem. Table 4.1 compares the performance of several block and convolutional codes. Note that in general, for a given code with a fixed block or constraint length, the lower the code rate, the larger the coding gain; for a fixed code rate, the larger the block or constraint length, the larger the coding gain. However, a low rate code utilizes the channel bandwidth least efficiently and a larger block or constraint length increases the system complexity. Decoding with soft decisions imposes an additional complexity penalty but with a performance improvement.

For code rates below $7 / 8$, decoders can be simple but their use results in significant bandwidth expansion. Changing the code rate from $3 / 4$ to $7 / 8$ improves the bandwidth utilization by almost 0.7 dB , while only 0.1 dB additional improvement would be obtained if the code rate is changed to $9 / 10$, but the system complexity increases enormously. Therefore, a code rate of $7 / 8$ would appear, initially at least, to represent a good trade-off between $\mathrm{E}_{\mathrm{b}} / \mathrm{N}_{\mathrm{o}}$ gain, bandwidth requirements and complexity [37].

| CODE |  | $\begin{aligned} & \text { CODE } \\ & \text { RATE } \end{aligned}$ | CODING <br> GAIN* <br> ( dB ) | ERRORCORRECTION CAPABILITY | REFERENCE |
| :---: | :---: | :---: | :---: | :---: | :---: |
| BLOCK | $\begin{aligned} & (63,30) \text { B.C.H. } \\ & (63,56) \text { B.C.H. } \\ & (127,64) \text { B.C.H. } \\ & (127,92) \text { B.C.H. } \\ & (1023,943) \text { B.C.H. } \\ & (15,11) \text { Hamming } \\ & (31,26) \text { Hamming } \\ & (23,12) \text { Golay } \end{aligned}$ | $\begin{aligned} & 0.48 \\ & 0.89 \\ & 0.50 \\ & 0.72 \\ & 0.92 \\ & 0.73 \\ & 0.84 \\ & 0.52 \end{aligned}$ | $\begin{aligned} & 3.3 \\ & 1.8 \\ & 4.3 \\ & 4.1 \\ & 4.7 \\ & 1.4 \\ & 1.8 \\ & 2.4 \end{aligned}$ | $\begin{array}{r} 6 \\ 1 \\ 10 \\ 5 \\ 8 \\ 1 \\ 1 \\ 3 \end{array}$ | $\begin{array}{lll} {[37]} & \text { Fig. } & 6.49 \\ {[38]} & \text { Table } 2 \\ {[39]} & \text { Fig. } & 5.11 \\ {[40]} & \text { Fig. } 4.7 \\ {[41]} & \text { Fig. } & 3 \\ {[40]} & \text { Fig. } & 4.4 \\ {[40]} & \text { Fig. } & 4.4 \\ {[39]} & \text { Fig. } & 1.10 \end{array}$ |
| CONVO- <br> LUTIONAL | $r=\frac{1}{2}, k=7,$ <br> Viterbi Decoding. $r=\frac{1}{2}, k=24$, <br> Viterbi Decoding. $r=3 / 4, k=9$, <br> Viterbi Decoding. $r=3 / 4, k=$, <br> Sequential. $\mathrm{r}=7 / 8, \mathrm{k}=? \text {, }$ <br> Sequential. $r=3 / 4, k=80,$ <br> Threshold (SPADE). $\mathrm{r}=7 / 8, \mathrm{k}=1176 \text {, }$ <br> Threshold (DITEC). $r=8 / 9, k=1233,$ <br> Threshold. | $\begin{aligned} & 0.50 \\ & 0.50 \\ & 0.75 \\ & 0.75 \\ & 0.875 \\ & 0.75 \\ & 0.875 \\ & 0.889 \end{aligned}$ | $\begin{aligned} & 3.4 \\ & 5.1 \\ & 2.8 \\ & 5.1 * * \\ & 4.1 * * \\ & 2.9 \\ & 3.2 \\ & 2.4 \end{aligned}$ | 5 7 5 | $\left[\begin{array}{lll}{[40]} & \text { Fig. } & 5.11 \\ {[40]} & \text { Fig. } & 5.29 \\ {[40]} & \text { Fig. } & 5.13 \\ {[42]} & & \\ {[42]} & & \\ {[43]} & \text { Fig. } 6 \\ {[43]} & \text { Fig. } 6 \\ {[43]} & \text { Fig. } 6\end{array}\right.$ |

*net $\mathrm{E}_{\mathrm{b}} / \mathrm{N}_{\mathrm{o}}$ gain
**soft decisions

Table 4.l:
Coding gain of several block and convolutional codes at output $B E R=10^{-7}$. Hard decision is used unless otherwise stated. Coding gain is based on CPSK modulation.

Table 4.1 also shows that, for convolutional codes, the $\mathrm{E}_{\mathrm{b}} / \mathrm{N}_{\mathrm{o}}$ gain depends on the decoding algorithm. Viterbi and sequential decoding can provide high $E_{b} / N_{o}$ gain, especially with soft decisions.

Viterbi decoding is a maximum likelihood decoding technique that is optimal for the additive white Gaussian noise (AWGN) channel. However, the implementation becomes impractical for constraint length $k \geqslant 10$ (approximately). For larger coding gains sequential decoding, although suboptimal, is more suitable. While Viterbi decoding is complex, present efforts are being directed towards LSI implementations [44]. However, these codecs are low rate.

It should be noted that the coding gains given in Table 4.1 are based on coherent PSK. The coding gain depends not only upon the FEC code employed but also on the modem [45]. An FEC codec is characterized by its input/output BER relationship. The $E_{b} / N_{o}$ difference for any codec input/output BER pair will depend upon the modem BER vs $E_{b} / N_{o}$ curve. Thus, the choice of modulation will affect the coding gain because each modulation technique has its own error rate performance (BER versus $E_{b} / N_{o}$ curve) on which the coding gain depends. Figure 4.1 shows the error rate performance curves for uncoded and coded CPSK and DPSK using a convolutional code with Viterbi decoder (constraint length $=7$, code rate $=\frac{1}{2}$, soft decision). It is seen from the figure that the coding gain at $B E R=10^{-5}$ is 5.2 dB for CPSK, 3.7 dB for DBPSK amd 4.4 dB for DQPSK.

There is a tradeoff between the number of errors within a block or constraint length that can be corrected and the complexity of the codec. Single-error-correcting codes use small block or constraint lengths and are simple to


Figure 4.1 Error Rate Performance With/Without R=1/2 Viterbi Codec

Coherent PSK = curve a-coded, curve
A-uncoded
Differential BPSK = Curve b-coded, curve B- uncoded.
Differential opSK = Curve c-coded, curve C-uncoded
implement. However, these codes are not applicable to PSK systems where differential encoding of data is used to combat recovered carrier phase ambiguity. In this case bit errors are likely to occur in pairs (burst errors) and it is necessary to find some means to separate paired errors. Columnwise writing and row-wise reading, symbol interleaving and $N$-symbol differential coding have been proposed [46]. The first method is appropriate for TDMA systems since this system requires compression and expansion buffers which can be utilized for writing and reading. Symbol interleaving is intended to eliminate the adverse effect of differential coding by encoding even and odd symbols independently so that a burst error is changed to a random error. Figure 4.2 shows the block diagram of a QPSK transmitter using the symbol interleaving method. The figure shows that the complexity of the system is doubled since two encoders and two decoders are needed for the transmit and receive sides, respectively. Figure 4.3 gives an example of bit correspondence of $a(4,3)$ code with symbol interleaving. Simulation results show that the symbol interleaving method improves the $E_{b} / N_{o}$ gain by about 0.5 dB [45, Figure 2]. Figure 4.4 shows the transmit system with $N$-Symbol differential coding. In this method, the $n$-th information symbol is recovered from the phases of the $(n-N)-t h$ and $n-t h$ received symbol. Thus, if the $n-$ th symbol is wrong, the $n$-th and ( $n+N$ )-th recovered symbols become erroneous. If $N$ is large, the $n$-th and ( $n+N$ )-th erroneous symbols may fall into two separate code blocks; they can then be corrected by single-error-correcting codes.

There are other ways to deal with the phase ambiguity problem. One way is as used in the MCS SLIM TDMA system [47] where a coherent QPSK modem is employed. A unique word (UW) at the beginning of the message is passed through


Figure 4.2: Application method with symbol interleaving.

| Serial input |  |  |  |  |  |  |  |  |  |  |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
| P |  | 1 |  | 3 | $s$ | 7 |  | 9 |  | 11 |
| Q |  | 2 |  | 4 | 6 | 8 |  | 10 |  | 12 |
| $\mathrm{P}^{\prime}$ | I1 |  | 3 | 5 | 91 | 7 | 9 |  | 1 | P3 |
| Q' | I2 |  | 4 | 6 | P2 | 8 | 10 |  | 12 | P4 |
| Al | 11 |  | 2 | 5 | 6 | 7 | 8 | 111 | 1 | 12 |
| A2 | I 3 |  | 4 | $\mathrm{P}_{1}$ | P2 | 9 | 10 | P3 | ${ }^{3}$ | $P_{4}$ |

Figure $4.3 \begin{aligned} & \text { An example of bit correspondence of (4, 3) code with } \\ & \text { symbol interleaving. }\end{aligned}$


Figure 4:4
Application method with N -symbol differential coding.

4 detectors (corresponding to 4 possible phases) to indicate the appropriate reference phase. Another approach is to use an FEC code which is transparent to phase ambiguity with separate codecs for each of the $I$ and $Q$ data channels $[46,48]$. A special transparent block code devised by Nakamura [49] permits the use of a single FEC decoder prior to the differential decoder. One more approach which is used in a Scientific Atlanta codec resolves the phase ambiguity by counting the number of l's in the syndrome register [50]. An excessive number of errors implies a phase ambiguity.

In the next section, the performance of FEC codes is examined with the assumption that only a number of the significant bits of a PCM word are coded. A suitable FEC code will be suggested in Section 4.1.3.
4.1.2 Error Correction of Most Significant PCM Bits

Since the bits in a PCM word have different significance, the FEC code need only be used for m most significant bits (MSB). In this section, the minimum number of significant bits that will be coded in order to give approximately the same $S / N$ output as in the case where all bits are error corrected will be determined.

Suppose that the PCM word is $n$ bits long. Of these bits, only the m most significant bits (including the sign bit) are coded. The remaining ( $n-m$ ) least significant bits are transmitted uncoded. Also suppose that the decoder output bit error rate corresponding to an input bit error rate $\mathrm{P}_{\mathrm{e}_{\mathrm{i}}}$ is $\mathrm{P}_{\mathrm{e}_{\mathrm{o}}}$ (see Figure 4.5). Following the same analysis as shown in Section 2.l.l, the signal-to-noise ratio for a PCM system with a $\mu=255-1$ aw compander and a folded binary code is given by


FIGURE 4.5 Application of FEC to most sigñificant bits in PCM word

$$
\begin{aligned}
& \left(\frac{S}{N}\right)_{\text {PCM-FB }}=\frac{1}{10.5 \times 2^{-2 n}+16\left[P_{e_{i}} \sum_{i=1}^{n-m} x_{i}^{2}+P_{e_{0}}\left(\frac{1}{4}+\sum_{i=n-m+1}^{n-1} x_{i}^{2}\right)\right]} \\
& \text { with } m \geqslant 1
\end{aligned}
$$

where

$$
x_{i}=\frac{1}{\mu}\left[(1+\mu)^{\left.2^{i-1} q_{-1}\right]}\right.
$$

(with $\mu=255$ ) is the error due to an error in the i-th least significant bit in the code word and $q=\frac{1}{2^{n-1}}$ is the uniform quantization step size.

Equation (4.1) is derived with the following assumptions:
(i) Each detected code word is not likely to contain more than one error.
(ii) The signal has Laplacian distribution.
(iii) The quantizer is $25 \%$ loaded.
(iv) Both $\mathrm{P}_{\mathrm{e}_{\mathrm{i}}}$ and $\mathrm{P}_{\mathrm{e}_{\mathrm{o}}}$ are much smaller than $\mathrm{l}_{\text {。 }}$

In (4.1), the noise components contributed by the FEC decoded bits and uncoded bits are weighted by their corresponding probabilities of error $\mathrm{P}_{\mathrm{e}_{\mathrm{o}}}$ and $\mathrm{P}_{\mathrm{e}_{\mathrm{i}}}$,
respectively. The first term in the denominator of (4.1) is the quantization noise. Using this equation, the $\mathrm{S} / \mathrm{N}$ is
plotted against the number of coded bits $m$ in Figure 4.6 for $n=9,12,15$ and for raw (uncoded) $P_{e_{i}}=10^{-3}, 10^{-4}$, $10^{-5}$ while the decoder output $B E R$ is fixed at $P_{e_{0}}=10^{-7}$ which is the required $B E R$ for $C A T V$ and high quality levels. Figure 4.7 shows the same curves with $P_{e_{o}}=10^{-6}$ for low quality level. It is seen from Figure 4.6 that it is not necessary to code every bit in a PCM word in order to obtain maximum $S / N$. In fact, for raw BER of $10^{-3}$, only about $m=\frac{n}{3}$ most significant bits need to be FEC coded to have the same $S / N$ as in the case where all bits are coded. For example, if the PCM word has $n=12$, only the 4 most significant bits require coding. The figure also shows that for lower raw BER (a lower gain FEC code is needed in this case), the number of most significant bits requiring coding is smaller. For example, while with $\mathrm{n}=12$, and a raw BER of $10^{-3}$, only the 4 most significant bits need to be coded with a coding scheme having a coding gain of 4.5 dB ; ; only the 2 most significant bits need to be coded with a coding scheme having a coding gain of $1.7 \mathrm{~dB} *$ if the raw BER is $10^{-5}$.

Comparing Figures 4.6 and 4.7 , it is seen that an increase in the decoder output BER to $10^{-6}$ from $10^{-7}$ results in negligible $S / N$ decrease for $n=9$, while there is a 6.6 dB $S / N$ decrease for $n=12$. The explanation is that, for larger $n$, the quantization noise is negligible when compared with

```
*from \(P_{e_{i}}=10^{-3}\) to \(P_{e_{0}}=10^{-7}\) results in a gross reduction
    in \(\mathrm{E}_{\mathrm{b}} / \mathrm{N}_{\mathrm{o}}\) of \(11.3-6.8=4.5 \mathrm{~dB}\) (CPSK).
\({ }^{*}\) From \(P_{e_{i}}=10^{-5}\) to \(P_{e_{o}}=10^{-7}\) results in a gross reduction
    in \(E_{b} / N_{i}^{i}\) of \(11.3 \mathrm{~dB}{ }^{{ }^{e}}{ }_{0} 9.6 \mathrm{~dB}=1.7 \mathrm{~dB}\) (CPSK).
```



the noise induced by digital errors; thus increasing the BER will decrease the $S / N$. In the case of small $n$, a increase in the BER has little effect on the $S / N$ since the quantization noise dominates the digital error noise.
4.1.3 Suitable FEC Code

There are three key aspects that are to be considered in choosing a suitable FEC code. They are (1) the number of (PCM) most significant bits to be coded, (2) the codec error-correcting capability, and (3) the codec BER performance.

As shown in Section 4.1.2 it is necessary to code only $m=n / 3$ most significant bits of an $n$-bit PCM word. This method reduces the redundancy relative to full word coding. Thus, a powerful low-rate code can be used to obtain high coding gain with possibly only a slight increase in transmission rate and hence transmission bandwidth. With $m=n / 3$, it is clear that the percentage increase in transmission rate is one-third of that for the case where all bits are coded. For example, suppose that the code rate is $\frac{3}{4}$, the increase in transmission rate is ll\%* when only one-third of the bits are coded.

If differential encoding-decoding is used to resolve phase ambiguity, then bit errors are likely to appear in pairs. If a single-error correcting codec is used then symbol interleaving would be required. Multi-error-correcting codes are suggested to be used to deal with

[^6]paired errors as well as to provide greater coding gain. Table 4.1 shows several multi-error-correcting block codes with error-correcting capability of 3 or greater. These block codes, however, are considered not suitable for the RPD application since they are either low-rate* or complex**. The table also shows three threshold-decodable convolutional codes with error-correcting capability of 5 or greater. These codes are convolutional self-orthogonal codes $[43,51,52]$. Such a class of codes have the following advantages:
(i) relatively simple implementation,
(ii) freedom from error propagation,
(iii) guaranteed correction capability beyond the minimum distance,
(iv) capability of operating at very high speed,

One disadvantage of the threshold decoding technique is that it yields a smaller coding gain than Viterbi or sequential decoding.

The last aspect considered in this section for choosing a suitable FEC code is the BER performance of the code. Section 2.1 shows that the required BER's to meet the $S / N$ specifications for high quality (including CATV) and low

[^7]quality levels are $10^{-7}$ and $10^{-6}$, respectively. It is seen from Figure 4.6 that if $m=n / 3$ most significant bits are coded, optimum $S / N$ performance can still be achieved if the raw BER input to the decoder is smaller than $10^{-3}$. Therefore, the objective for the codec is to correct a raw BER of $10^{-3}$ to $10^{-7}$ and $10^{-6}$ for high quality (including CATV) and low quality levels, respectively. These correspond to coding gains (assuming CPSK) of 4.5 dB and 3.7 dB , respectively. The three threshold-decodeable convolutional codes shown in Table 4.1 do not meet the required $B E R$ performance since they have coding gains ranging from 2.9 dB to 4.1 dB . The, other codes which meet the required $B E R$ performance objective are either low-rate or require complex decoding schemes (Viterbi decoding and sequential decoding with soft decisions). Table 4.2* compares the performance of two threshold-decodeable convolutional self-orthogonal codes with code rates of $3 / 4$ and $7 / 8$ (see Table 4.1). The table shows that the rate- $3 / 4$ code requires a transmission rate about 6\% larger than that for the $7 / 8$-code but it allows higher raw BER.
Furthermore, the constraint length of the rate- $3 / 4$ code is only 80 which is far less than that of the rate- $7 / 8$ code. Thus, the rate- $3 / 4$ codec is less complex and therefore it is proposed rather than the rate-7/8 code**.

[^8]| $\begin{gathered} \text { QUALITY } \\ \text { LEVEL } \end{gathered}$ | $\begin{array}{\|c} \text { NUMBER } \\ \text { OF } \\ \text { BITS } \\ \text { PER } \\ \text { SAMPLE } \\ n \end{array}$ | NUMBER OF CODED MSB's m | CAN- <br> DIDATE <br> CODE | INPUT <br> BER ${ }^{P} e_{i}$ | OUTPUT <br> BER <br> ${ }^{P} e_{0}$ | $\begin{aligned} & \text { CODING } \\ & \text { GAIN }^{1} \\ & (\mathrm{~dB}) \end{aligned}$ | BAND- <br> WIDTH <br> EX- <br> PANSION2 |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
| CATV | 12 | 4 | $\begin{aligned} & \mathrm{r}=3 / 4 \\ & \mathrm{k}=80 \end{aligned}$ | $6 \times 10^{-4}$ | $10^{-7}$ | $\begin{aligned} & 3.64 \\ & 3.14 \\ & 3.64 \end{aligned}$ | 11\% |
|  |  |  | $\begin{aligned} & \mathrm{r}=7 / 8 \\ & \mathrm{k}=1176 \end{aligned}$ | $4 \times 10^{-4}$ | $10^{-7}$ | $\begin{aligned} & 3.61 \\ & 3.14 \\ & 3.61 \end{aligned}$ | 5\% |
| HIGH | 11 | 4 | $\begin{aligned} & r=3 / 4 \\ & \mathrm{k}=80 \end{aligned}$ | $6 \times 10^{-4}$ | $10^{-7}$ | $\begin{aligned} & 3.61 \\ & 3.11 \\ & 3.61 \end{aligned}$ | 12\% |
|  |  |  | $\begin{aligned} & \mathrm{r}=7 / 8 \\ & \mathrm{k}=80 \end{aligned}$ | $4 \times 10^{-4}$ | $10^{-7}$ | $\begin{aligned} & 3.59 \\ & 3.14 \\ & 3.59 \end{aligned}$ | 5\% |
| LOW | 9 | 3 | $\begin{aligned} & r=3 / 4 \\ & \mathrm{k}=80 \end{aligned}$ | $10^{-3}$ | $10^{-6}$ | $\begin{aligned} & 3.24 \\ & 2.84 \\ & 3.24 \end{aligned}$ | 11\% |
|  |  |  | $\begin{aligned} & r=7 / 8 \\ & \mathrm{k}=1176 \end{aligned}$ | $7 \times 1.0^{-4}$ | $10^{-6}$ | $\begin{aligned} & 3.19 \\ & 2.84 \\ & 3.19 \end{aligned}$ | 5\% |

${ }^{1}$ Figure on top is for coherent PSK and MSK, the middle figure is for differential. BPSK and differential MSK, the bottom figure is for differential QPSK.
${ }^{2}$ This is the excess bandwidth with FEC code relative to that without coding. It is given by $\frac{m}{n}\left(\frac{1}{r}-1\right) \times 100 \%$.

Table 4.2: Comparison between the performance of rate-3/4 and rate $7 / 8$ convolutional self-orthogonal codes with threshold decoding. Note that only m most significant bits are coded.

So far, only high-rate convolutional codes have been discussed. Lower-rate codes with Viterbi or sequential decoding may give higher $E_{b} / N_{o}$ gain but larger bandwidth expansion. Another fact in using a powerful low-rate code is that more most significant bits need to be coded in order to obtain maximum $\mathrm{S} / \mathbf{N}^{*}$. Table 4.3 shows the number of most significant bits that require coding and the $E_{b} / N_{o}$ gain (based on CPSK) of a rate $-\frac{1}{2}$ convolutional code with Viterbi decoding. Comparing Tables 4.2 and 4.3 , it is seen that the rate- $\frac{1}{2}$ code with Viterbi decoding gives an $E_{b} / N_{o}$ gain about 1.3 dB higher than the rate $-\frac{3}{4}$ code with threshold decoding (for CATV quality level), but the bandwidth expansion for the rate $-\frac{1}{2}$ code is about $28 \%$ (or 1.1 dB ) larger than that for the other code. Furthermore, the rate $-\frac{1}{2}$ code uses the Viterbi decoding scheme which requires a much more complicated decoder than codes with threshold decoding.

Thus it is not worthwhile to use a low rate code in the case where the receiver cost must be kept low and where the power and bandwidth utilization factors are of the same significance. However, if the bandwidth utilization factor is far below the power utilization (i.e. satellite power is severely limited) then a powerful, low-rate code may be used to increase the system capacity at the cost of receiver complexity.

[^9]| QUALITY LEVEL | NUMBER OF BITS PER SAMPLE <br> n | NUMBER OF CODED MSB's <br> m | $\begin{aligned} & \text { INPUT } \\ & \text { BER } \\ & { }^{P_{e}} e_{i} \end{aligned}$ | OUTPUT <br> BER $P_{e_{o}}$ | CODING <br> GAIN <br> (dB) | BANDWID'TH <br> EXPANSION $\frac{m}{n}\left(\frac{1}{r}-1\right)$ |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: |
| CATV | 12 | 5 | $6.5 \times 10^{-3}$ | $10^{-7}$ | 4.89 | $41.7 \%$ |
| HIGH | 11 | 5 | $6.5 \times 10^{-3}$ | $10^{-7}$ | 4.77 | 45.5\% |
| LOW | 9 | 4 | $1.1 \times 10^{-2}$ | $10^{-6}$ | 4.30 | 44.4\% |

Table 4.3: Coding gain and bandwidth expansion of rate $-\frac{1}{2}$ convolutional code ( $k=7$ ) with Viterbi decoding. Coding gain is based on CPSK.

The effect of a bit error is to cause the $D-t o-A$ converter to output the wrong level, and when the bit in error is at or near the most significant bit, pops or clicks may be heard in the audio. A BER of $10^{-9}$ or $10^{-10}$ would be required to eliminate the clicks and pops [17]. Fortunately, a technique called error concealment. (EC) provides an inexpensive method for removing clicks and pops.

With error concealment, a parity bit is appended to each PCM word in the transmitter. In the receiver's demultiplexer, if a parity error is detected in the received word, then the previous valid word from the demultiplexer or an interpolated value is output instead of the current erroneous word. This form of "error correction" serves to reduce the perceived effects of channel bit errors, and provides a subjective enhancement.

Table 4.4 shows the bit error probability of an odd number of bits in error (estimated by one bit error) and the probability of an even number of bits in error, estimated by 2 bits in error. With error concealment the odd errors are concealed. However, while even errors can cause clicks and pops in the audio they occur very infrequently. For low frequency components in the audio, error concealment (holding the previous sample) will effectively eliminate pops and clicks; low frequency components cannot change abruptly between samples, and holding the previous sample when a parity error is detected essentially eliminates pops and clicks. The majority of energy in high-quality audio is contained in the low frequencịes. CCIR documentation indicates that the peak energy on average in high-quality

| Quality <br> Level | Bit Error Probability ( $P_{e}$ ) | Odd Probability <br> 1 Bit in Error ${ }^{1}$ | Even Probability <br> 2 Bits in Error ${ }^{2}$ | Word Error Probability ${ }^{3}$ |
| :---: | :---: | :---: | :---: | :---: |
| CATV <br> $\mathrm{n}=12$ bits <br> per PCM <br> word | $\begin{aligned} & 10^{-5} \\ & 10^{-6} \\ & 10^{-7} \end{aligned}$ | $\begin{aligned} & 1.2 \times 10^{-4} \\ & 1.2 \times 10^{-5} \\ & 1.2 \times 10^{-6} \end{aligned}$ | $\begin{aligned} & 6.6 \times 10^{-9} \\ & 6.6 \times 10^{-11} \\ & 6.6 \times 10^{-13} \end{aligned}$ | $\begin{array}{lll} 1.2 \times 10^{-4} \\ 1.2 \times 10^{-5} \\ 1.2 \times 10^{-6} \end{array}$ |
| ```HIGH n=ll bits per PCM word``` | $\begin{aligned} & 10^{-5} \\ & 10^{-6} \\ & 10^{-7} \end{aligned}$ | $\begin{aligned} & 1.1 \times 10^{-4} \\ & 1.1 \times 10^{-5} \\ & 1.1 \times 10^{-6} \end{aligned}$ | $\begin{aligned} & 5.5 \times 10^{-9} \\ & 5.5 \times 10^{-11} \\ & 5.5 \times 10^{-13} \end{aligned}$ | $\begin{array}{lll} 1.1 & \times & 10^{-4} \\ 1.1 & \times & 10^{-5} \\ 1.1 & \times & 10^{-6} \end{array}$ |
| ```LOW n=9 bits per PCM word``` | $\begin{aligned} & 10^{-5} \\ & 10^{-6} \\ & 10^{-7} \end{aligned}$ | $\begin{aligned} & .9 \times 10^{-4} \\ & .9 \times 10^{-5} \\ & .9 \times 10^{-6} \end{aligned}$ | $\begin{array}{lll} 3.6 & \times 10^{-9} \\ 3.6 \times 10^{-11} \\ 3.6 & \times 10^{-13} \end{array}$ | $.9 \times 10^{-4}$ <br> $.9 \times 10^{-5}$ <br> $.9 \times 10^{-6}$ |

1 Odd Probability $=C_{n}^{l} P_{e}\left(1-P_{e}\right)^{n-1}$
2 Even Probability $=C_{n}^{2} \cdot P_{e}^{2}\left(1-P_{e}\right)^{n-2}$
n
3 Word Error Probability $=\sum_{i=1} C_{n}^{i} P_{e}^{i}\left(l-P e^{n-l}\right.$ with $C_{n}^{m}=\frac{n!}{m!(n-m)!}$

Table 4.4: Odd, even, and word error probabilities versus bit error probability for 3 quality levels.
audio programs is below 200 Hz . The maximum change possible between $32-\mathrm{kHz}$ samples for a 200 Hz test tone with peak-to-peak variation is about $4 \%$ of peak value. This maximum change occurs at the maximum signal level, and the transient caused by missing a $4 \%$ level is attenuated due to the low-pass filtering at the output and masked by the high-level fundamental component.

When very high frequency audio components dominate the program interval (they are at nearly peak-to-peak levels), the time between noticeable clicks would be the time corresponding to the time between 2 bit (even) errors in Table 4.4. Table 4.5 illustrates the time between noticeable level changes versus the bit error probability when error concealment is employed.

Because of the longer time between samples. for the low quality $5-\mathrm{kHz}$ channel ( $90 \mu \mathrm{~s}$ instead of $31.25 \mu \mathrm{~s}$ for the $15-\mathrm{kHz}$ channel), the $5-\mathrm{kHz}$ digital audio channel with error concealment will be more susceptible to the effect of bit errors in the higher frequency signal components than the $15-\mathrm{kHz}$ channel unit, with the same BER.

Figure 4.8 provides the results of subjective tests conducted by the $\operatorname{BBC}[53,54]$ in order to grade the impairment due to channel errors with and without error concealment. It is noticed that for $P e^{\leqslant l 0^{-6}}$, error concealment brings the effect of channel errors to an imperceptible level. This error concealment technique is referred to as "zero order extrapolation", and is employed in the BBC NICAM-3 system [8] and by the Scientific Atlanta system [17].

| Quality <br> Level | Transmision <br> Rate (kbps) | $\begin{array}{r} \text { Bit Error } \\ \text { Probability } \end{array}$ | Mean Time Between <br> 2 Bits in Error |
| :---: | :---: | :---: | :---: |
| $\begin{aligned} & \text { CATV } \\ & \mathrm{n}=12 \end{aligned}$ | $32 \times(12+1)=$ | $\begin{aligned} & 10^{-5} \\ & 10^{-6} \\ & 10^{-7} \end{aligned}$ | 6.1 minutes 10.1 hours 42 days |
| $\begin{aligned} & \text { HIGH } \\ & \mathrm{n}=1 \mathrm{l} \end{aligned}$ | $\begin{array}{r} 32 \times(11+1)= \\ 384 \end{array}$ | $\begin{aligned} & 10^{-5} \\ & 10^{-6} \\ & 10^{-7} \end{aligned}$ | 7.9 minutes <br> 13.1 hours <br> 55 days |
| $\begin{aligned} & \text { LOW } \\ & \mathrm{n}=9 \end{aligned}$ | $32 \times(9+1)=$ | $\begin{aligned} & 10^{-5} \\ & 10^{-6} \\ & 10^{-7} \end{aligned}$ | $\begin{aligned} & 14.5 \text { minutes } \\ & 1.0 \text { day } \\ & 100 \text { days } \end{aligned}$ |

Table 4.5: Mean time between noticeable level shifts for low frequency components.

SUBJECTIVE IMPAIRMENT GRADE


SUBJECTIVE EFFECT OF BIT ERRORS IN PRESENCE OF PROGRAM (SOLO TRUMPET)
(1) J. R. CHEW AND M.E.B. MOFFAT, "PULSE-CODE" MODULATION FOR HIGH-QUALITY SOUND-SIGNAL DISTRIBUTION: PROTECTION AGAINST DIGIT ERRORS," BRITISH BROADCASTING CORPORATION REPORT NO. 1972/18 UDC 534.86.626376.56

Figure 4.8 = Error Concealment Performance

As the errors experienced in the system tend to be "bursty", that is, several errors followed by a relatively long period before another error, the technique called by the BBC "first order interpolation" may be of value. It will be tried in a subsequent model of the DATE [9] system. In this technique, three successive sample words are stored in chronological order. Parity checks are made on the middle word of the three. If there is an error detected, the middle word is replaced by the average of the first and last word. Using this technique, the "just perceptible impairment" error rate may be relaxed to $10^{-5}$ from $10^{-7}$.
4.3 Comparison Between FEC and EC

The advantages and disadvantages of FEC and EC are summarized as follows:
(i) With the use of the rate-3/4 FEC, the carrier-tonoise density ratio is reduced by typically 3.6 dB coding gain (depending on the modem). The error concealment technique reduces the carrier-to-noise density ratio only about 1.4 dB .
(ii) If FEC coding of the MSB's of the PCM word is used the transmission bit rate increases less than 12\% from that with no coding. If EC is used the bit rate increases by $8.3 \%$ for CATV quality; $9.1 \%$ for high quality; $11 \%$ for low quality.
(iii) The average time between errors with $B E R=10^{-7}$ is 26 seconds for CATV quality ( $384 \mathrm{kbits} / \mathrm{s}$ ); 28 seconds

FIt is shown in Figure 4.8 that with the use of EC, the BER can increase to $10^{-5}$ from $10^{-7}$ for CATV and high quality levels. This corresponds to a gross $\mathrm{E}_{\mathrm{b}} / \mathrm{N}_{\mathrm{O}}$ gain of 1.7 dB for CPSK modulation. Subtracting the energy per bit reduction of 0.3 dB due to the extra parity bit which is added to each PCM word ( 12 bits/word) leaves 1.4 dB as the net $E_{b} / N_{o}$ gain.
for high quality ( $352 \mathrm{kbits} / \mathrm{s}$ ) . For low quality ( 288 kbits/s), the required BER is $10^{-6}$ for which the average time between errors is 3.5 seconds. Without using EC, clicks or pops due to digital errors may be perceived at these time intervals. With the use of EC, only the clicks or pops due to even numbers of bits in error within a PCM word can be perceived. The mean time between even errors is on the order of minutes for all three quality levels with $\operatorname{BER}=10^{-5}$ (see Table 4.5).
(iv) Hardware implementation for the EC circuit is very simple while that for the FEC circuit can be very complex.

From (i) to (iv) above, it is seen that implementation of EC is simpler than that of $F E C$ but the latter provides a larger $E_{b} / N_{o}$ gain. Furthermore, the use of $F E C$ would realize the required $S / N$ values (i.e. objective performance) while EC could only improve the subjective performance without indicating whether the required $\mathrm{S} / \mathrm{N}$ values are achieved. Further study would be necessary to relate objective and subjective performance measures.

Thus, in order to meet the $\mathrm{S} / \mathrm{N}$ requirements, FEC can be used. EC can also be included to improve the subjective performance. Note that $F E C$ automatically provides parity. Provided that this parity can be related to errors on a PCM word-by-word basis then an extra EC parity bit would not be required. Also note that $E C$ can be used for audio signals only. For data transmission, only FEC can be employed to reduce the required carrier-to-noise ratio at the receiver.

A quantitative comparison of satellite power and bandwidth utilization efficiencies for radio program distribution systems using the modulation candidates identified in Section l. 2 is presented in this section. For the dedicated transponder case, the satellite power and bandwidth required to support the radio carrier(s) are compared to those required for a single video carrier. For the shared transponder case, the satellite power and bandwidth required to support both the radio and video carriers are compared to those required for a single video carrier.
5.1 Power and Bandwidth Utilization Efficiencies for Dedicated Transponder

The power and bandwidth required for the radio program carrier(s) relative to those for a single video carrier are:

$$
\begin{align*}
& \text { Relative power } r_{d}=\frac{\left(\frac{C}{N_{O}}\right)}{\left(\frac{C}{N_{O}}\right)}  \tag{5.1}\\
& \text { Relative bandwidth } q_{d}=\frac{B W_{R}}{B W_{V}} \tag{5.2}
\end{align*}
$$

where

$$
\begin{aligned}
\left(\frac{C}{N_{O}}\right)_{R}= & \text { carrier-to-noise density ratio for the } \\
& \text { composite radio program signal at the DBS } \\
& \text { receiver input }
\end{aligned}
$$

$$
\begin{aligned}
\left(\frac{\mathrm{C}}{\mathrm{~N}_{\mathrm{O}}}\right)_{\mathrm{V}}= & \text { carrier-to-noise density ratio for the video } \\
& \text { signal at the } \mathrm{DBS} \text { receiver input } \\
= & 87 \mathrm{~dB}-\mathrm{Hz} \\
\mathrm{BW}_{\mathrm{R}}= & \text { bandwidth required for the composite program } \\
& \text { radio signal. } \\
\mathrm{BW}_{\mathrm{V}}= & \text { video receiver noise bandwidth. } \\
= & 18 \mathrm{MHz}
\end{aligned}
$$

The carrier-to-noise density ratio for the digital radio signal required by the demodulator to attain a specified BER value is given by (in $d B$ ):

$$
\begin{equation*}
\left(\frac{C}{N_{0}}\right)_{R}=\frac{E_{b}}{N_{0}}-D+\log \left(N \alpha_{i} R_{i}\right)+M+L \tag{5.3}
\end{equation*}
$$

where

$$
\begin{aligned}
\frac{\mathrm{E}_{\mathrm{b}}}{\mathrm{~N}_{\mathrm{O}}}= & \text { energy per bit-to-noise density ratio required } \\
& \text { by the demodulator to meet a specified } B E R \\
& \text { value. }
\end{aligned}
$$

$D=$ net coding gain with forward error correction.
$N=$ number of multiplexed radio channels.
$\alpha_{i}=$ multiplexing overhead.
$R_{i}=$ encoding rate for one radio channel (i stands for one of the three quality levels).

$$
\begin{aligned}
M= & \text { modem implementation margin. } \\
\mathrm{L}= & \text { loading degradation (not including } \\
& \text { intermodulation and interference which will be } \\
& \text { discussed in section } 6.0 \text { ). This value reflects } \\
& \text { the fact that the satellite operating point } \\
& \text { deviates from saturation for the new modes of } \\
& \text { operation. }
\end{aligned}
$$

This equation applies to both SCPC and TDM'd signals.

The bandwidth required for the radio signal (see Section 3.3) is given by:

$$
B W_{R}=\left[\begin{array}{ll}
1.50 N \alpha_{i} R_{i} x \quad 1.2 & \text { for BPSK }  \tag{5.4}\\
0.75 N \alpha_{i} R_{i} x 1.2 & \text { for QPSK } \\
1.25 N \alpha_{i} R_{i} x_{i} 1.2 & \text { for MSK }
\end{array}\right.
$$

The factor of 1.2 in (5.4) is for the guard band between the carriers and is $20 \%$ of the occupied bandwidth. It should be noted that if the transponder is occupied by only one TDM'd carrier, then the guard band is no longer needed.

The relative power $r_{d}$ and relative bandwidth $q_{d}$ for the dedicated transponder case are plotted against the number of 15 kHz radio channels in Figures 5.1 and 5.2. Both SCPC and TDM cases are considered for three quality levels (CATV, high quality, lower quality). For each of these cases, coherent and differential PSK and MSK modulations are evaluated. The following assumptions apply to the results shown in Figures 5.1 and 5.2:




Figure 5.2b: Relative Bandwidth for High Quality
Figure 5.2c: Relative Bandwidth for Iow Quality

- transmission bit rate per 15 kHz radio channel (no overhead bits assumed) and the BER performance of the digital modems are shown in Table 3.2.
- No FEC coding is used.
- 2 dB modem implementation margin for BPSK, QPSK, MSK, DPSK modems, 2.5 dB for DMSK modem.
- In the TDM case, all radio channels are multiplexed to modulate a single carrier.
- In the SCPC case, all carriers are equally spaced.
- 20\% guard band between the SCPC carriers. Note that no guard band is used in the TDM case since only one TDM'd carrier is considered and the occupied bandwidth will be less than the transponder bandwidth.
- 1 dB satellite output backoff (loading degradation) for TDM case.
- 4.5 dB satellite output backoff (loading degradation) for SCPC case (for 3 carriers or more).
- Degradations due to intermodulation and interference are not considered.

The following observations are derived from the results shown in Figures 5.1 and 5.2:
(i) Single TDM carrier operation is more power and bandwidth efficient than SCPC for a dedicated transponder at maximum capacity. About 3.5 dB of

[^10]satellite power and $20 \%$ of transponder bandwidth are saved if the digitized radio signals are multiplexed to modulate a single carrier* as compared to the case where each radio signal modulates a single carrier. TDM operation utilizes the transponder power more efficiently (single TDM carrier) because it requires less transponder backoff and bandwidth more efficiently than SCPC because it requires less aggregate guard band. For multiple rim carrier operation, the power and bandwidth utilization efficiencies are expected to be the same as SCPC operation since the $I M$ interference level is approximately the same in both cases (both represent multi-carrier operation).
(ii) Modems which employ coherent detection require less power than those which employ differential detection. Differential QPSK is the least power efficient, but differential and coherent QPSK utilize bandwidth most efficiently.
5.2 Power and Bandwidth Utilization Efficiencies for Shared Transponder

Two access methods are considered for the shared transponder case: SCPC and subcarrier. The relative power and bandwidth ratios for the two methods are given below.
5.2.1 Shared Transponder SCPC Case

$$
\begin{equation*}
\text { Relative power } r_{s}=\frac{\left(\frac{C}{N_{O}}\right)+\left(\frac{C}{N_{O}}\right)_{R}}{\left(\frac{C}{N_{O}}\right)} \tag{5.5}
\end{equation*}
$$

$$
\text { Relative Bandwidth } q_{S}=\frac{B W_{V}+B W_{R}+\text { Guard Band }}{B W_{V}} \text { (5.6) }
$$

where

$$
\begin{aligned}
& \left(\frac{\mathrm{C}}{\mathrm{~N}_{\mathrm{O}}}\right)_{\mathrm{V}}=87.0 \mathrm{~dB}-\mathrm{Hz} \\
& \left(\frac{\mathrm{C}}{\mathrm{~N}_{\mathrm{O}}}\right)_{\mathrm{R}} \text { is given by (5.3) } \\
& \mathrm{BW} W_{\mathrm{V}}=18 \mathrm{MHz} \\
& \mathrm{BW}_{\mathrm{R}} \text { is given by }(5.4)
\end{aligned}
$$

In this case, a guard band equal to the bandwidth occupied by the SCPC carriers $\left(B W_{R}\right)$ is used between the video carrier and radio carriers to avoid intermodulation products $\mathrm{f}_{\mathrm{V}}+\mathrm{f}_{\mathrm{r}_{\mathrm{l}}}-\mathrm{f}_{r_{2}}$ (see Figure 5.3) [55].

Figures 5.4 and 5.5 show the relative power and the excess bandwidth respectively required for coherent detection of PSK and MSK. Figure 5.6 shows the relative power required for the differential detection of PSK and MSK. The following assumptions apply to the results shown in Figures 5.4, 5.5 and 5.6:

- Transmission rate per 15 kHz radio channel (no overhead bits are assumed) and the BER performance of the digital modems are shown in Table 3.2.
- No FEC coding is used.
- 2 dB modem implementation margin for BPSK, QPSK, MSK, DPSK modems, 2.5 dB for DMSK modem.


FIGURE-5.3 TV PLUS LARGE NUMBER OF SCPC CARRIFRS TRANSPONDER SHARTNG ARRANGEMENT



Pigure 5. 6: Relative power for pileferental Detection of

- Guard band equal to the bandwidth occupied by the SCPC radio carriers is used between the video carrier and the SCPC carriers (see Section 6.1.2.3).
- SCPC carriers are equally spaced.
- 20\% guard band between the SCPC radio carriers.
- 2.0 dB composite satellite output backoff (loading degradation) to reduce intermodulation.
- Degradations due to intermodulation and interference are not considered。

It is seen from Figures 5.4 and 5.6 that, in terms of relative power required, DQPSK is the least efficient, next is DMSK and then DBPSK. Coherent detection of PSK (both binary and quadrature $P S K$ ) and MSK requires the least additional power. In terms of additional bandwidth required, Figure 5.5 shows that (coherent and differential) QPSK is the most efficient, next is MSK and then BPSK.
5.2.2 Shared Transponder Sub-carrier Case Without Video Associated Audio Subcarrier

In this case a single TDM'd subcarrier is used. The relative power and bandwidth are given by:

Relative power $r_{S}=\frac{\left(\frac{C}{N_{O}}\right)_{R F}}{\left(\frac{C}{N_{O}}\right)}$

Relative Bandwidth $q_{S}=\frac{B W_{I}}{B W_{V}}$
where

$$
\begin{aligned}
&\left(\frac{\mathrm{C}}{\mathrm{~N}_{\mathrm{O}}}\right)_{\mathrm{RF}}= \text { the total } \mathrm{RF} \text { power to noise density } \\
& \text { ratio required to support both the video } \\
& \text { and radio signals. } \\
&\left(\frac{\mathrm{C}}{\mathrm{~N}_{\mathrm{O}}}\right)_{\mathrm{V}}= 87.0 \mathrm{dB-Hz} \\
& \mathrm{BW}_{\mathrm{T}} \quad \text { is the Carson's rule bandwidth of the } \mathrm{RF} \text { signal } \\
& \quad \text { including both } \mathrm{TV} \text { and radio. }
\end{aligned}
$$

$$
\mathrm{BW}_{\mathrm{V}}=18 \mathrm{MHz}
$$

The total RF carrier to noise density ratio $\left(C / N_{o}\right)_{R F}$ is found by solving the following simultaneous equations [56]:

$$
\begin{align*}
& \left(\frac{S}{N_{O}}\right)_{V}=\sigma\left(\frac{C}{N_{O}}\right) \frac{\Delta f_{V}^{2}}{\bar{f}_{V}^{2}}  \tag{5.9a}\\
& \left(\frac{C}{N_{O}}\right)_{R}=\frac{1}{2}\left(\frac{C}{N_{O}}\right)_{R F} \frac{\Delta f_{R}^{2}}{\mathrm{f}_{R}^{2}}  \tag{5.9b}\\
& B W_{T}=2\left(\Delta f_{V}+\Delta f_{R}+f_{V}\right) * \tag{5.9c}
\end{align*}
$$

*Due to the assumption that the Carson's rule bandwidth is calculated by adding the peak frequency deviations, the results in this section are conservative.
where
$\left(\frac{S}{N_{o}}\right)$ is the video signal-to-noise ratio.
$\Delta f_{V}$ is the peak frequency deviation of the carrier due to video signal.
$\mathrm{f}_{\mathrm{V}}$ is the peak baseband video frequency (4.2 MHz).
$\left(\frac{C}{N_{0}}\right)$ is given by $(5.3)$ and refers to the subcarrier.
$\Delta f_{R}$ is the peak frequency deviation of the carrier due to radio signals.
$f_{R}$ is the subcarrier frequency carrying the radio signals.

Solving. (5.9a), (5.9b), and (5.9c) for ( $\left.C / N_{o}\right)_{R F}$ we obtain:

$$
\begin{equation*}
\left(\frac{C}{N_{O}}\right)_{R F}=\frac{\left[f_{V} \sqrt{\frac{1}{6}\left(\frac{S}{N_{O}}\right)}+f_{R} \downarrow \overline{2\left(\frac{C}{N_{O}}\right)}\right]_{R}^{2}}{\left[\frac{B W_{T}}{2}-f_{V}\right]^{2}} \tag{5.10}
\end{equation*}
$$

It should be noted that if there is no radio signal, ie. $\left(\mathrm{C} / \mathrm{N}_{\mathrm{O}}\right)_{\mathrm{R}}=0$, then $\left(\mathrm{C} / \mathrm{N}_{\mathrm{O}}\right)_{\mathrm{RF}}=\left(\mathrm{C} / \mathrm{N}_{\mathrm{O}}\right)_{\mathrm{V}}=87 \mathrm{~dB}-\mathrm{Hz}$, then $\left(S / N_{o}\right)_{V}=97.88$ dB if $\Delta f_{V}=6 \mathrm{MHz}$. Thus, (5.10) can be written as:

$$
\begin{equation*}
\left(\frac{C}{N_{O}}\right)_{R F}=\frac{\left[1.343 \times 10^{11}+f_{R} \downarrow \overline{2\left(\frac{C}{N_{O}}\right)}\right]^{2}}{\left[\frac{B W_{T}}{2}-f_{V}\right]^{2}} \tag{5.11}
\end{equation*}
$$

Using (5.7) and (5.11), the relative power is plotted against the number of radio channels for several values of relative bandwidth $\mathrm{g}_{\mathrm{S}}=\mathrm{BW}_{\mathrm{T}} / \mathrm{BW}_{\mathrm{V}}$ and for three radio quality levels in Figures 5.7 and 5.9. Figures 5.8 and 5.10 show relative bandwidth as a function of the number of radio channels for several values of relative power $r_{s}$ and for CATV quality level. The following assumptions apply to (5.11) and the results shown in Figures 5.7 to 5.10:

- Transmission rate per 15 kHz radio channel (no overhead bits is assumed) an the BER performance of the digital modems are shown in Table 3.2.
- No FEC coding is used.
- 2 dB modem implementation margin for BPSK, QPSK, MSK, DPSK modems, 2.5 dB for DMSK modem.
- In the Carson's rule bandwidth equation, the top baseband frequency is fixed at 4.2 MHz and peak deviations of the main carrier due to video and sub-carrier are added.
- Only video signal is considered (i.e. no audio subcarrier*) and the Carson's rule bandwidth for the video signal alone is 18 MHz .

[^11]

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- All radio channels are TDM'd to modulate a single subcarrier having a nominal frequency of 5.68 MHz which is half-way between the highest baseband video frequency ( 4.2 MHz ) and the frequency of the second harmonic of the colour sub-carrier ( 7.16 MHz ) to make the most efficient use of the available baseband bandwidth (see Figure 5.11).
- Transponder is operated at saturation (i.e. no output back-off).
- Degradations due to intermodulation and interference are not considered.

Although the results shown in Figures 5.7 to 5.10 are somewhat conservative, it still can be seen that the subcarrier technique is more power efficient than the shared transponder SCPC case when BPSK is used. For example, Figure 5.8 shows that, for BPSK with coherent detection, only the transmission bandwidth has to be increased by about $39 \%$ (no increase in transmitted power) in order to transmit 5 CATV quality radio channels in addition to the video signal using subcarrier technique without degrading the performance of the video.signal. On the other hand, if the shared transponder SCPC technique is used to transmit 5 CATV quality BPSK radio channels in addition to a video signal, the transmission bandwidth has to be increased by about $39 \%$ (only $19 \%$ if QPSK is used) and the transmitted power is increased by 0.53 dB (see Figures 5.4 and 5.5). Note that the results predicted by (5.11) apply only to a small number of radio channels*, and the analysis in this section assumes that all radio channels are multiplexed to
*If a large number of radio channels are carried by the sub-carrier, the bandwidth required for the radio signal would be wide enough so that the video signal and the second harmonic of the colour sub-carrier will cause interference to the radio signal which is not accounted for in (5.ll).


Figure 5.11: Video baseband and radio sub carrier spectra.
(digitally) modulate a single sub-carrier. Furthermore, it is assumed that the demodulator is operated above threshold.

It can be seen that coherent BPSK, QPSK and MSK require the same RF power and bandwidth. However, it is possible to transmit more QPSK radio channels within the constrained baseband bandwidth with the cost of more RF power. and bandwidth. Furthermore, Figures $5.9 a, \quad b, c$ show that differential detection of QPSK (DQPSK) requires more power than DMSK and DBPSK.
5.2.3 Shared Transponder Subcarrier Case With Video Associated Audio Subcarrier

The relative power and bandwidth with a TV-associated audio sub-carrier added are given by (5.7) and (5.8). In (5.7), the term $\left(C / N_{o}\right)_{R F}$ now is the total RF power required to support the video, audio and the radio signals and is found by solving the simultaneous equations (5.9a), (5.9b) and the following:

$$
\begin{align*}
& \left(\frac{C}{N_{O}}\right)_{A}=\frac{1}{2}\left(\frac{C}{N_{O}}\right)_{R F} \frac{\Delta f_{A}^{2}}{f_{A}^{2}}  \tag{5.12a}\\
& B W_{T}=2\left(\Delta f_{V}+\Delta f_{A}+\Delta f_{R}+f_{V}\right) \tag{5.12b}
\end{align*}
$$

where
$\left(C / N_{o}\right)_{A}$ is the carrier-to-noise density of the audio signal.
$\Delta f_{A} \quad$ is the peak frequency deviation of the main carrier due to the audio signal.
$f_{A}$. is the audio sub-carrier frequency.

Similar to the preceding section, we obtain:

$$
\begin{equation*}
\left(\frac{\mathrm{C}}{\mathrm{~N}_{\mathrm{O}}}\right)_{\mathrm{RF}}=\frac{\left[1.469 \times 10^{11}+\mathrm{f}_{\mathrm{R}} \sqrt{2\left(\frac{\mathrm{C}}{\mathrm{~N}_{\mathrm{O}}}\right)}\right]^{2}}{\left[\frac{B W_{T}}{2}-f_{V}\right]^{2}} \tag{5.13}
\end{equation*}
$$

The relative power is found from (5.7) and (5.13), and is plotted against the number of radio channels for several values of relative bandwidth and for CATV quality level only in Figures 5.12 and 5.13. The relative bandwidth is plotted in Figures 5.14 and 5.15. The same assumptions in Section 5.2.2 apply here except that the nominal radio subcarrier frequency is fixed at 6.08 MHz which is half way between the audio sub-carrier at 5 MHz and the second harmonic of the colour sub-carrier (see Figure 5.16). The choice of 5 MHz for the audio sub-carrier is arbitrary, but this choice yields conservative use of baseband bandwidth and high audio signal-to-noise ratio if the video signal is low-pass filtered (see Figure 5.17). Furthermore, the audio carrier-to-noise density ratio is assumed to be 65.0 $\mathrm{dB}-\mathrm{Hz}$.

It is seen from Figures 5.7a (or 5.9a) and 5.12 (or 5.14) that with the addition of an audio sub-carrier, the RF power has to be increased by about 0.9 dB (for 5 TDM'd radio channels) in order to meet the required video and radio carrier-to-noise density ratios. For coherent detection, Figure 5.12 shows that if the RF noise bandwidth is kept fixed at 18 MHz , about 5.4 dB excess power is needed to support one audio and five CATV quality radio channels in addition to the video signal. If




Figure 5.16: Video baseband and audio and radio subcarriers spectra


NOTE: This figure suggests that the optimum placement of the audio subcarrier is as close to the upper video baseband as possible when overdeviation is not used. This may not be true of an overdeviated system.

Figure 5.17: Audio Subcarrier Placement in Relation to Video Baseband Filtering (ref. TV Sound Subcarrier Transmission, Johannsen et al.y IEEE TBC Sept. 1974)
the power is to be unchanged, about $47 \%$ excess $R F$ bandwidth (see Figure 5.13) is needed which is $5 \%$ above that required in the case where the audio signal is not considered.

As in all previous cases, DQPSK again requires more power than DMSK and DBPSK as shown in Figure 5.14. It should be noted that the baseband bandwidth between the audio signal and the second harmonic of the colour subcarrier is 7.16 $5.0=2.16 \mathrm{MHz}$ which could be too small to accomodate a certain number of radio channels. Table 5.1 gives the upper limit of the number of radio channels that can be added within this bandwidth.
5.3 Power and Bandwidth Utilization Efficiencies With FEC Coding

In the above sections, the relative power and bandwidth utilizations with the addition of radio programs are quantified without FEC coding. This section evaluates these factors with FEC coding of the digitized radio programs. It is assumed that the $F E C$ code used here is the rate $-\frac{3}{4}$ convolutional code with threshold decoding as proposed in Section 4.l.3. The code gives a coding gain of 3.6 dB based on coherent PSK modulation.

The relative power and bandwidth utilizations for a dedicated transponder with no FEC coding are shown in Figures 5.1 and 5.2. With the use of the rate $-\frac{3}{4}$ code, the relative power utilization is decreased by the coding gain and the bandwidth utilization is increased by the bandwidth expansion of the code (coding gain and bandwidth expansion of the rate $-\frac{3}{4}$ code are given in Table 4.2).

For the shared transponder case, the relative power (for coherent detection only) and bandwidth utilizations are shown in Figures 5.18 and 5.19 for FMTV/SCPC case and

|  |  | Maximum number of radio channels for |  |  |
| :---: | :---: | :---: | :---: | :---: |
| Quality | Bit Rate <br> Per Channel | BPSK | MSK | QPSK |
| CATV | 384 kbps | 3 | 3 | 6 |
| HIGH | 352 kbps | 3 | 4 | 6 |
| LOW | 288 kbps | 4 | 5 | 8 |

Table 5.1: Maximum number of radio channels that can be accommodated within 2.16 MHz . Transmission bandwidth per channel = 1.5 x bit rate $+20 \%$ guard band ( $10 \%$ between the radio and audio, $10 \%$ between radio and the second harmonic of the colour sub-carrier. See Figure 5.16).



Figures 5.20 and 5.21 for FMTV/Subcarrier case. Comparing the results with FEC coding and those without coding, it is seen that the power utilization improvement due to coding is higher for the FMTV/subcarrier case than the FMTV/SCPC case. For example, for 5 radio programs to be transmitted, FEC coding with the rate $-\frac{3}{4}$ code decreases the composite RF (video plus radio) power requirement by 0.9 dB for the FMTV/Subcarrier case (see Figures 5.12 and 5.21), and by only 0.3 dB for $\mathrm{FMTV} / \mathrm{SCPC}$ case (see Figures 5.4 and 5.18).
5.4 Summary

From the results and discussions in Sections 5.1 and 5.2, we arrive at the following remarks:

For dedicated transponder:
(i) The use of the time division multiplexing (TDM) technique results reduces up to 3.5 dB the power required over the use of SCPC.
(ii) Differential QPSK requires about 2.3 dB more power than coherent QPSK. However, the bandwidth requirement is the same for both differential and coherent detections.
(iii) QPSK utilizes the bandwidth most efficiently. Time division multiplexing of the QPSK signal requires the least bandwidth.
(iv) With the use of FEC coding, the power requirement is decreased by the coding gain and the bandwidth requirement is increased by the bandwidth expansion of the code.

For shared transponder:
(i) While the use of SCPC signals always requires more power and bandwidth (i.e. relative power is always greater than 0 dB ), the use of the subcarrier technique can trade power for bandwidth and vice versa.
(ii) SCPC/QPSK requires the least bandwidth expansion, followed by SCPC/MSK and then SCPC/BPSK.
(iii). The sub-carrier technique is considered more power efficient than the SCPC technique. However, this technique is applicable only for a small number of radio channeḷs (less than 8 TDM/QPSK or 4 TDM/BPSK channels) due to limited baseband bandwidth.
(iv) Addition of an audio sub-carrier increases the power requirement by 0.9 dB or the bandwidth requirement by about 5\% for 5 TDM'd radio channels.
(v) Differential QPSK requires the most additional power in both shared transponder SCPC and subcarrier cases. In the latter case, DQPSK also requires more transmission bandwiath.

The above remarks suggest that, in terms of power and bandwidth utilization efficiencies, TDM is better than SCPC for a dedicated transponder. For a shared transponder, a TDM sub-carrier is favoured over SCPC. Furthermore, for both dedicated and shared transponders, coherent QPSK is the most power and bandwidth efficient. The use of differential detection increases the power requirement and, for the shared transponder subcarrier case only, the transmission bandwidth requirement is also increased. Thus the use of differential detection is less desirable based on for high efficiency in power and bandwidth utilization decision criteria only.
6.0 INTERFERENCE AND TRANSPONDER SHARING ANALYSIS

This section gives a quantitative analysis of the interference to both the video and radio signals and the resulting effect on system capacity. There are two interference problems to be discussed. The first is internal interference which includes direct interference between the video and/or radio signals and interference due to intermodulation products generated by video and/or radio carriers sharing the same transponder. The second is external interference which includes interference from staggered cross-polarized transponders on the same satellites and from co-channel (co-polarized and crosspolarized) transponders on adjacent satellites. The analysis is performed with the assumption that no FEC coding is employed for the radio program. However, the same approach undertaken here can be applied to radio program distribution with FEC coding.

## 6.1 <br> Internal Interference: Intra-transponder Interference

This section considers interference between signals within the wanted transponder. The two scenarios to be considered are: dedicated transponder (radio signals only) and shared transponder (FMTV plus radio signals). In the following sections, the effects of interference on the performance of the wanted $T V$ and radio signals are quantitatively analyzed.
6.1.1 Interference Between Radio Signals Within a Dedicated Transponder

The channel arrangement of SCPC carriers in a dedicated transponder must consider the following interference problems:
(i) adjacent radio channel interference, and
(ii) interference due to intermodulation products.
6.1.1.1 Adjacent Radio Channel Interference

The adjacent channel interference ( $A C I$ ) to the wanted digital radio channel from the upper and lower adjacent radio channels is evaluated here for different digital demodulator receive $I F$ filters. For this analysis it is assumed that:

- the wanted PSK signal power spectrum input to the receive IF filter has raised cosine shape with rolloff factor $\alpha$ as follows

$$
P(\omega)=\left[\begin{array}{rr}
1, & \text { for } \quad 0 \leqslant|\omega| \leqslant \frac{\pi}{T}(1-\alpha) \\
\frac{1}{2}\left[1-\sin \frac{T}{2 \alpha}\left(\omega-\frac{\pi}{T}\right)\right], & \text { for } \\
\frac{\pi}{T}(1-\alpha) \leqslant|\omega| \leqslant \frac{\pi}{T}(1+\alpha)
\end{array}\right.
$$

where $\omega$ is the frequency relative to the IF carrier frequency and $T$ is the symbol duration.

- the adjacent channel signals have the same power and symbol rate as the wanted one.

The output power spectrum $P_{o}(\omega)$ of a filter is related to the input power spectrum $P_{i}(\omega)$ and the filter frequency response $H(\omega)$ by [5]

$$
P_{0}(\omega)=P_{i}(\omega)|H(\omega)|^{2}
$$

Thus, the received wanted and interfering power (from upper and lower adjacent channels) spectra are given by

$$
\begin{aligned}
& P_{W}(\omega)=P(\omega)|H(\omega)|^{2} \\
& P_{I}(\omega)=P\left(\omega-\omega_{A}\right)|H(\omega)|^{2}+P\left(\omega+\omega_{A}\right)|H(\omega)|^{2}
\end{aligned}
$$

where
${ }^{\omega_{A}}$ is the frequency spacing between adjacent channels, and
$P(w)$ is the input power spectrum.

Therefore, the received $I F$ carrier-to-interference ratio is

$$
\frac{C}{I}=\frac{\int_{-\infty}^{\infty} P_{W}(\omega) d \omega}{\int_{-\infty}^{\infty} P_{I}(\omega) d \omega}
$$

Assuming, that the receive $I F$ filter is symmetrical about the $I F$ carrier frequency, $C / I$ is given by

$$
\frac{C}{I}=\frac{1}{2} \frac{\int_{-\infty}^{\infty} P(\omega)|H(\omega)|^{2} d_{\omega}}{\int_{-\infty}^{\infty} P\left(\omega-\omega_{A}\right)|H(\omega)|^{2} d_{\omega}}
$$

The carrier-to-interference ratio is plotted versus the adjacent channel spacing in Figure 6.1 for transmitted spectrum with $\alpha=0.5$ and for $3 r d, 4$ th and 5 th order Butterworth receive filters* having $3-\mathrm{dB}$ bandwidths of $\frac{l}{T}$.

* The amplitude response of the Butterworth filter of order $n$ and 3 dB bandwidth $\omega_{3}$ is given by

$$
|H(\omega)|^{2}=\frac{1}{1+\left(\omega / \omega_{3}\right)^{2 n}}
$$

Figure 6.2 shows the $E_{b} / N_{o}$ degradation at $B E R=10^{-6}$ for QPSK due to upper and lower adjacent channel interference. The degradation results are obtained from a simulation program developed by MCS for NICSMA [57]. From the figure, it is seen that the degradation at $B E R=10^{-6}$ for QPSK is less than 0.15 dB if the $\mathrm{C} / \mathrm{I}$ is about 30 dB or greater. As seen from Figure 6.1, $C / I$ of 30 dB can be obtained by using a $50 \%$ rolloff transmit filter and a 3 rd order Butterworth receive filter and a spacing between adjacent channels of 1.8 times the symbol rate*. With smaller adjacent channel spacing, a higher order receive filter must be used to maintain the same $E_{b} / N_{o}$ degradation.

Rosenbaum [58] has derived formulas to evaluate the expected error performance for coherent and differential PSK (CPSK and DPSK) with the presence of thermal noise and peaklimited interference. He has shown that the BER performance of binary PSK (BPSK) is degraded less than quadrature PSK (QPSK) by the addition of interference and also that differential detection suffers more degradation than coherent detection.

As discussed in Section 3.2.4, QPSK performs better than MSK in tightly bandlimited applications with and without adjacent channel interference (channel spacings \& 1.3 R approximately). As will be shown in the next section the dedicated transponder case is power limited. Thus there is more bandwidth available than required. Therefore, the use of MSK for this transponder category is not constrained by adjacent channel interference.

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Figure 6.1: Carrier-to-adjacent channe1 interference as a function of channe 1 spacing for $3 r d$, 4 th and 5 th order Butterworth filters.


Figure 6.2 Eb/No degradation at $B E R=10^{-5}$ for OPSK due to adjacent channel interference.

A mechanism that can cause the adjacent channel interference to increase is intermodulation (for multi-carrier operation) or spectral spreading (for single carrier operation - see Figure 6.3 [59], also [60, 61]) due to the nonlinear characteristic of the earth station HPA. Extensive computer simulation [62] shows that in order to avoid adjacent channel interference due to spectrum spreading, the HPA must not be operated at saturation even for single carrier transmission. Alternatively, if the HPA is driven to saturation, a filter (or filters) should be placed after the HPA to reduce the out-of-band power which may cause interference into adjacent channels on the uplink.
6.l.1.2 Interference Due to Intermodulation in Satellite TWT

In the dedicated transponder case, SCPC carriers will share a common non-linear amplifier (e.g. TWT) resulting in intermodulation (IM) products which cause interference unless sufficient input backoff is provided for the IWT to operate in the linear region or the carriers are assigned in such a way that the IM products fall into unoccupied frequency bands. The effect of IM interference on the system capacity is evaluated in this section. Frequency assignment plans for no or minimum IM interference are also discussed.

The performance degradation for each signal (i.e. $E_{b} / N_{o}$ degradation) due to IM interference is evaluated by solving the set of link equations. Assuming that the system is homogeneous, the link equations are:


HPA output spectra for filtered OPSK, filter (a), 30-percent rolloff Nyquist ( $30 \mathrm{Msymbols} / \mathrm{s}$ ) operated at $B T=1.00$.


HPA output spectra for filtered QPSK, filter (a), 30-percent rolloff Nyquist ( $30 \mathrm{Msymbol} / \mathrm{s}$ ) operated at $B T=1.15$.

Figure 6.3 Spectrum spreading due to non-linear amplifier.

$$
\begin{align*}
& \left(\frac{\mathrm{C}}{\mathrm{~N}_{0}}\right)_{\text {up }}=\phi_{s}-\log \left(\frac{4 \pi}{\lambda^{2}}\right)-\mathrm{B}_{\mathrm{i}}-\mathrm{k}+\left(\frac{\mathrm{G}}{\mathrm{~T}}\right)_{\mathrm{s}}-\log \mathrm{log}(6.1) \\
& \left(\frac{C}{N_{o}}\right)_{\text {down }}=E I R P-L_{d}-B_{o}-k+\left(\frac{G}{T}\right)_{E}-10 \log N  \tag{6.2}\\
& \frac{C}{I_{O}} \simeq\left\{\begin{array}{l}
\left(\frac{C}{I}\right)_{C C}+\log \log _{N}+\left[10 \log \left(\frac{\mathrm{BW}}{\mathrm{NBW}}\right)-1.5\right](6.3) \\
\text { for } \leqslant 71 \% \text { bandwidth occupancy } \\
\left(\frac{C}{I}\right)_{C C}+10 \log B W_{N} \\
\text { for } \geqslant 71 \% \text { bandwidth occupancy }
\end{array}\right.
\end{align*}
$$

where
$\phi_{s} \quad$ is the saturating flux density in $d B W / \mathrm{m}^{2}$.
$\lambda$ is the unlink wavelength in metres.
$B_{i}$ is the satellite input backoff in $d B$.
k is Boltzmann's constant $=-228.6 \mathrm{~dB}-\mathrm{K}$.
$\left(\frac{G}{T}\right)_{S}$ is the satellite receive $G / T$ in $d B / K$.
$N$ is the number of homogeneous carriers.

EIRP is the satellite transmitted EIRP in dEW.
$I_{d} \quad$ is the downlink free space path loss in $d B$.
$B_{0} \quad$ is the satellite output backoff.
$\left(\frac{G}{T}\right)_{E}$ is the earth station receive $G / T$ in $d B / K$.
> $\frac{C}{I_{0}} \quad$ is the carrier-to-spectral interference density ratio.

$\left(\frac{C}{I}\right)_{\text {cc }}$ is the centre channel carrier-to-IM noise for equal and equally spaced carriers.
$\mathrm{BW}_{\mathrm{N}}$ is the noise bandwidth of the receiver.
$B W_{R F}$ is the total usable RF bandwidth.
In (6.3), the last term on the right hand side is the IM interference improvement factor for a random selection of SCPC carrier frequencies over the available transponder bandwidth (this is termed the staggering advantage).

The total carrier-to-noise density is then:

$$
\begin{equation*}
\left(\frac{C}{N_{0}}\right)_{\text {total }}=\left[\left(\frac{C}{N_{0}}\right)_{\text {up }}^{-1}+\left(\frac{C}{N_{0}}\right)_{\text {down }}^{-1}+\left(\frac{C}{I_{0}}\right)^{-1}\right]^{-1} \tag{6.4}
\end{equation*}
$$

The link equations (6.1) to (6.4) are used to find the optimum operating point* for the link parameters and for the satellite transponder multi-carrier output power and $C / I$ versus input power transfer curves as shown in Table 6.l and Figure 6.4 [63], respectively. Once the maximum number of channels is found the carrier-to-intermodulation interference ratio can be evaluated according to (6.3).

The optimum input backoff is given as a function of the earth station receive antenna $G / T$ in Table 6.2. The table also shows the carrier-to-noise degradation due to intermodulation interference. It is seen that the $C / N$

[^13]| Satellite Parameters | DBS | ANIK C |  | UNIT |
| :---: | :---: | :---: | :---: | :---: |
| Saturating flux density | -85.0 | -80.7 |  | $\mathrm{dBW} / \mathrm{m}^{2}$ |
| Uplink frequency | 14 | 14 |  | GHz |
| Downlink frequency | 12 | 12 |  | GHz |
| Satellite G/T | $-4.0$ | 1.6 |  | $d B / K$ |
| EIRP | 54.5 | 49.3 |  | dBW |
| Usable RF bandwidth | 27 | 54 |  | MHz |
| Earth Station Parameters* | CATV | HIGH | LOW | UNIT |
| Receiver noise bandwidth |  |  |  |  |
| BPSK, DMSK | 384 | 352 | 288 | kHz |
| QPSK, CMSK | 192 | 176 | 144 | kHz |
| Required Radio $\mathrm{C} / \mathrm{N}_{\mathrm{O}}$ : |  |  |  |  |
| Coherent PSK, MSK | 69.1 | 68.8 | 67. | $\mathrm{dB}-\mathrm{Hz}$ |
| Differential BPSK | 69.7 | 69.4 | 67. | $\mathrm{dB}-\mathrm{Hz}$ |
| Differential MSK | 70.2 | 69.9 | 68. | $\mathrm{dB}-\mathrm{Hz}$ |
| Differential QPSK | 71.4 | 71.1 | 69. | $\mathrm{dB}-\mathrm{Hz}$ |

* See Section 2.l.1. Assuming 2 dB implementation margin for CPSK, CMSK, DPSK modems; 2.5 dB for DMSK modem.

Table 6.l: Parameters for link calculation.

|  | $\begin{aligned} & (G / T)_{E} \\ & (\mathrm{~dB} / \mathrm{K}) \end{aligned}$ | Optimum <br> Input Back- <br> off (dB) | Output Backoff <br> (dB) | C/N Degradation due to $I M$ ( dB ) | Transponder <br> Capacity <br> (Carriers) |
| :---: | :---: | :---: | :---: | :---: | :---: |
| DBS | 4 | 4.3 | 1.9 | 1.3 | 6 |
|  | 6 | 4.9 | 2.0 | 1.7 | 8 |
|  | 8 | 5.6 | 2.2 | 2.2 | 10 |
|  | 10 | 6.7 | 2.5 | 2.4 | 13 |
|  | 12 | 7.5 | 2.9 | 2.5 | 15 |
|  | 14 | 8.0 | 3.0 | 2.6 | 18 |
| ANIK C | 10 | 4.0 | 1.8 | 1.0 | 8 |
|  | 14 | 5.3 | 2.1 | 1.7 | 16 |
|  | 18 | 8.5 | 3.2 | 1.9 | 29 |
|  | 22 | 10.0 | 4.0 | 2.2 | 49 |

Table 6.2: Transponder capacity for CATV quality level with CPSK or CMSK modulation (i.e. for required $C / N_{o}=69.1 \mathrm{~dB}-\mathrm{Hz}$ )
degradation due to IM interference increases with increasing $(G / T)_{E}$ despite the fact that the required input power always decreases (i.e. more back-off). For example, in the DBS case, if $(G / T)_{E}=8 \mathrm{~dB} / \mathrm{K}$ the $\mathrm{C} / \mathrm{N}$ degradation with respect to single carrier operation is 4.4 dB . If the $(G / T)_{E}$ value increases to $12 \mathrm{~dB} / \mathrm{K}$, the degradation increases to 5.4 dB (i.e. an increase of 1.0 dB ) but the system capacity increases to 15 from 10 (i.e. an increase of 1.76 dB ). However, in order to keep the cost of the receive earth station low, the receive antenna $G / T$ value has to be kept small. A typical receiver with a 1 m antenna and $360^{\circ} \mathrm{K}$ noise temperature has a $G / T$ value of about $13.8 \mathrm{~dB} / \mathrm{K}$ which allows about 18 CATV quality radio carriers to be transmitted through a DBS satellite transponder using CPSK or CMSK modulation and SCPC/FDMA (16 radio carriers can be transmitted if ANIK $C$ is used).

Note that the capacity as well as the optimum input backoff and the $C / N$ degradation values given in Table 6.2 are for a required $C / N_{o}=69.1 \mathrm{~dB}-\mathrm{Hz}$ only. However, it was found that the optimum input backoff and the $C / N$ degradation values do not change with the required $C / N_{o}$ for all modulation and quality levels considered in the report. The transponder capacity is plotted against earth station receive antenna $G / T$ for coherent and differential PSK, MSK and for three quality levels in Appendix B.

This analysis applies to multi-carrier operation. In the case where the radio channels are grouped together with each group time division multiplexed to modulate a single carrier, the amount of IM interference is expected to be smaller since there will now be fewer carriers which carry the same composite number of radio channels as in the SCPC case. For 3 or more TDM carriers sharing the same
transponder, the $C / N$ degradation due to $I M$ interference given in Table 6.2 can be used as an upper bound. In the case of 2 wideband TDM carriers sharing the same transponder, IM interference can be avoided by leaving a guard band equal to (at least) the larger carrier bandwidth between the TDM carriers. However, the $2 A-B$ and $2 B-A$ IM products may fall into the adjacent transponder causing interference to the adjacent transponder traffic. The degree of this type of interference depends on the satellite output multiplexer filter. Figure 6.5 [64, Figure 4] plots the $2 A-B$ and $2 B-A$ IM product levels against the carrier difference at the input for several total input backoff values. The figure shows that there is a $d B-f o r-d B$ tradeoff between the input backoff and the IM product levels. At $0 d B$ input backoff and $0 d B$ carrier difference at input, the $2 A-B$ (or $2 \mathrm{~B}-\mathrm{A}$ ) product level is -15 dB . To reduce this IM product level to below say -40 dB in order to avoid interference to an adjacent transponder, an output multiplexer filter with stopband attenuation of at least 25 $d B$ is required. For a power limited system, if the TDM carriers are placed at the extreme ends of the transponder (ensuring the IM products well into the adjacent transponders), this is not a stringent requirement.

It is shown in Appendix $B$ that the DBS and ANIK C systems both operate under a power limited condition in the dedicated transponder case. Thus, there is more bandwidth available than required. Therefore, a specific frequency plan for the SCPC carriers can be employed to further reduce IM interference*.

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Frequency plans for IM-free operation [65, 66] provide frequency arrangements for the carrier centre frequencies such that no $I M$ products will fall within the carrier bandwidths. These plans, however, are quite inefficient in bandwidth utilization. For example, in one case cited 284 consecutive channel slots are required in order to allocate only 20 third-order IM-free carriers, i.e. the total bandwidth required is 14.2 times that occupied by the carriers alone. As a result, these frequency plans are not widely used. Another approach for this problem is called the IM-minimum Channel Allocation [67] which is very attractive from the viewpoint of bandwidth utilization efficiency. This is a computerized iterative approach to minimizing IM products. Table 6.3 [67, Table 4] gives some examples of IM-minimum Channel Allocation and IM-advantage factor. The IM-advantage factor shown in the table ranges from 2.5 dB to 4.0 dB higher than the approximate improvement factor $\left[\log \left(\frac{\mathrm{BW}_{\mathrm{RF}}}{\mathrm{N} . \mathrm{BW}_{\mathrm{N}}}\right)-1.5\right]$ used in (6.3). Note that the bandwidth ratio given in the table is the ratio $\frac{\mathrm{BW}_{R F}}{\mathrm{~N} \cdot \mathrm{BW}_{\mathrm{N}}}$.
6.1.2 Interference in a Shared Transponder - FMTV Plus Radio SCPC Case

This is the case where an FMTV signal and radio SCPC carriers share the same satellite transponder. Two kinds of interference mechanism are considered in this section: direct interference and intermodulation interference.
6.1.2.1 Direct Interference from SCPC to FMTV

Only radio SCPC carriers which are passed by the TV IF receive filter will cause interference to the TV signal.

TABLE 6.3
Optimum Channel Allocations for Single-level SCPC Systems
(a) NUMBER OF CARRIERS $=20$

| $\begin{gathered} \text { BANDWIDTH } \\ \text { RATIO } \end{gathered}$ | AVAILABLE No. OF SLOTS | $\begin{gathered} \text { IM-ADVANTAGE } \\ (\mathrm{OB}) \end{gathered}$ | CHANNEL ALLOCATION |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
| 2 | 40 | 4.45 | 1 | 2 | 3 | 5 | 6 | 7 | 10 | 16 | 17 | 19 | 24 | 26 | 27 | 32 | 34 | 36 | 37 | 38 | 39 | 40 |
| 3 | 60 | 6.73 | 1 | 2 | 3 | 6 | 9 | 12 | 13 | 19 | 22 | 27 | 32 | 41 | 43 | 47 | 54 | 55 | 56 | 58 | 59 | 60 |
| 4 | 80 | 8.18 | 1 | 2 | 3 | 4 | 8 | 11 | 16 | 28 | 30 | 41 | 45 | 52 | 57 | 61 | 62 | 70 | 73 | 76 | 79 | 80 |
| 5 | 100 | 9.54 | 1 | 2 | 3 | 5 | 10 | 17 | 29 | 31 | 42 | 45 | 48 | 65 | 67 | 80 | 85 | 89 | 90 | 96 | 98 | 100 |

(b) NUMBER OF CARRIERS $=30$

| BANDWIDTH <br> RATIO | AVAILABLE NO. OF SLOTS | IM-ADVANTAGE ( CB ) | CHANNEL ALLOCATION |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
| 2 | 60 | 4.28 | 47 | 2 49 | 3 50 | 54 | 55 | 6 56 | 7 57 | 10 58 | 13 59 | $\begin{aligned} & 15 \\ & 60 \end{aligned}$ | 18 | 22 | 26 | 27 | 28 | 32 | 36 | 40 | 42 | 43 |
| 3 | 90 | 6.37 | 70 | $7{ }^{2}$ | 3 79 | $\begin{gathered} 4 \\ 8 i \end{gathered}$ | 6 83 | $8{ }^{9}$ | 15 86 | $\begin{aligned} & 16 \\ & 88 \end{aligned}$ | $\begin{aligned} & 19 \\ & 89 \end{aligned}$ | $\begin{aligned} & 22 \\ & 90 \end{aligned}$ | 28 | 31 | 36 | 39 | 42 | 47 | 59 | 60 | 64 | 68 |
| 4 | 120 | 7.85 | 90 | ${ }_{9}^{2}$ | 102 | 108 | 6 110 | 9 113 | $\begin{array}{r} 13 \\ 117 \end{array}$ | $\begin{array}{r} 17 \\ 118 \end{array}$ | $\begin{array}{r} 19 \\ 119 \end{array}$ | $\begin{array}{r} 32 \\ 120 \end{array}$ | 37 | 39 | 48 | 51 | 56 | 62 | 69 | 76 | 77 | 86 |
| 5 | 150 | 8.95 | 117 | $122^{2}$ | 3 135 | 140 | 10 143 | 14 145 | $\begin{array}{r} 18 \\ 147 \end{array}$ | 20 148 | $\begin{array}{r} 29 \\ 149 \end{array}$ | $\begin{array}{r} 30 \\ 150 \end{array}$ | $40$ | 45 | 52 | 66 | 70 | 73 | 81 | 87 | 104 | 116 |

(c) NUMBEA OF CARRIERS $=40$

| $\begin{aligned} & \text { BANDWIDTH } \\ & \text { RATIO } \end{aligned}$ | AVAILABLE NO. OF SLOTS | IM-A DVANTAGE <br> (dB) | CHANNEL ALLOCATION |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
| 2 | 80 | 4.22 | 43 | 2 | $4{ }^{3}$ | 4 48 | 52 | 66 | 62 | 8 64 | 10 65 | 13 66 | 16 70 | 18 71 | 22 72 | .24 74 | 25 75 | 27 76 | 28 77 | 32 78 | 35 79 | 39 <br> 80 |
| 3 | 120 | 6.24 | 67 | 2 70 | $77^{\frac{3}{3}}$ | 4 78 | 82 | 6 88 | 97 | 888888 | 12 97 | 15 100 | 17 105 | 21 109 | 1111 | 34 112 | 39 113 | 45 116 | 51 117 | 59 118 | 57 119 | 65 120 |
| 4 | 160 | 7.63 | 71 | 2 72 | $7 \stackrel{3}{9}$ | 5 92 | 7 7 | 8 108 | 10 115 | 13 117 | $\begin{array}{r} 20 \\ i 26 \end{array}$ | $\begin{array}{r} 21 \\ 132 \end{array}$ | $\begin{array}{r} 25 \\ 136 \end{array}$ | $\begin{array}{r} 30 \\ 145 \end{array}$ | $\begin{array}{r} 31 \\ 148 \end{array}$ | $\begin{array}{r} 37 \\ 154 \end{array}$ | $\begin{array}{r} 40 \\ 152 \end{array}$ | $\begin{array}{r} 45 \\ 154 \end{array}$ | $\begin{array}{r} 56 \\ 156 \end{array}$ | $\begin{array}{r} 58 \\ +59 \end{array}$ | $\begin{array}{r} 66 \\ 159 \end{array}$ | $\begin{array}{r} 70 \\ 160 \end{array}$ |
| 5 | 200 | 8.74 | 104 | $10{ }^{2}$ | 117 | 128 | 6 132 | 144 | 10 146 | 163 | 20 164 | 26 175 | 32 176 | 38 184 | 45 185 | 54 190 | 58 192 | 71 194 | $\begin{array}{r}75 \\ \hline 95\end{array}$ | $\begin{array}{r} 83 \\ 198 \end{array}$ | $\begin{array}{r} 85 \\ 199 \end{array}$ | $\begin{array}{r} 86 \\ 200 \end{array}$ |


| $\begin{aligned} & \text { BAHDHIDTH } \\ & \text { RATIO } \end{aligned}$ | AVA ILABLE NO. OF SLOTS | $\left\|\begin{array}{c} \text { IM-ADVANTA GE } \\ (\mathrm{dB}) \end{array}\right\|$ | CHANNEL ALLOCATION |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
| ? | 100 | 4.13 | $\begin{array}{ll}1 & 2 \\ 38 & 41 \\ 90 & 91\end{array}$ | 3 42 92 | 4 44 93 | 5 46 94 | 6 53 96 | 7 56 97 | 8 58 98 | 9 62 99 | $\begin{array}{r} 10 \\ 64 \\ 100 \end{array}$ | 16 67 | 17 68 | 18 73 | 19 75 | 22 77 | 24 81 | 25 85 | 28 86 | 31 88 | 36 89 |
| 3 | 150 | 6.15 | $\begin{array}{rr} 1 & 2 \\ 56 & 60 \\ 138 & 139 \end{array}$ | $\begin{array}{r} 3 \\ 64 \\ 141 \end{array}$ | $\begin{array}{r} 4 \\ 66 \\ 142 \end{array}$ | $\begin{array}{r} 5 \\ 76 \\ 144 \end{array}$ | $\begin{array}{r} 6 \\ 77 \\ 146 \\ \hline \end{array}$ | $\begin{array}{r} 7 \\ 85 \\ 147 \end{array}$ | $\begin{array}{r} 8 \\ 90 \\ 148 \end{array}$ | $\begin{array}{r} 9 \\ 96 \\ 149 \end{array}$ | $\begin{array}{r} 12 \\ 98 \\ 150 \\ \hline \end{array}$ | $\begin{aligned} & 20 \\ & 99 \end{aligned}$ | $\begin{array}{r} 21 \\ 105 \end{array}$ | 24 109 | 29 113 | 33 118 | 36 123 | 38 128 | $\begin{array}{r} 43 \\ 131 \end{array}$ | $\begin{array}{r} 46 \\ 134 \end{array}$ | 48 135 |
| 4 | 200 | 7.57 | $\begin{array}{rr} 1 & 2 \\ 76 & 78 \\ 184 & 185 \end{array}$ | $\begin{array}{r} 27 \\ 188 \end{array}$ | $\begin{array}{r} 4 \\ 92 \\ 192 \\ \hline \end{array}$ | $\begin{array}{r} 5 \\ 98 \\ 195 \\ \hline \end{array}$ | $\begin{array}{r} 6 \\ 106 \\ 106 \\ \hline \end{array}$ | $\begin{array}{r} 10 \\ 110 \\ 197 \end{array}$ | $\begin{array}{r} 11 \\ 116 \\ 198 \\ \hline \end{array}$ | $\begin{array}{r} 14 \\ 120 \\ 199 \\ \hline \end{array}$ | $\begin{array}{r} 15 \\ 136 \\ 200 \\ \hline \end{array}$ | $\begin{array}{r} 20 \\ 140 \end{array}$ | $\begin{array}{r}27 \\ 142 \\ \hline\end{array}$ | $\begin{array}{r}33 \\ 43 \\ \hline\end{array}$ | 35 145 | 40 157 | 45 164 | 51 171 | 52 173 | 60 179 | 74 180 |
| 5 | 250 | 8.59 | $\begin{array}{rr} 1 & 2 \\ 81 & 91 \\ 23 . & 237 \end{array}$ | $\begin{array}{r} 3 \\ 106 \\ 240 \end{array}$ | $\begin{array}{r} 4 \\ 111 \\ 24 ? \end{array}$ | $\begin{array}{r} 5 \\ 129 \\ 244 \end{array}$ | $\begin{array}{r} 6 \\ 131 \\ 245 \end{array}$ | $\begin{array}{r} 10 \\ 139 \\ 247 \end{array}$ | $\begin{array}{r} 11 \\ 150 \\ 248 \end{array}$ | $\begin{array}{r} 18 \\ 15 ? \\ 249 \end{array}$ | $\begin{array}{r} 20 \\ 156 \\ 250 \\ \hline \end{array}$ | 24 165 | 26 171 | 183 | 42 191 | 47 197 | 57 203 | 61 208 | 72 214 | 73 216 | 77 226 |

the noise bandwidth of the $T V$ receiver is 18 MHz , any SCPC carrier located less than 9 MHz away from the $T V$ carrier is a potential interferor. The SCPC interferor will appear at the output of the $T V$ demodulator as non-stationary interference as a result of the non-linear nature of broadband frequency modulation. That is, if the TV carrier remains stationary for some period of time near the SCPC carrier, interference which falls within the $T V$ baseband frequency range will result. However, if the separation between the instantaneous TV carrier and the SCPC carrier is sufficiently large, the resulting interference to the video will fall outside the $T V$ baseband frequency range and its effect will be negligible.

An approximate analysis of this interference mechanism was performed in detail in [68]. The analysis consists of
(i) calculating the conditional degradation of (weighted and de-emphasized) baseband (video and associated audio) SNR as a function of the instantaneous TV carrier frequency deviation, i.e. the SNR degradation for a given frequency deviation;
(ii) de-conditioning the degradation with respect to the frequency deviation distribution to obtain the first order statistics of the "instantaneous" degradation, i.e. fraction of time (or probability) that degradation of $\ell d B$ is exceeded as a function of $\ell$; and
(iii) trading the degradation level against the probability of occurance according to a reasonable rule $\left(k / \ell^{2}\right.$ bound) to combine the results into an equivalent stationary degradation.

This analysis is performed with the assumption that both the video and audio demodulators operate above threshold. The process is illustrated in Figure 6.6. Numerical evaluation for this interference problem is not performed here since it is very unlikely that the SCPC carriers will fall within the passband of the TV receive filter (see Section 6.1.2.2). As a result, there is no direct interference from SCPC to FMTV.
6.1.2.2 Direct Interference from FMTV to SCPC

The RF spectrum corresponding to a stationary picture such as a colour bar pattern consists of discrete components due to the line repetitive nature of the signal. Hence, interference in any narrow band consisting of several lines will appear stationary and noise-like, and will depend only on the interfering spectrum in the band of interest. Thus, in order to evaluate the interference from FMTV to the radio SCPC carriers, it is necessary to know the RF spectrum of the TV signal. Figure 6.7a shows a plot of the RF power spectrum of an NTSC TV signal generated by a simulation program (see Appendix C) for a TV signal represented by 3 sinusoidal tones. The figure also shows the approximate trapezoidal spectrum shape derived from the calculated one. The mathematical representation of the approximated spectrum normalized to peak value is given by

$$
S_{n}(f)=\left[\begin{array}{lr}
400 e^{.95133 f ;} ; & f \leqslant-6.3 \mathrm{MHz}  \tag{6.5}\\
1.0 & |f| \leqslant 6.3 \mathrm{MHz} \\
400 e^{-.95133 f ;} & f \geqslant 6.3 \mathrm{MHz}
\end{array}\right.
$$


(a)
(b)


Figure 6.6: video SNR degradation statistics
(a) conditional degradation $L(f i)$ as a function of the TV carrier frequency deviation $6 i$.
(b) Commulative distribution function $F(f i)$. of the frequency deviation of the TV signal.
(c) De-condi tioned probability Pd that $\ell$ is exceeded and the equivalent stationary degradation leq $=\mathrm{P}_{\mathrm{t}} \ell_{t}$.

$-175472$


Figure 6.7.: (a) Calculated TV power spectrum with peak deviation $=6 \mathrm{MHz}$. Spacing between 2 consecutive points in the plot-vertical: 5.24 dB ; horizontal: 1.26 MHz . The approximate spectrum is shown by the dotted line.
(b) arrangement for FMTV and radio SCPC carriers.
$S_{n}(f)$ is the normalized power density given in absolute units (not in $d B$ ) and $f$, in $M H z$, is the frequency relative to the centre frequency of the carrier. The actual power density spectrum is

$$
S(f)=\frac{C_{V}}{\int S_{n}(f) d f} S_{n}(f)=\frac{C_{V}}{14.7} S_{n}(f)
$$

where $C_{V}$ is the total power of the $T V$ signal.

From Figure 6.7b, the total interference power is

$$
\begin{equation*}
I_{V}=\int_{D_{f}-\frac{B}{2}}^{D_{f}+\frac{B}{2}} S(f) d f=\frac{C_{V}}{14 \cdot 7} \int_{f}+\frac{B}{2} S_{n}(f) d f \tag{6.7}
\end{equation*}
$$

where $D_{f}$ is the separation between the TV and SCPC centre frequencies and $B$ is the noise bandwidth of the radio receiver.

If we let

$$
\phi(f)=\int_{-\infty}^{f} S_{n}(f) d f
$$

then (6.7) becomes

$$
\begin{equation*}
I_{V}=\frac{C_{V}}{14 \cdot 7}\left[\phi\left(D_{f}+\frac{B}{2}\right)-\phi\left(D_{f}-\frac{B}{2}\right)\right] \tag{6.9}
\end{equation*}
$$

where


Thus the radio carrier-to-interference ratio is

$$
\begin{align*}
\frac{C_{R}}{I_{V}} & =\frac{C_{R}}{\frac{C_{V}}{14.7}\left[\phi\left(D_{f}+\frac{B}{2}\right)-f\left(D_{f}-\frac{B}{2}\right)\right]} \\
& =\frac{14.7 \beta}{\phi\left(D_{f}+\frac{B}{2}\right)-\phi\left(D_{f}-\frac{B}{2}\right)} \tag{6.11}
\end{align*}
$$

where $\beta=\frac{C_{R}}{C_{V}}$ is the received SCPC-to-TV power ratio.
Equation (6.11) is evaluated for three quality levels and PSK, MSK modulations using the radio receiver noise bandwidth and SCPC-to-TV power ratios as shown in Table 6.4. The SCPC-to-TV power ratio is derived from the analysis given in Section 2.l.l. The results are shown in Figures 6.8 to 6.10 as plots of radio carrier-to-interference ratio versus separation between $T V$ and SCPC centre frequencies. From the figures it is seen that coherent BPSK has the lowest C/I. Differential QPSK gives the highest C/I since DQPSK requires the most receive power but the least receive bandwidth compared to other modulations. The figures also show that the radio carrier must be placed at least 13 MHz away from the TV carrier in order to obtain a C/I of at least 25 dB. For differential QPSK, this frequency spacing gives a $C / I$ of about 30 dB .

|  |  | QUALITY LEVEL |  |  |
| :---: | :---: | :---: | :---: | :---: |
|  | Modulation | CATV | High | Low |
| ```SCPC-to-TV power ratio \beta (dB)``` | CPSK, CMSK <br> DBPSK <br> DMSK <br> DQPSK | $\begin{aligned} & -17.9 \\ & -17.3 \\ & -16.8 \\ & -15.6 \end{aligned}$ | $\begin{aligned} & -18.2 \\ & -17.6 \\ & -17.1 \\ & -15.9 \end{aligned}$ | $\begin{aligned} & -19.9 \\ & -19.2 \\ & -18.7 \\ & -17.6 \end{aligned}$ |
| Receiver Noise bandwidth B ( kHz ) | $\begin{array}{ll}\text { BPSK, DMSK } \\ \text { QPSK, } & \text { CMSK }\end{array}$ | $\begin{aligned} & 384 \\ & 192 \end{aligned}$ | $\begin{aligned} & 352 \\ & 176 \end{aligned}$ | $\begin{aligned} & 288 \\ & 1.44 \end{aligned}$ |

Table 6.4: SCPC-to-TV power ratio and receiver noise bandwidth for 3 quality levels.




Table 6.5 gives the Eb/No degradation at $B E R=10^{-6}$ for a low quality radio signal located 13 MHz away from the $T V$ centre frequency assuming that the $T V$ interference is similar to that of a tone interferor for the same $C / I$. Low quality radio was used for the calculation of $E_{b} / N_{0}$ degradation since it gives the lowest $C / I$ as compared with the other two quality levels. Thus, the $\mathrm{E}_{\mathrm{b}} / \mathrm{N}_{\mathrm{o}}$ degradations for CATV and high quality levels are bounded by those given in Table 6.5. Note that even though QPSK results in a high $C / I$ for a given frequency spacing, its BER performance is degraded the most.

### 6.1.2.3 Interference Due to Intermodulation

Since the FMTV and the radio SCPC carriers share the same transponder, power sharing and interference due to intermodulation could be factors that limit the capacity of the system. In this section, both factors are evaluated for $T V$ and radio signals. $T V$ and radio carrier output powers are plotted against the total input power in Figures 6.11 and 6.12 , respectively. These results were obtained by a computer program [55]. The TV and radio output powers versus input power are to be used in the link calculations to compute the required earth station receive antenna $G / T$, the TV carrier-to-noise degradation due to intermodulation interference and the available carrier power for the radio signal(s). Note that the small signal suppression phenomenon is shown in Figure 6.12 where the radio output power decreases as the transponder is driven into saturation.

Figure 6.13 shows the intermodulation products generated by the $T V$ and radio carriers [63]. The levels of significant


Table 6.5: $E_{b} / N_{o}$ degradation at $B E R=10^{-6}$ due to tone interference for low quality radio located 13 MHz away from TV centre frequency.

| $\because$ | - | - | - |  | - |  | - |  |  | $\square$ | $\underline{-7}$ | - | - |  |  | - | - | + |  |  |  |  | - | +i |  |  |  |  |  |  | $\square$ | $\cdots$ | $\square$ |  |  |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
| C...: | $\cdots$ | -1. |  |  | + |  |  | - | $\cdots$ |  | -18 | $\cdots$ | + + | ... |  | $\square$ | $\cdots$ |  | - |  | L- |  | - | -1. | $\ddagger$ | + |  |  |  | --- | - | $\rightarrow$ |  | +-.. |  |
|  |  | $\cdots$ |  |  |  |  | - |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |
| $\cdots$ | 7:-1. | $\square$ | -- | $+$ | - |  | - |  | - |  |  |  |  |  |  |  |  |  | + |  |  | - |  |  |  |  |  |  | - |  |  |  | - |  |  |
| - ..... |  | $\cdots$ | ....- |  | - |  | - | - | - |  |  |  | 6- | - |  | 1 --0 | Ut | Ott | D0 | 440 | r | Ven | St |  | - | , |  |  | + |  | - | $\pm$ | $\square$ | - |  |
|  | -- |  |  |  |  |  | - |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  | T- | - |  | $\cdots$ | P- | - |  |  |  |  |  |  |



|  | $\pm$ |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  | + | + |  |  |  |  |  |  |  |  |  |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
| $\cdots$ |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |
| - | $\underline{\square}$ | $\bigcirc$ |  | Eigu | ne | -6. | 1 | 14- |  | $V=$ | carr | nrie | er | -to | -1:MEin | inte | terif | repre | renc | c- | - | , |  |  |  |  |  |  |  |  |  |
| - | $\cdots$ |  |  |  |  |  |  |  |  |  |  |  |  |  | - |  |  |  |  | - | , |  |  |  | - | , | - | - | - |  |  |
|  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |
| $\pm$ |  | - |  |  |  |  |  | $\cdots$ |  | ati | 10 | Ver | rsu | us =to | total | 71 In | inpa |  | pow | wer | r |  |  |  |  |  |  |  |  |  |  |
|  |  |  |  | - |  |  |  | $\cdots$ |  | $\cdots$ | $\cdots$ | $\cdots$ |  |  | - |  | $\cdots$ | - | $\pm$ |  |  |  | $\cdots$ | - | $\square$ |  | - |  | - | - |  |
|  | - |  |  | . |  |  |  | + |  | con | mpu | ute | ed | 1-0 | For An | Anill | i-k | A | tyi | pe |  | -a | anis | Spro | Ond | dex | C- |  |  |  |  |
|  |  |  |  | - |  |  |  |  |  |  | - |  |  |  |  |  |  |  |  |  |  |  |  | - |  |  |  |  |  |  |  |
| $\pm$ | $\square$ | - |  |  |  |  |  | - | - | , |  |  |  |  | - | - | - | $\cdots$ | - | - | - | - | $\bigcirc$ | $\geq$ | - |  | - | - | $\underline{-}$ |  |  |
|  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |
|  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |

Th eariveroly interference ratio
-

| $\underline{0}$ | I | - |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  | - |  |  | - |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
| $\cdots$ | $\underline{+}$ | - | H | - | $\cdots$ | - | - | $\bigcirc$ |  | - |  | - | - | - | , | $\cdots$ |  | - |  |  |  | , |  |
| - | - | - | + | - | - | - | $\bigcirc$ | - |  | - |  |  |  | - |  | - |  | - |  |  |  | - |  |
|  | $\underline{-1}$ | - | - | + | $\cdots$ | - | - | $\underline{+}$ |  | - | - | - | - | - |  |  |  | - | - | $\cdots$ |  | - |  |


| $2 f v-f r$ | $-22 d B$ |
| :--- | ---: |
| $2 f r-f v$ | -42 dB |
| $3 f v-2 f r$ | -30 dB |
| $3 f r-2 f r$ | -50 dB |



Figure 6.13 Multiple radio program and TV carrier within an RF channel

IM products are also given in the figure for a received radio/TV power ratio of -20 dB . The figure suggests the following points:
(i) Radio carriers must be grouped to one side of the TV carrier since the power levels of the $2 f_{v}-f_{r} I M$ products are comparable to those of the radio carriers.
(ii) A guard band between the radio and TV carriers should be used in order to avoid the $f_{v}+f_{r_{1}}-f_{r_{2}}$ type IM products falling into the radio program bandwidth. This guard band should be at least equal to the bandwidth occupied by the radio programs.
(iii) The $2 f_{r_{1}}-f_{r_{2}}$ and $f_{r_{1}}+f_{r_{2}}-f_{r_{3}}$ IM products which are within the radio program bandwidth are small and can be neglected since the radio carriers are well backed-off.

Hence, it is seen that intermodulation interference is negligible for the radio program carriers if sufficient guard band between radio and $T V$ is used and the radio carriers are grouped to one side of the TV carrier.

The link analysis for the shared transponder-SCPC case is as follows.

The link equations for the TV program carrier are:

$$
\begin{equation*}
\left(\frac{C_{V}}{\mathrm{~N}_{\mathrm{O}}}\right)_{u p}=\phi_{s}-\log \left(\frac{4 \pi}{\lambda^{2}}\right)-k+(G / \mathrm{T})_{s}+P_{V} \tag{6.12}
\end{equation*}
$$

$$
\begin{align*}
& \left(\frac{C_{V}}{N_{o}}\right)_{\text {down }}=E I R P-L_{d}-k+(G / T)_{E}+P^{\prime} V  \tag{6.13}\\
& \left(\frac{C_{V}}{I_{O}}\right)=\left(\frac{C_{V}}{I}\right)+10 \log (18 \mathrm{MHz}) \tag{6.14}
\end{align*}
$$

where $P_{V}$ and $P^{\prime} V_{V}$ are the $T V$ powers (relative to single carrier saturation) at the input and output of the transponder, respectively. The value for $C_{V} / I$ is obtained from Figure 6.14. The available $T V\left(\frac{C}{N_{O}}\right)$ is then given by

$$
\left(\frac{C_{V}}{N_{0}}\right)_{\text {available }}=\left[\left(\frac{C_{V}}{N_{0}}\right)_{\text {up }}^{-1}+\left(\frac{C_{V}}{N_{0}}\right)_{\text {down }}^{-1}+\left(\frac{C_{V}}{I_{0}}\right)^{-1}\right]^{-1}(6.15)
$$

Similarly, the link equations for the radio program carrier are:

$$
\begin{align*}
& \left(\frac{C_{R}}{N_{O}}\right)_{\text {up }}=\phi_{S}-\operatorname{lolog}\left(\frac{4 \pi}{\lambda^{2}}\right)-k+(G / T)_{S}+P_{R}  \tag{6.16}\\
& \left(\frac{C_{R}}{N_{O}}\right)_{\text {down }}=E I R P-L_{d}-k+(G / T)_{E}+P_{R}^{\prime}  \tag{6.17}\\
& \left(\frac{C_{R}}{I_{O}}\right) \text { is very large. }
\end{align*}
$$

where $P_{R}$ and $P^{\prime}{ }_{R}$ are the radio power (relative to single carrier saturation) at the input and output of the transponder, respectively. The available carrier-to-noise density ratio for the radio program is given by

$$
\begin{equation*}
\left(\frac{C_{R}}{N_{0}}\right) \text { available }=\left[\left(\frac{C_{R}}{N_{0}}\right)_{\text {up }}^{-1}+\left(\frac{C_{R}}{N_{0}}\right)_{\text {down }}^{-1}\right]^{-1} \tag{6.18}
\end{equation*}
$$

Using (6.12) to (6.15), the earth station receive antenna G/T required for $\left(\frac{C_{V}}{N_{0}}\right)$ available $=\left(\frac{C_{V}}{N_{0}}\right)_{\text {required }}=$ $87.0 \mathrm{~dB}-\mathrm{Hz}$ is calculated for the link parameters given in Table 6.1 and shown in Tables 6.6 and 6.7 for DBS and ANIK $C$ cases, respectively. The tables also show the $T V$ carrier-to-noise ratio degradation due to IM interference and the total carrier-to-noise density ratio available for the radio program carriers which is calculated by using (6.16) to (6.18). It is seen from the tables that the transponder should be operated near saturation since this operating point almost always requires the smallest $(G / T)_{E}$ except for the case where there is a large number of SCPC carriers sharing the same transponder with FMTV*. Backingoff the transponder means that the TV carrier-to-noise ratio degradation due to $I M$ is decreased but the required $(G / T)_{E}$ increases and hence the available radio carrier-tonoise density ratio also increases. However, the trade-off between $(G / T)_{E}$ and the degradation or the available radio carrier-to-noise density ratio is not $d B$ for $d B$. For example, in Table 6.7, for an input radio/TV power ratio of $-12 d B$, in changing the input back-off from $2 d B$ to $6 d B$ (i.e. changing the total input power from -2 dB to -6 dB ) the required $(G / T)_{E}$ increases by 1.3 dB but the degradation decreases by only 0.32 dB and the available radio $\mathrm{C} / \mathrm{N}_{\mathrm{o}}$ increases by only 0.9 dB .

[^15]| Total <br> Radio/TV <br> Power Ratio <br> On The <br> Uplink <br> (dB) | Total Input Power (dB) | TV C/No $\mathrm{N}_{\mathrm{o}}$ de gradation due to IM ( dB ) | $\begin{aligned} & \text { Req' } \mathrm{d}(\mathrm{G} / \mathrm{T})_{\mathrm{E}} \\ & \text { for } \mathrm{TV} \mathrm{C} / \mathrm{N}_{\mathrm{O}} \\ & =87 \mathrm{~dB}-\mathrm{Hz} \end{aligned}$ | Total available Radio C/No ( $\mathrm{dB}-\mathrm{Hz}$ ) |
| :---: | :---: | :---: | :---: | :---: |
| -18 | $\begin{aligned} & -2 \\ & -4 \\ & -6 \end{aligned}$ | $\begin{aligned} & 0.19 \\ & 0.15 \\ & 0.10 \end{aligned}$ | $\begin{aligned} & 12.61 \\ & 13.97 \\ & 16.72 \end{aligned}$ | $\begin{aligned} & 66.32 \\ & 66.78 \\ & 67.92 \end{aligned}$ |
| -16 | $\begin{aligned} & -2 \\ & -4 \\ & -6 \end{aligned}$ | $\begin{aligned} & 0.28 \\ & 0.24 \\ & 0.15 \end{aligned}$ | $\begin{aligned} & 12.81 \\ & 14.37 \\ & 17.11 \end{aligned}$ | $\begin{aligned} & 68.43 \\ & 69.65 \\ & 70.49 \end{aligned}$ |
| -14 | $\begin{aligned} & -2 \\ & -4 \\ & -6 \end{aligned}$ | $\begin{aligned} & 0.43 \\ & 0.36 \\ & 0.23 \end{aligned}$ | $\begin{aligned} & 13.14 \\ & 14.74 \\ & 17.60 \end{aligned}$ | $\begin{aligned} & 70.49 \\ & 71.63 \\ & 72.45 \end{aligned}$ |
| -12 | $\begin{aligned} & -2 \\ & -4 \\ & -6 \end{aligned}$ | $\begin{aligned} & 0.67 \\ & 0.53 \\ & 0.35 \end{aligned}$ | $\begin{aligned} & 13.64 \\ & 15.11 \\ & 18.49 \end{aligned}$ | $\begin{aligned} & 73.07 \\ & 73.92 \\ & 74.75 \end{aligned}$ |

Table 6.6: Required earth station antenna G/T, TV (C/No) degradation due to intermodulation and total available radio $C / N_{o}$ as functions of total input power and the Radio/TV power ratio at the input of the transponder for DBS case.

| Total <br> Radio/TV <br> Power Ratio <br> On The Uplink <br> (dB) | Total Input Power (dB) | TV C/No degradation due to IM (dB) | $\left\|\begin{array}{l} \text { Req' } \mathrm{d}(\mathrm{G} / \mathrm{T})_{\mathrm{E}} \\ \text { for } \mathrm{TV} \mathrm{C/N} \\ =87 \mathrm{~dB}-\mathrm{Hz} \end{array}\right\|$ | Total available Radio $\mathrm{C} / \mathrm{N}_{\mathrm{o}}$ $(\mathrm{dB}-\mathrm{Hz})$ |
| :---: | :---: | :---: | :---: | :---: |
| -14 | $\begin{aligned} & -2 \\ & -4 \\ & -6 \end{aligned}$ | $\begin{aligned} & 0.44 \\ & 0.36 \\ & 0.23 \end{aligned}$ | $\begin{aligned} & 17.13 \\ & 17.68 \\ & 18.50 \end{aligned}$ | $\begin{aligned} & 69.88 \\ & 70.72 \\ & 71.40 \end{aligned}$ |
| -12 | $\begin{aligned} & -2 \\ & -4 \\ & -6 \end{aligned}$ | $\begin{aligned} & 0.67 \\ & 0.54 \\ & 0.35 \end{aligned}$ | $\begin{aligned} & 17.51 \\ & 18.01 \\ & 18.81 \end{aligned}$ | $\begin{aligned} & 72.75 \\ & 73.14 \\ & 73.65 \end{aligned}$ |
| -10 | $\begin{aligned} & -2 \\ & -4 \\ & -6 \end{aligned}$ | $\begin{aligned} & 1.03 \\ & 0.80 \\ & 0.54 \end{aligned}$ | $\begin{aligned} & 18.01 \\ & 18.35 \\ & 19.23 \end{aligned}$ | $\begin{aligned} & 74.94 \\ & 75.37 \\ & 75.86 \end{aligned}$ |
| -8 | $\begin{aligned} & -2 \\ & -4 \\ & -6 \end{aligned}$ | $\begin{aligned} & 1.74 \\ & 1.27 \\ & 0.82 \end{aligned}$ | $\begin{aligned} & 19.23 \\ & 19.23 \\ & 19.75 \end{aligned}$ | $\begin{aligned} & 78.09 \\ & 78.07 \\ & 78.14 \end{aligned}$ |
| -6 | $\begin{aligned} & -2 \\ & -4 \\ & -6 \end{aligned}$ | $\begin{aligned} & 2.89 \\ & 2.06 \\ & 1.33 \end{aligned}$ | $\begin{aligned} & 20.83 \\ & 20.54 \\ & 20.83 \end{aligned}$ | $\begin{aligned} & 81.72 \\ & 81.28 \\ & 81.00 \end{aligned}$ |

Table 6.7: Required earth station antenna G/T, TV (C/No) degradation due to intermodulation and total available radio $C / N_{o}$ as functions of total input power and the Radio/TV power ratio at the input of the transponder for ANIK $C$ case.

Note that the $C / N_{o}$ degradations due to $I M$ shown in Tables 6.6 and 6.7 may translate into somewhat larger video SNR degradations. This is because the dominant interfering IM product, of the form $f_{v}+f_{r_{1}}-f_{r_{2}}$ (see Figure 6.13), is coherent with the $T V$ carrier and hence will be demonstrated into a narrowband interferer centred at $f_{r_{1}}-f_{r_{2}}$. The $I M$ interference will therefore receive less $F M$ plus noise weighting advantage than will noise. For a small number of radio program carriers at sufficiently high level, it may also be subjectively disturbing. This will to some extent mitigate the above comments concerning optimum backoff.

The number of radio channels supported by the available values of radio $\left(C / N_{o}\right)$ shown in Tables 6.6 and 6.7 for input backoff $=2 \mathrm{~dB}$ is given in Appendix $B$ for three quality levels and PSK, MSK modulations.
6.1.3 Interference in a Shared Transponder - Radio Subcarrier Case

In the subcarrier case, the radio signals are time division multiplexed to modulate a subcarrier located above the TV baseband spectrum. Direct interference and interference due to intermodulation will be discussed in this section.
6.1.3.1 Direct Interference to Baseband TV from the Radio
Subcarrier

The direct interference level depends mostly on how sharply the $T V$ and radio signals are filtered. An approximation for the direct interference level is given below. The following assumptions apply to the discussion:

- TV audio and radio program subcarriers arranged as shown in Figure 5.16
- Peak composite frequency deviation $\left(\Delta f_{v}\right)=6 \mathrm{MHz}$
- Video baseband bandwidth $\left(\mathrm{f}_{\mathrm{v}}\right)=4.2 \mathrm{MHz}$
- TV receiver IF noise bandwidth $=18 \mathrm{MHz}$
- Required TV Carrier-to-noise density ratio $=87.0$ dB-Hz
- Required radio carrier-to-noise density ratio $=69.1$ $+\operatorname{lolog} \mathrm{N} \mathrm{dB}-\mathrm{Hz}$, where N is the number of multiplexed radio channels.

The video signal-to-noise density ratio (unweighted) with de-emphasis advantage $P=2.5 d B[70]$ is given by

$$
\begin{align*}
& \frac{S_{V}}{N_{O}}=6\left(\frac{C_{V}}{N_{o}}\right) \frac{\Delta f_{V}^{2}}{\mathrm{E}_{V}} P \\
& =6\left(10^{8.7}\right)\left(\frac{6.0 \times 10^{6}}{4.2 \times 10^{6}}\right)^{2}\left(10^{0} 25\right) \\
& =1.091 \times 10^{10} \\
& =100.38 \mathrm{~dB}-\mathrm{Hz} \tag{6.19}
\end{align*}
$$

This corresponds to an unweighted signal-to-noise ratio $S_{V} / N=34.1 \mathrm{~dB}$.

Suppose that the TV low-pass filter rejects $R d B$ of the radio subcarrier power, then the (unweighted) TV signal-toradio interference ratio is

$$
\begin{align*}
& \frac{S_{V}}{I}=\frac{S_{V} / N_{O}}{S_{R} / N_{O}}+R=100.38-(69.1+10 \log N)+R \\
& \frac{S_{V}}{I}=31.28+R-10 \log N d B \tag{6.20}
\end{align*}
$$

If the radio subcarrier is narrowband, the value of $R$ can be estimated by the attenuation of the TV low-pass filter at the subcarrier centre frequency. For example, for a 4 thorder, 0.5 dB ripple Chebyshev low-pass filter with cut-off frequency of 4.2 MHz , the attenuation at 6 MHz (which is the assumed radio subcarrier centre frequency) is about 20.3 dB . Thus, the TV signal-to-radio interference ratio for $N=1$ is $S_{V} / I_{R}=51.58 d B$ which degrades the $T V S / N$ by an amount of $\log \log \left(1+\frac{\mathrm{S}_{\mathrm{V}} / \mathrm{N}}{\mathrm{S}_{\mathrm{V}} / \mathrm{I}_{\mathrm{R}}}\right)=0.08 \mathrm{~dB}$. For $\mathrm{N}=2$ and. 4 , the degradation increases to 0.15 dB and .30 dB , respectively. Therefore it is seen that the radio subcarrier at 6 MHz has small direct interference on the video signal. However, it may cause significant interference to the audio subcarrier since the two subcarriers would be separated by only $l \mathrm{MHz}$. As a result, a higher order filter and/or a larger spacing between the audio and radio subcarriers should be used to reduce direct interference from radio to TV audio. It should be noted that a higher order filter will likely cause more signal distortion and a higher radio subcarrier frequency requires more RF power and bandwidth. Further study is necessary to find the optimum tradeoff between the factors mentioned above.
6.1.3.2 Interference to Baseband TV Due to Intermodulation

Another problem with the use of a radio subcarrier is the problem of intermodulation between the dominant baseband line components (colour, audio and radio subcarriers) resulting from distortion of the FM carrier. The demodulated video signal-to-IM interference is calculated in this section by using a simulation program (see Appendix C). The simulated baseband $T V$ signal is modelled by 3 randomphase sinusoids having frequencies $15 \mathrm{kHz}, 3.58 \mathrm{MHz}$ and 5 MHz which represent the luminance, chrominance and the associated audio subcarriers, respectively. A fourth sinewave located at 6 MHz is used to simulate the radio
subcarrier*. The simulated composite baseband signal is pre-emphasized, and the result $F M$ modulates the RF carrier with a peak frequency deviation of 6 MHz . The modulated RF

* more accurate model for the composite (TV + Radio) signal is

3

$$
x(t)=\sum_{i=1} A_{i} \cos \left(\omega_{i} t+\phi_{i}\right)+R(t)
$$

The first term on the right hand side is the 3 tone TV signal. The radio signal $R(t)$ is given by:

$$
\begin{aligned}
R(t)= & \sum_{n} a_{n} g(t-n T) \cos \left(\omega_{c} t+\phi\right) \\
& +\sum_{n} b_{n} g(t-n T) \sin \left(\omega_{c} t+\phi\right) \text { for QPSK } \\
\text { or } R(t)= & \sum_{n} a_{n} g(t-n T) \cos \left(\omega_{c} t+\phi\right) \\
& +\sum_{n} b_{n} g\left(t-n T+\frac{T}{2}\right) \sin \left(\omega_{c} t+\phi\right) \text { for MSK }
\end{aligned}
$$

where

$$
a_{n}, b_{n} \text { are } p n \text { sequences, }
$$

$$
T \text { is the symbol interval, }
$$

${ }^{\omega} \mathrm{c}$ is the radio subcarrier frequency,

$$
g(t)=\left[\begin{array}{l}
\text { square-root raised-cosine pulse for PSK } \\
\text { modulation, or } \\
\cos \left(\frac{2 \pi t}{T}\right) ;-\frac{T}{2} \leqslant t \leqslant \frac{T}{2} \text { for MSK modulation }
\end{array}\right.
$$

carrier is then filtered by a 4 th-order, 0.5 dB ripple Chebyshev bandpass filter with $3-d B$ bandwidth of 18 MHz . Nonlinear $A M / P M$ conversion of $5^{\circ} / \mathrm{dB}$ is simulated after the filter. The result after the nonlinearity is then demodulated by an ideal $F M$ demodulator. Finally, deemphasis is applied to the signal. The video signal-toweighted intermodulation noise ratio is calculated by taking the ratio of the power of the luminance and chrominance subcarriers to that of the intermodulation products lying within the 4.2 MHz video noise bandwidth. The process is illustrated in Figure 6.15.

Table 6.8 shows the levels of the demodulated de-emphasized signal and those of the second and third order IM products within the 4.2 MHz band for 0,1 , and 4 TDM'd CATV-quality radio channels. The video signal-to-weighted IM noise is also shown in the table. The degradation due to IM interference for the video signal can be estimated by converting the IM power to equivalent thermal noise. By using this approach, the degradation for the video signal with weighted $\mathrm{S} / \mathrm{N}=44.6 \mathrm{~dB}$ is $0.20 \mathrm{~dB}, 0.74 \mathrm{~dB}$ and 1.87 dB for $N=0$, 1 , and 4 TDM'd radio channels, respectively. Recall that the degradation due to direct interference to video from radio is 0.08 dB and 0.30 dB for $\mathrm{N}=1$ and 4 , respectively. The TV associated audio may be degraded by direct interference from the radio carrier. However, it is not subject to. IM interference. A radio subcarrier frequency higher than 6 MHz may be needed to avoid excessive direct interference to the audio signal.

[^16]

Figure 6.15: Simulation. Block Diagram used to Calculate the Video Signal-toweighted IM noise ratio.

| Description of Tones | Frequency (MHz) | Noise Weighting ( dB ) | Power of Tone in $d B$ for $N$ tDM'd Radio Channels |  |  |
| :---: | :---: | :---: | :---: | :---: | :---: |
|  |  |  | $\mathrm{N}=0$ | $\mathrm{N}=1$ | $\mathrm{N}=4$ |
| Luminance <br> Chrominance A <br> Audio <br> B <br> Radio (CATV)C | $\begin{aligned} & 0.015 \\ & 3.58 \\ & 5.00 \\ & 6.00 \end{aligned}$ | Not applicable | $\begin{gathered} 0.0 \\ -6.92 \\ -22.50 \end{gathered}$ | $\begin{gathered} 0.0 \\ -6.93 \\ -22.38 \\ -17.89 \end{gathered}$ | $\begin{gathered} 0.0 \\ -6.9 \\ -22.34 \\ -11.70 \end{gathered}$ |
| $\begin{array}{r} C-B \\ 2 A-C \\ B-A \\ 2 A-B \\ C-A \\ A+B-C \\ 2 B-C \end{array}$ | $\begin{aligned} & 1.00 \\ & 1.16 \\ & 1.42 \\ & 2.16 \\ & 2.42 \\ & 2.58 \\ & 4.00 \end{aligned}$ | $\begin{array}{r} -4.8 \\ -5.5 \\ -6.2 \\ -8.5 \\ -9.1 \\ -9.7 \\ -13.0 \end{array}$ | $\begin{aligned} & -53.97 \\ & -51.24 \end{aligned}$ | $-61.3$ <br> $-50.4$ <br> $-54.4$ <br> $-50.3$ <br> $-44.5$ <br> -55.1 <br> -62.1 | $\begin{aligned} & -58.6 \\ & -46.0 \\ & -56.0 \\ & -54.1 \\ & -39.8 \\ & -52.1 \\ & -63.3 \end{aligned}$ |
| Video signal power ( dB ) |  |  | 0.80 | 0.80 | 0.81 |
| Total weighted IM noise power ( AB ) |  |  | $-56.83$ | -50.1 | -46.5 |
| Video-to-weighted IM noise ratio |  |  | 57.63 | 51.9 | 47.3 |

Table 6.8: Signal and IM products power levels calculated by simulation for 6 MHz peak deviation and 4 pole, $0.5-d B$ ripple Chebyshev filter plus $5^{\circ} / d B$ satellite and demodulator AM/PM.

Table 6.8 shows that the highest IM product level for $N=0$ (i.e. no radio subcarrier) is 51.24 dB below the luminance level. However, with the addition of a radio subcarrier carrying 4 radio programs, the highest IM product level now is 39.8 dB below the luminance level. Though the weighted video $\mathrm{S} / \mathrm{N}$ is not severely degraded by this IM level (the weighted video $\mathrm{S} / \mathrm{N}$ degradation was found to be about 1.87 $\mathrm{dB})$, the $I M$ product may cause degradation to the subjective performance of the video signal. Tests should be performed to determine what subjective effect results. Note that the generated IM products involving the TV associated audio subcarrier are small and can be neglected. Thus, only the radio subcarrier is responsible for the generation of significant IM products because of its high power level. Therefore, in order to reduce IM interference in this case, the power level of the radio subcarrier has to be kept small, i.e. only a small number of radio programs should be transmitted and/or use only low quality level radio programs should be used.

It should also be noted that there are no IM products falling within the neighborhood of 5 MHz and 6 MHz which are the frequencies of the audio and radio subcarriers. Thus, these signals are not subjected to IM interference.

Furthermore, the locations of the audio and radio subcarriers which have been used so far were optimized based on the utilization efficiency of the baseband bandwidth. Optimization of the locations of these subcarriers based on interference as well as power utilization efficient is not carried out in this report because the process may involve subjective performance evaluation which should be done by hardware testing and beyond the scope of this study.

This section analyzes downlink interference to the wanted signal (TV or radio) from signals in adjacent transponders for the DBS case only*. Two scenarios are considered: intra-system interference where the interfering signal originates from the same satellite as the wanted one and (2) inter-system interference where the interfering and wanted signals come from different satellites. In the first scenario, the signals in the staggered cross-polarized transponders are considered as potential interferors. In the second scenario, the interfering signal which comes from an adjacent satellite is assumed to be co-polar and cochannel with the wanted one. Furthermore, the following assumptions apply to the analysis:

Transponder frequency plan (see Figure 6.16):

- Transponder usable bandwidth $=27 \mathrm{MHz}$ (DBS case).
- Spacing between co-polarized adjacent transponder = 39 MHz .
- Spacing between cross-polarized (staggered) adjacent transponders $=19.5 \mathrm{MHz}$.

TV signal format (see Figure 6.17):

- Video baseband bandwidth $=4.2 \mathrm{MHz}$ (NTSC).
- Audio sub-carrier frequency = 5 MHz .

[^17]

Figure 6.16:- DBS transponder frequency plan


Figure 6.17: TV baseband and RF. power. spectra.

```
- Peak deviation = 6 MHz (which results in Carson's
        rule bandwidth = 22 MHz).
- TV receiver noise bandwidth = 18 MHz.
```

Other assumptions are as follows:

- Uplink fade is neglected in the analysis*.
- Downlink fade is assumed to be the same for both the wanted and interfering signals.
- The interfering signal serves the same coverage area as the wanted one.


### 6.2.1 Intra-system Interference

Only interference from cross-polarized signals in staggered transponders is evaluated in this section. Interference from signals in co-polarized adjacent transponders is small and can be neglected since the spacing between such transponders and the wanted transponder is sufficiently large ( 39 MHz ).

The following cases are to be investigated:
(i) Dedicated transponder where the full transponder usable bandwidth is dedicated to the use of radio sCPC carriers only.
(ii) Shared transponder - SCPC case where the TV signal (FM modulated) and a number of radio SCPC carriers share the same transponder.

* This section considers downlink interference which depends on the power levels of the received wanted and interfering signals. For systems with automatic gain control (AGC), the effect of uplink fade on the downlink interference can be neglected.
(iii) Shared transponder - Sub-carrier case where the radio signals are carried by a sub-carrier located above the TV baseband spectrum.

In case (i), the only interference problem is the interference between the SCPC carriers. In case (ii), there are several interference problems which are interference from FMTV to FMTV, from FMTV to SCPC, from SCPC to FMTV and SCPC to SCPC.

### 6.2.1.1 Dedicated Transponder Case

It has been shown in Section 6.1.1 that the satellite capacity in this transponder category is power limited. Thus, there is more bandwidth available than needed. Therefore, the SCPC carriers can be arranged in such a way that the carriers in the staggered transponder fall into unused frequency slots in the wanted transponder (see Figure 6.18). As a result, interference from a staggered adjacent channel is eliminated.

Where the traffic in the staggered transponder is not SCPC but rather wideband, mutual interference between the signals can still be completely avoided by squeezing the SCPC carriers. within the 12 MHz bandwidth at the centre of the transponder. This is illustrated in Figure 6.19.
6.2.1.2 Shared Transponder - FMTV Plus SCPC Case

As suggested in Section 6.1.2, the radio SCPC carriers have to be grouped to one side of the TV carrier in order to avoid the comparably sized $2 A$ - B IM products from falling into the radio carrier bandwidth. There are two possible ways to arrange the $T V$ and SCPC carriers for the staggered adjacent transponders as shown in Figure 6.20. In both.


Figure 6.18:- SCPC carrier allocation to avoid adjacent transponder interference.


Figure 6.19: Centre 12 MHz bandwidth is used for the SCPC carriers to avoid interference from wideband signals in adjacent transponder.

arrangements, the $T V$ spectrum is centred at 2 MHz away from the center of the transponder so that some RF bandwidth is available on one side of the $T V$ spectrum for the radio SCPC carriers. These SCPC carriers are assumed to be at least 12.5 MHz away from the centre of the TV spectrum* and 1 MHz away from the transponder edge. The only difference between the two frequency arrangements is the location of the SCPC carriers with respect to the $T V$ carrier in the transponders. Figure 6.20 shows that the SCPC carriers are on the same side of the TV carrier in both wanted and staggered adjacent transponders in frequency arrangement \#l. In frequency arrangement \#2; the SCPC carriers are on the opposite side of the $T V$ carrier in the staggered adjacent transponders.

The following interference problems are analyzed for both frequency arrangements: interference from FMTV to FMTV, from FMTV to SCPC and from SCPC to FMTV. Note that there is no interference from SCPC to SCPC in both arrangements since these carriers do not overlap each other. Also note that in the analysis given below only signals falling within the noise bandwidth of the wanted receiver are considered as potential interferors and it is assumed that the downlink powers of the wanted and interfering TV signals are the same.

Since the interfering signal is cross-polarized with respect to the wanted one, then the received interference power is determined by two significant interfering components.[71] which are (1) component induced by rain depolarization [72] of the interfering signal and (2) component which is part of the interfering signal received through the receive antenna cross-polar gain. If the cross-polar discrimination due to rain depolarization is $X P D$ (in $d B$ ) and the cross-polar gain

[^18]relative to the on-axis (co-polar) gain is $G_{x}$ (see Section 6.2.2) then the interfering signal power at the receiver input normalized to the received wanted carrier power is
\[

$$
\begin{equation*}
I_{N}=r+\log \left(10^{\frac{-X P D}{10}}+10^{\frac{G}{10}}\right), d B \tag{6.21}
\end{equation*}
$$

\]

where $r(d B)$ is the ratio of the interfering signal power to the wanted signal power at the receiver input. Figure 6.21 plots the cross-polar discrimination, rain attenuation, normalized power of the interfering signal ( $I_{N}$ ) for $r=0 d B$ against the time percentage of the worst month. The values of rain attenuation $A$, cross-polar discrimination $X P D$ and $I_{N}$ are interpreted as follows. For a given time percentage of the worst month, say $p \%$, the actual rain attenuation will exceed the value shown in the figure for $p \%$ of the time and the actual cross-polar discrimination will exceed the value indicated in the figure for ( $100-\mathrm{p}$ ) \% of the time. Therefore, since $G_{x}$ is time invariant, $\dot{I}_{N}$ will not be greater, than the indicated value for ( $100-\mathrm{p}$ ) \% of the time but will never be less than $G_{x}$ (for $r=0 d B$ ).

In the case of clear weather (i.e. no clouds, rain), XPD is infinite and (6.21) becomes

$$
\begin{equation*}
I_{N}(\text { clear weather })=r+G_{x} \tag{6.22}
\end{equation*}
$$

The total received interference power at the output of the receiver filter is found by integrating the interfering power spectrum over the noise bandwidth of the filter. Note that if the interfering spectrum is normalized to the received wanted power, the integration will result in the reciprocal of the carrier-to-interference ratio.


Figure 6.21 - Cross-polar discrimination (XPD), receive antenna on-axis gain (GX), normalized cross-polarized interfering signal power ( $\mathrm{I}_{\mathbf{M}}$ ) for $r=0 \mathrm{~dB}$ and the rain attenuation ( $A$ ) versus time percentage of the worst month. Values of $A$ and XPD are for CCIR climatic region $K$ and signal path elevation angle of 25 degrees.

Interference from FMTV to FMTV:

Assume that the interfering TV power spectrum is approximately given by (6.6) and that the spectrum is bounded by the transponder bandwidth of 27 MHz , then the TV carrier-to-interference ratio is given by (see Figure 6.20)

$$
\begin{align*}
& \frac{C_{V}}{I_{V}}=\left[\begin{array}{c}
-\log \frac{1}{14.7}\left[\int_{10.5}^{15.5} S_{n}(f) d f+\int_{-11.5}^{-10.5} S_{n}(f) d f\right]-I_{N^{\prime}} \\
\text { for arrangement } \# 1
\end{array}\right. \\
& -10 \log \frac{1}{14.7} \int_{D^{11.5}} S_{n}(f) d f+\int_{n}^{-14.5} S_{n}(f) d f-I_{N^{\prime}} \\
& 6.5-15.5 \\
& \text { for arrangement \#2 } \tag{6.22}
\end{align*}
$$

where $I_{N}$ is given by (6.21) with $r=0 d B$ (see Figure 6.21) and $S_{n}(f)$ is given by (6.5).

After integrating the right-hand-side of (6.22), we obtain

$$
\frac{C_{V}}{I_{V}}=\left[\begin{array}{ll}
26.7-I_{N} & \text { for arrangement \#1 }  \tag{6.23}\\
12.3-I_{N} & \text { for arrangement \#2 }
\end{array}\right.
$$

With a clear-weather $T V$ total $C / N_{0}$ of $87 \mathrm{~dB}-\mathrm{Hz}$, the received TV $C / N_{o}$ (with the effect of rain attenuation) and $C / N_{o}$ degradation* are plotted against time percentage of the worst month in Figure 6.22. It can be that the $\mathrm{C} / \mathrm{N}_{\mathrm{o}}$ degradation due to interference to an FMTV signal from

[^19]

Time percentage of the worst month

Figure 6.22: FMTV carrier-to-noise density ratio and degradation due to (FMTV to FMTV) cross-polarized adjacent channe1 interference versus time percentage of the worst month.
cross-polarized FMTV signals in the staggered adjacent transponders is negligible for both. frequency arrangements shown in Figure 6.20.

It is worthwhile to note that the $C / N_{o}$ degradation calculated above is not necessarily the same as the baseband video (or audio) $S / N$ degradation since an interfering TV carrier is demodulated into a baseband interferor only when the difference between the instantaneous frequencies of the wanted and interfering carriers is less than the noise bandwidth of the baseband video lowpass filter, i.e. less than about 4.2 MHz . Analysis for baseband $\mathrm{S} / \mathrm{N}$ degradation is not undertaken here because the degradation is negligible and, furthermore such analysis is beyond the scope of this study. Reference [68] presents an approximate solution to this analysis.

## Interference from FMTV to SCPC

The analysis given in Section 6.1.2.2 can be applied to this interference problem. Similar to (6.7), the total interference power received is

$$
\begin{equation*}
I_{V}=\frac{I_{N} C_{R}}{14.7} \int_{D_{f}-\frac{B}{2}}^{D_{n}+\frac{B}{2}} S_{n}(f) d f \tag{6.23}
\end{equation*}
$$

where
$I_{N} \quad$ is the interfering signal power normalized to the radio signal power.
$C_{R} \quad$ is the radio signal power.
$D_{f} \quad$ is the spacing between the $T V$ carrier centre frequency and that of the radio carrier.

B
is the noise bandwidth of the radio receiver.
$S_{n}(f)$ is the approximate $T V$ spectrum and is. given by (6.5).

Rearranging (6.23), we obtain the radio carrier-tointerference ratio in $d B$ :

$$
\begin{equation*}
\frac{C_{R}}{I_{V}}=-I_{N}-\operatorname{lolog} \frac{1}{14.7}\left[\phi\left(D_{f}+\frac{B}{2}\right)-\phi\left(D_{f}-\frac{B}{2}\right)\right] \tag{6.24}
\end{equation*}
$$

where

$$
\phi(f)=\int_{-\infty}^{f} S_{n}(f) d f
$$

The worst case radio channel is the channel that is closest to the centre of the interfering $T V$ power spectrum in the staggered adjacent transponder. Thus, the worst case $C_{R} / I_{V}$ given by ( 6.24 ) corresponds to $D_{f}=5 \mathrm{MHz}$ and 9 MHz for frequency arrangement \#1 and \#2, respectively. Using these $D_{f}$ values, the clear weather $C_{R} / I_{V}$ for the worst case radio channel with CATV quality level and with CBPSK modulation is found to be 22.9 dB and 34.1 dB for frequency arrangements \# 1 and \#2, respectively*. By converting the interference power to equivalent thermal noise power, the

[^20]$\mathrm{E}_{\mathrm{b}} / \mathrm{N}_{\mathrm{o}}$ degradation is estimated at 0.45 dB for arrangement \#1 and 0.04 dB for arrangement \#2. (Note that the required $\mathrm{E}_{\mathrm{b}} / \mathrm{N}_{\mathrm{o}}$ for the radio channel is 13.3 dB which includes 11.3 $d B$ required for $a \operatorname{BER}=10^{-7}$ plus 2 dB implementation margin). Thus, it is seen that the radio SCPC carriers could suffer severe interference from the TV.signal in an orthogonally polarized staggered-cochannel transponder if frequency arrangement $\# 1$ is used. The degradation due. to ${ }^{\text {• }}$ this type of interference is negligible for frequency arrangement \#2 in clear weather. Nevertheless, it is found that $C_{R} / I_{V}$ for this frequency arrangement remains above 28 dB for $99.9 \%$ of the worst month.

Interference from SCPC to FMTV

The analysis for this interference problem was briefly described in Section 6.1.2.1. Numerical evaluation for the TV signal-to-noise degradation is rather lengthy and is not carried out in this section. An estimate of the TV carrier-to-interference is carried out as follows.

For frequency arrangement \#1, all SCPC carriers in the staggered adjacent channel fall within the wanted:TV receiver noise bandwidth. The carrier-to-interference ratio in clear weather is

$$
\begin{equation*}
\frac{C_{V}}{I_{R}}=-I_{N}=-r-G_{x} \tag{6.25}
\end{equation*}
$$

where

$$
\begin{aligned}
& r=\beta+10 \log N \\
& G_{x}=-25 d B
\end{aligned}
$$

and $\quad \beta$ is the radio $S C P C-t o-T V$ power ratio and $N$ is the number of radio channels.

As an example, suppose that the interfering radio signals are CATV quality level and that CPSK or CMSK is used, then $\beta=-17.9 \mathrm{~dB}$. As a result, $C_{V} / I_{R}=36.9 \mathrm{~dB}$ in clear weather for $N=4$. This low power interferor would cause negligible interference into the baseband video signal. However, the audio signal is more susceptible to this type of
interference (i.e. narrowband interference) than the video signal [68]. More detailed analysis is necessary to quantify the audio performance degradation.

Only one radio SCPC carrier falls within the wanted TV receiver noise bandwidth in frequency arrangement \#2. In this case, the clear weather $C_{V} / I_{R}$ for the same example above is 42.9 dB . This SCPC interferor would not degrade the subjective performance of the wanted TV signal because the interfering signal power is low and the interfering carrier is located far away from the centre of the wanted $T V$ signal ( 9 MHz spacing) so that it is demodulated into a baseband interferor which falls far above the highest frequency of the $T V$ baseband signal for most of the time.

Note that rain does have a significant impact on the crosspolarized interfering signal power level because of the depolarization of the interfering signal caused by rain. But, its impact on the additional performance degradation is not important since the degradation due to rain attenuation is the dominant factor.

### 6.2.1.3 Shared Transponder - Subcarrier Case

For the subcarrier case assume both wanted and interfering signals are the same. Also assume that there are no SCPC carriers sharing the same transponder with the carrier which is frequency modulated by the combined ( $T V$ and radio) signal. Also assume that the Carson's rule bandwidth in this case is the same as the case where no radio subcarrier
is used i.e., the peak frequency deviation due to video is reduced such that the total occupied $R F$ bandwidth is unchanged when a radio subcarrier is added. Reduction in peak frequency deviation requires increases in carrier power in order to maintain the same signal-to-noise ratio. Section 5.2.2 investigates this matter in detail.

Since the transponders each carry only one FM carrier, the carriers can be located at the centres of their respective transponders. Similar to the analysis for FMTV to FMTV interference in the preceding section, the carrier-tointerference ratio in this case is given by

$$
\begin{align*}
\frac{C_{V}}{I_{V}} & =-\log \frac{1}{14.7}\left[\int_{-13.5}^{-10.5} S_{n}(f) d f+\int_{10.5}^{13.5} S_{n}(f) d f\right]-I_{N} \\
& =26-I_{N} \tag{6.26}
\end{align*}
$$

where $I_{N}$ is the interfering signal power normalized to the wanted signal power and is given by (6.2l) with $r=0 d B$.

It can be seen that the interference level as given in (6.26) is negligible. As an example, the clear weather $C_{V} / I_{V}$ is $26+25=51 \mathrm{~dB}$.

For the case where the carrier power is kept fixed, i.e. the RF bandwidth has to be expanded in order to accomodate a radio subcarrier a higher interference level results due to the fact that both the $R F$ spectrum and the receiver noise bandwidth are wider. In this case, additional analysis similar to the one described in Section 6.1.2.1 is required to quantify the performance degradation for the TV (both video and audio) and the radio signals.

The amount of co-channel interference on the downlink from an adjacent satellite depends on the spacing between the wanted and interfering satellites and the side-lobe gain of the receive antenna. This section provides a series of equations for the calculation of the levels of co-channel (co-polar and cross-polar) interference from an adjacent satellite. A minimum satellite spacing for a given minimum C/I will be obtained by using these equations.

In addition to the assumptions listed at the beginning of Section 6.2, the following assumptions are also applied:

- The reference gain patterns for the 1 m ( 3 dB beamwidth $\phi_{o}=1.8^{\circ}$ ) DBS individual reception antenna as a function of the off-axis angle $\phi$ is given as follows [73]

Co-polar gain:

$$
G_{C}(\phi)=\left[\begin{array}{llr}
0.0 & \text { for } & 0 \leqslant \phi \leqslant .25 \phi_{O}  \tag{6.27}\\
-12\left(\phi / \phi_{o}\right)^{2} & \text { for } .25 \phi_{o} \leqslant \phi \leqslant .707 \phi_{O} \\
-\left[9+20 \log \left(\phi / \phi_{O}\right)\right] & \text { for } .707 \phi_{0} \leqslant \phi \leqslant 1.26 \phi_{0} \\
-\left[8.5+25 \log \left(\phi / \phi_{o}\right)\right] & \text { for } 1.26 \phi_{o} \leqslant \phi \leqslant 9.55 \phi_{0} \\
-33 & \text { for } & \phi \geqslant 9.55 \phi_{0}
\end{array}\right.
$$

Cross-polar gain:

$$
G_{X}(\phi)=\left[\begin{array}{lll}
-25.0 & \text { for } & 0 \leqslant \phi \leqslant .25 \phi_{O} \\
-\left[30+40 \log \left(l-\phi_{0} / \phi_{O}\right)\right] & \text { for } .25 \phi_{O} \leqslant \phi \leqslant .44 \phi_{O} \\
-20 & \text { for } .44 \phi_{O} \leqslant \phi \leqslant 1.4 \phi_{O} \\
-\left[30+25 \log \left(\phi / \phi_{O}-1\right)\right] & \text { for } 1.4 \phi_{O} \leqslant \phi \leqslant 2 \phi_{O} \\
-30 & \text { for } 2 \phi_{O} & \leqslant \phi \leqslant 7.24 \phi_{O} \\
G_{C}(\phi) & &
\end{array}\right.
$$

- In calculating the antenna gains, the geocentric angle (angle between the satellites as seen from the centre of the earth) is used rather than the topocentric angle (angle as seen from the receiving earth station). Conservative results will be obtained because the latter is always greater than the former.

The following are the downlink $C / I$ equations for co-channel interference from an adjacent satellite located at $\phi$ degrees from the wanted one. Recall that the two satellites are assumed to serve the same coverage area.

## FMTV to FMTV or SCPC to SCPC interference

$$
\frac{C_{V}}{I_{V}} \text { or } \frac{C_{R}}{I_{R}}=\left[\begin{array}{lc}
-G_{C}(\phi) & \text { Co-polar } \\
-G_{C}(\phi)-\log \left[10^{\frac{-X P D}{10}}+10^{\left.\frac{G_{x}(\phi)}{10}\right]}\right.
\end{array}\right.
$$

Cross-polar

FMTV to SCPC interference

The interference power is
where $\frac{C_{V}}{14.7} S_{n}(f)$ is the approximate interfering FMTV power spectrum and $B$ is the radio receiver noise bandwidth.

B/2
If $B \leqslant 6.3 \mathrm{MHz}$, then $\int S_{n}(f) d f=B$, thus (6.30) becomes $-B / 2$

$$
I_{V}=\left[\begin{array}{lc}
\log \left(\frac{C_{V} B}{14}\right)+G_{C}(\phi) & \text { Co-polar } \\
C_{B} & \\
\log \left(\frac{V}{14.7}\right)+G_{C}(\phi)+\log \left[10^{\frac{-X P D}{10}}+10^{\frac{G_{x}(\phi)}{10}}\right]
\end{array}\right.
$$

Therefore, if the radio carrier power is $C_{R}$, then the radio carrier-to-TV interference ratio is

$$
\frac{C_{R}}{I_{V}}=\left[\begin{array}{ll}
\beta-\log \frac{B}{14.7}-G_{C}(\phi) & \text { Co-polar } \\
\beta-\log \frac{B}{14.7}-G_{C}(\phi) \\
-\log \left[10^{\frac{-X P D}{10}}+10^{\frac{G_{X}(\phi)}{10}}\right] & \text { Cross-polar }
\end{array}\right.
$$

where $\beta=C_{R} / C_{V}$ is the SCPC-to-TV power ratio and is given in Table 6.4.

## SCPC-to-FMTV Interference

If there are N radio $\operatorname{SCPC}$ carriers falling within the wanted $T V$ receiver noise bandwidth, then the interference power is

$$
I_{R}=\left[\begin{array}{cc}
\log N C_{R}+G_{C}(\phi) & \text { Co-polar } \\
\log \log C_{R}+G_{C}(\phi) & \\
+10 \log \left[10^{\frac{-X P D}{10}}+10^{\frac{G_{X}(\phi)}{10}}\right] & \text { Cross-polar }
\end{array}\right.
$$

The TV carrier-to-radio interference ratio is

$$
\frac{C_{V}}{I_{R}}=\left[\begin{array}{cc}
-\beta-\log N-G_{C}(\phi) & \text { Co-polar } \\
-\beta-\log N-G_{C}(\phi) & \\
-10 \log \left[10^{\frac{-X P D}{10}}+10^{\frac{G_{X}(\phi)}{10}}\right] & \text { Cross-polar }
\end{array}\right.
$$

Equations (6.29), (6.32) and (6.34) are calculated for $\phi=$ 1, 2, 3, 5, 7, 10 and 15 degrees with $\mathrm{XPD}=\infty \mathrm{dB}$ (i.e. clear weather), $\beta=-17.9 \mathrm{~dB}, \mathrm{~B}=0.384 \mathrm{MHz}$ and $\mathrm{N}=10$. The results are shown in Table 6.9. It is seen from the table that if the clear weather interference level is to be kept at about 30 dB or above, the, the spacing between the satellites has to be $2.5^{\circ}$ or greater if the wanted and interfering signals are orthogonal (i.e. cross-polarized), otherwise the spacing has to be at least $15^{\circ}$.

It should be noted that the co-polar $C / I$ value is not affected by rain, but the cross-polar C/I value is. A detailed analysis of co-channel (co-polar and cross-polar) interference with the effects of rain attenuation and depolarization is given in [71].

In Sections 6.1 and 6.2, the internal and external interference problems for dedicated and shared transponder cases were examined in detail. The following remarks are drawn from the results of the analysis.


Table 6.9: Clear weather C/I resulting from co-channel adjacent satellite interference as a function of satellite spacing angle $\phi$.

For mutual interference between the $T V$ and radio signals within the wanted transponder (i.e. internal interference).

- Direct interference between the radio SCPC carriers in a dedicated transponder and between the FMTV and SCPC carriers in a shared transponder does not represent a serious interference problem since the interference level can be made as small as necessary by using sharp rolloff filters or/and sufficient spacing between the carriers.
- In a shared transponder, the radio SCPC carriers should be grouped to one side of the TV carrier to avoid comparably sized $2 f_{v}-f_{r}$ IM products. Furthermore, a guard band equal to the total bandwidth occupied by the radio SCPC carriers is needed between the $T V$ and SCPC carriers in order to avoid the $f_{v}+f_{r_{i}}-f_{r_{j}}$ IM products falling into the radio carriers. The radio carriers are thus only
 products.
- Interference due to intermodulation products requires the satellite transponder to operate in a backed-off mode (typically 4.5 dB output relative to single saturation) in the dedicated transponder case. Assuming the above frequency plan constraints; a shared transponder can be operated near saturation if total power assigned is well below that of the TV carrier (implying modest variation of the resultant signal envelope). Composite output backoff (typically $2.5-3.0 \mathrm{~dB}$ ) is required to reduce IM levels when radio carrier power approaches that of the TV carrier.
- Direct interference from radio subcarrier to baseband video signal or vice verse depends on the position of the subcarrier and the sharpness of the filters used.. It has been shown that a radio subcarrier located at 6 MHz causes neglegible direct interference into the baseband video.
- The radio subcarrier was found to be the main cause of baseband IM interference due to distortion of the FM carrier. As a result, the power level of the radio subcarrier has to be kept small to avoid excessive IM interference. It was also found that IM interference caused by the video associated audio is negligible.

For interference from signals in transponders other than the wanted one (i.e. external interference)*:

- Cross-polar interference from signals in the staggered transponders (in the same satellite, i.e. intra-system interference) is small in most cases except the shared transponder - SCPC case where the SCPC carriers may suffer severe interference from the adjacent transponder FMTV signal if the FMTV and SCPC frequency arrangement in the adjacent transponder is the same as in the wanted one (see Figure 6.20a). The frequency arrangement shown in Figure 6.20 b may be used in which the FMTV to SCPC interference level is reduced by about ll dB as compared to that in the other arrangement.

[^21]- Co-channel with adjacent satellite interference was also analyzed for a 1 m antenna ( $1.8^{\circ} 3 \mathrm{~dB}$ beamwidth) receive antenna. It was found that two satellites located $2.5^{\circ}$ apart can serve the same service area without causing more than -30 dB of interference power into each other under clear weather conditions if the satellites transmit orthogonal (crosspolarized) signals. If the satellites transmit signals of the same polarization (co-polarized signals) to the same service area, a spacing between the satellites of about $15^{\circ}$ is needed to achieve the same interference level.

It should be noted that the analysis performed in this section assumes no FEC coding employed for the radio program. The analysis can be repeated with the assumption that FEC coding is used. However, note that with the use of FEC coding, the required radio power is decreased by the coding gain and the required (noise) bandwidth is increased by the bandwidth expansion of the code. Thus, interference to a TV signal from radio signal(s) will be smaller compared to that in the no FEC coding case. Nevertheless,
interference from $T V$ to radio will be larger due to lower power and wider receiver noise bandwidth required for the radio signal.

SYSTEM IMPLEMENTATION, FLEXIBILITY AND GROWTH CONSIDERATIONS

In this section, block diagrams for the transmit and receive stations are provided for the two transponder categories (dedicated and shared). Other factors such as compatibility with TV direct broadcast systems (for redistribution) and with $T V$ and radio receive equipment, system flexibility and growth capability are also discussed.

## 7.1

7.1.1 Radio SCPC, TDM/FDMA Implementation

Figures 7.1 show the basic earth station transmit and receive block diagrams SCPC and/or TDM/FDMA operation.

The SCPC transmitter will transmit as many carriers (N) as required using iN channel units each processing a 15 kHz program channel with a PCM encoder (number of levels determined by required quality level), FEC channel encoder and the desired modulator. The low level modulator outputs will be combined; the result is frequency translated to the final RF and amplified in an HPA. For $N>3$, the HPA will have to be backed off ( 3.5 dB output back-off minimum).

The SCPC receiver configuration will depend upon the application. For CATV applications all programs may need to be re-distributed simultaneously and $N$ receive channel units would be required. For direct-to-home applications a single channel unit could be used to reduce cost, with program selection based on carrier frequency selection using a tuneable down-converter or local oscillator (see Figure 7.3).

(a) Transmit Station

Figure 7.1: Block diagrams for $T x / R x$ stations for radio SCPC carriers (dedicated transponder case)

CHANNEL UNIT \#1

(b) Receive Station

Figure 7.1 (cont'd)



The TDM transmit and receive block diagrams differ from the SCPC ones in that program channels are combined and received digitally and only single equipment chains are needed following (preceding) the multiplexor (demultiplexor).

With the use of SCPC, intermodulation interference and frequency uncertainty of system oscillators are especially of concern. While interference due to intermodulation can be minimized by backing off the satellite transponder, the composite frequency uncertainty of the oscillators could be equal to the bandwidth of the SCPC carrier and therefore automatic frequency control (AFC) is required. One common AFC technique is to send in addition to the SCPC carriers a pilot tone which is tracked to the phase-locked loop (PLL) at the receiver

If a pilot tone is sent by each station originating a number of SCPC carriers, then transmitter, satellite and receiver. frequency offsets can be tracked and spectrum centering may be performed on the SCPC carriers. While AFC is required for TDM carriers as well, the degree of compensation decreases as the number of carriers and quality level increases i.e. the TDM bandwidth to frequency uncertainty ratio increases.

Radio programs transmitted using TDM/FDMA require multiplexing of the data bits which increases the system complexity as compared to SCPC/FDMA. However, this technique utilizes the transponder power and bandwidth more efficiently (see Section 5.l). As a result, more radio programs can be transmitted by using TDM/FDMA than SCPC/FDMA.

At the receive side, the equipment required to receive the radio programs transmitted by SCPC could be relatively simple. Figure 7.3 illustrates a possible equipment arrangment for this case. A tunable local oscillator is needed to select the individual radio program.SCPC carrier and converts it to a fixed $I F$ frequency, say 70 MHz , for demodulation. The recovered radio baseband signal can be amplified to drive a speaker or fed to a re-modulator for redistribution. The receiver for TDM/FDMA is more complicated since it involves demultiplexing before the individual radio programs are recovered. However, while the TDM/FDMA technique allows several radio programs to be accessed simultaneously by a single receiver, the SCPC receiver as shown in Figure 7.3 can recover only one radio program at a time. Thus, in terms of the complexity of the receiver, SCPC is more suitable for individual reception (i.e. home receiver) where it is likely that only one radio program is accessed at a time. For redistribution media, TDM/FDMA is suggested since in this case, several radio programs need to be (remodulated and) re-transmitted simultaneously.
7.1.2 Radio SCPC, TDM/FDMA System Flexibility and Growth

The use of SCPC for radio program distribution would probably provide the most flexibility as compared to TDM/FDMA and subcarrier techniques since the radio programs can be transmitted independently can originate from different transmit stations. Furthermore, the radio program slot(s) in the satellite transponder can be accessed by any transmit station. For each TDM carrier in the TDM/FDMA mode, all radio programs must be routed (by
terrestrial microwave links, satellite links, or any other means) to a central station where the multiplexing process takes place before the radio programs can be transmitted.

It has been shown in Section 6.1.1.2 that the system capacity for the dedicated transponder case is power limited. Thus, system growth is limited by the available satellite power and the earth station G/T. By increasing the earth station $G / T$, more radio programs can be transmitted.

It should be noted that in a power-limited situation, the available power and capacity can be allocated between SCPC and TDM carriers on a proportionate basis. If the multicarrier capacity of the transponder is $N$ program channels, then these can be grouped as SCPC and TDM carriers are desired.
7.1.3 Subsystem Specifications

Table 7.1 summarizes the technical parameters for the proposed digital radio subsystem. Rate $-3 / 4$ FEC coding is suggested to be used to code the most significant, bits (MSB) of the PCM word. The FEC code reduces the required carrier power by an amount of 3.6 dB for coherent PSK modulation. In the case where the system capacity is severely power-limited, a lower rate code with higher coding gain (the complexity of the codec is of course higher) can be used. The rate $-1 / 2$ convolutional code with Viterbi decoding was suggested in Section 4.3 for use in such a case. The use of error concealment is also suggested. This technique improves the subjective

| QUALITY LEVEL | CATV | HIGH | LOW |
| :---: | :---: | :---: | :---: |
| Signal Baseband Bandwidth (kHz): | 15 | 15 | 15 |
| Sampling Rate ( kHz ) : | 32 | 32 | 32 |
| Source Encoding: | Companded PCM | Companded PCM | Companded PCM |
| Number of Bits per Sample: | 12 | 11 | 9 |
| Bit Rate per Channel (kbps) | 384 | 352 | 288 |
| Error Control- |  |  |  |
| - Number of FEC Coded MSB's | 4 | 4 | 3 |
| - FEC Code Rate: | 3/4 | 3/4 | 3/4 |
| - Coding Gain(dB) for CPSK(DBPSK/DMSK): | 3.3(2.8) | 3.3(2.9) | 3.2(2.4) |
| - Error Concealment | YES | YES | YES |
| Transmission Rate per Channel (kbps) | 458.7 | 426.7 | 352.0 |
| Modulation: | Coherent BPSK; or <br> differential BPSK, MSK |  |  |
| Channel Spacing (in multiple of symbol rate) | 31.8 | >1.8 | \$1.8 |
| Receive IF Filter - |  |  |  |
| - Amplitude Response: | Resembles 3rd or 4th-order |  |  |
| - Group delay at 3 dB point [34]: | Butterworth filter. <br> $\leqslant 10 \%$ of the bit interval. |  |  |
| -3 dB Bandwidth: | Equal to the symbol rate. |  |  |
| Required Unfaded $C / N_{o}(d B-H z)$ per program channel*- |  |  |  |
| - Differential BPSK Modulation: | 66.9 | 66.5 | 65.3 |
| - Differential MSK Modulation: | 67.4 | 67.0 | 65.8 |
| Bit Error Rate | $10^{-7}$ | $10^{-7}$ | $10^{-6}$ |
| Unweighted $\mathrm{S} / \mathrm{N}(\mathrm{dB})$ : | 60 | 55 | 43 |

[^22]performance of the PCM encoded radio signal. Note that while the FEC code operates only on the MSB's, normally error concealment checks the parity of the whole PCM word. Further investigation is required to determine the possible use of both these techniques.

Coherent BPSK, differential BPSK and MSK can be used in this transponder category. BPSK modulation was chosen over QPSK modulation since both give the same BER performance but the BPSK modem is relatively simpler than QPSK modem. Furthermore, since differential detection is less sensitive to oscillator phase noise than coherent detection, then a differential receiver can be built with less strigent stability requirements on the local oscillator. Hence, DBPSK modulation is suggested for use in the case where the cost of the receiver is to be kept low.
7.2 Shared Transponder: FMTV Plus Radio SCPC
7.2.1 FMTV Plus Radio SCPC Implementation

The block diagrams of the the transmit and receive systems are given in Figure 7.4.

The same problems mentioned above, namely, IM interference and oscillator frequency stability need to be considered in the design of this system. The radio SCPC carriers have to be grouped to one side of the TV carrier and a guard band between the TV and SCPC is necessary to avoid large IM products falling into the bandwidth occupied by the SCPC carriers. The pilot tone approach used to correct local


Figure 7.4: Block diagrams for $T x / \mathrm{Rx}$ stations for FMTV plus radio SCPC (shared transponder FMTV/SCPC case)
oscillator frequency shift may not be economically feasible in this case. Because the TV signal spectrum occupies most of the (DBS) transponder bandwidth leaving a small
unoccupied bandwidth for the radio SCPC carriers. Thus, sending a pilot tone can be very costly in such a bandwidth constrainted system. An alternative solution to this problem is to recover the coherent carrier from the TV wideband FM signal. (see Appendix D). Spectrum centering then can be done with this recovered carrier.

One of the advantages of this system is the TV and radio receivers can share $R F$ equipment to the second downconverter output to provide simultaneous TV and radio reception. Figure 7.4 b shows the block diagram of such receivers.
7.2.2 FMTV Plus Radio SCPC System Flexibility and Growth

Because this system uses SCPC access for the radio carriers the inherent flexibility associated with SCPC is retained:

- the TV and radio transmission systems can be designed independently,
- the TV and radio programs can originate from different transmit stations, and
- the receivers need not be constrained to receive a TV signal in order to receive radio programs (as in the subcarrier technique). However, both the TV and radio programs can be received simultaneously by $T V$ and radio receivers sharing the same second downconverter.

The design of the (cross-polarized) staggered adjacent transponder in this sytem, however, is constrained by the adjacent channel interference to the radio SCPC carriers which are located at the edge of the transponder. It has been shown in Section 6.2.2.2 that the radio SCPC carriers may suffer severe interference from FMTV signal in the staggered transponder if the TV signal is placed close (in terms of frequency spacing) to the wanted radio signals: An appropriate plan for arranging the FMTV and SCPC carriers in the staggered adjacent transponders is shown in Figure 6.20b.

Another factor that limits the flexibility and growth of the system is the limitation of both satellite power and bandwidth. For example, the frequency arrangmeent as shown in Figure 6.20 b for the DBS satellite allows only about 3 MHz transponder bandwidth for the radio SCPC traffic. This bandwidth accommodates only $4 \mathrm{BPSK} / \mathrm{SCPC}$ CATV quality radio channels. For the ANIK C satlelite, the system capacity is power limited. See Section 6.1.2.3 and Appendix B for more details.

Radio Subcarrier Implementation

The hardware implementation of the transmit and receive. stations in the subcarrier implementation is more complicated than in the previous case due to the need

| Baseband Video (NTSC): | 4.2 | MHz |
| :---: | :---: | :---: |
| Baseband TV associated Audio: | 5 | kHz |
| Audio Subcarrier Frequency: | 5 | MHz |
| Peak Frequency Deviation due to Video: | 6 | M zz |
| Peak Frequency Deviation due to Audio Subcarrier: | . 56 | MHz |
| Peak Frequency Deviation of TV-associated Audio FM Subcarrier: | 25 | kHz |
| Required RF Bandwidth | 22 | MHz |
| Receiver IF Filter Noise Bandwidth: | 18 | M z |
| Required $\mathrm{C} / \mathrm{N}_{0}$ | 87 | $\mathrm{dB}-\mathrm{Hz}$ |
| Video S/N (13 dB de-emphasis and noise weighting advantage [70]): | 44.6 | dB |
| Audio S/N (10.5 dB de-emphasis and noise weighting advantage [70]): | 49.4 | dB |

Table 7.2: TV subsystem specifications
for multiplexing and demultiplexing. As indicated in Figure 7.5 , the signals are filtered by sharp rolloff filters at both the transmit at receive sides in order to reduce direct interference between each other (see Section 6.1.3.1). Intermodulation products generated by the subcarriers were also found to be a potential problem in this FDM technique.

Intermodulation interference onto the video and the associated audio signals can be kept small by using only. one radio subcarrier which is modulated by TDM'd radio signals. Using more than one subcarrier for the radio programs increases not only the IM interfernce level but also the complexity of the transmit station. However, the use of multiple subcarriers eliminates the need for multiplexing and demultiplexing if each subcarrier carries only one radio program. As a result, the radio programs can be accessed by a relative simple receiver.

There is another problem that should be considered in the design of the subcarrier system; that is, the frequency deviation of the RF carrier may be increased in order to accommodate a radio subcarrier without any increase in the required power (see section 5.2.2). In this case, the receiver IF bandwidth has to be increased to avoid signal distortion caused by bandwidth limitation. However, an increase in receiver IF bandwidth will result in more received thermal and interfering noise power. Furthermore, this would make the receiver become incompatable with FMTV systems which use an 18 MHz noise bandwidth. Thus, provided that the bandwidth is kept fixed, the radio channels which are carried by a subcarrier can be accessed

(a) Transmit Station
$\begin{aligned} \text { Figure 7.5: } & \begin{array}{l}\text { Block diagrams for the } T x / R x \text { stations for } T V \text { and radio } \\ \text { subcarrier (shared transponder subcarrier case) }\end{array}\end{aligned}$


Figure 7.5 (cont'd)
by a plug-in unit (which includes PSK/MSK demodulator, demultiplexer, $D / A$ converter) added to the TVRO terminal.
7.3.2 Radio Subcarrier System Flexibility and Growth

The subcarrier case is considerably less flexible than the SCPC case due to the following reasons:

- The TV and radio programs have to be transmitted from the same transmitting centre,
- the receiver has to demodulate the TV signal in order to receive radio programs,
- failure of the TV signal transmitter or FM demodulator will result in a loss of radio signal.

The subcarrier system growth is limited by the satellite power and bandwidth and also the baseband bandwidth available between the TV associated audio subcarrier and the second harmonic of the colour subcarrier.

### 7.3.3 Subsystem Specifications

The radio subcarrier is located at 6 MHz which is about half way between the audio subcarrier and the second harmonic. The addition of this subcarrier will increase the required RF power or/and bandwidth as shown in Section 5.2.3. This section gives the values of peak frequency deviations due to the video and subcarriers as functions of the number of TDM'd radio channels for two cases: (l) fixed RF bandwidth (i.e. the RF bandwidth remains unchanged), and (2) fixed RF power equivalent to $C / N_{o}=87$ $\mathrm{dB}-\mathrm{Hz}$. Note that QPSK modulation with coherent detection
which requires the least bandwidth and power is recommended for the subcarrier case since there is a limited bandwidth available for the radio subcarrier. Thus, this section considers only coherent QPSK. Results for coherent BPSK and MSK are identical to those for coherent QPSK, but there are smaller numbers of radio programs that can be transmitted by using BPSK modulation.

Assume that the performance of the radio and TV are as shown in Tables 7.1 and 7.2, the peak frequency deviations due to video and the subcarriers are given in Tables 7.3 and 7.4 for CATV quality level and coherent PSK, MSK. The values in the tables are found by using (5.9a), (5.9b), (5.12a), and (5.12b). It is seen from the tables that if the RF bandwidth is kept fixed, the peak deviations due to video and audio must be reduced and the required RF power (i.e. power of the modulated carrier) has to be increased in order to accomodate the radio subcarrier. On the other hand, if the $R F$ power is kept fixed, the peak deviations due to video and audio must remain unchanged to maintain the required performance. The $R F$ bandwidth is then increased (i.e. more composite deviation) to accommodate the radio subcarrier.

|  | NUMBER <br> OF RADIO <br> CHANNELS | REQUIRED <br> TOTAL <br> RADIO <br> $\mathrm{C} / \mathrm{N}$ 。 <br> (dB) | $\begin{gathered} \Delta \mathrm{f}_{\mathrm{V}} \\ (\mathrm{MHz}) \end{gathered}$ | $\begin{gathered} \Delta f_{A} \\ (\mathrm{MHz}) \end{gathered}$ | $\begin{gathered} \Delta \mathrm{f}_{\mathrm{R}} \\ (\mathrm{MHz}) \end{gathered}$ | ```REQUIRED RF C/No (dB)``` | REQUIRED <br> RF BAND- <br> WIDTH <br> ( MHz ) |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
| No <br> FEC <br> CODING | 1 | 69.1 | 5.15 | 0.48 | 0.93 | 88.32 | 21.5 |
|  | 3 | 73.9 | 4.67 | 0.44 | 1.46 | 89.18 | 21.5 |
|  | 5 | 76.1 | 4.38 | 0.41 | 1.76 | 89.73 | 21.5 |
| RATE- $\frac{3}{4}$ <br> FEC <br> CODING | 1 | 65.8 | 5.39 | 0.50 | 0.66 | 87.94 | 21.5 |
|  | 3 | 70.6 | 5.02 | 0.47 | 1.07 | 88.55 | 21.5 |
|  | 5 | 72.8 | 4.79 | 0.45 | 1.32 | 88.95 | 21.5 |

Table 7.3: Peak deviations as functions of number of CATV quality radio channels for a fixed $R F$ bandwidth. $\Delta f_{V}, \Delta f_{A}$ and $\Delta f_{R}$ are the peak deviation due to video, audio subcarrier and radio subcarrier, respectively. Note that if there is no radio subcarrier then $\Delta f_{V}=6.0 \mathrm{MHz}$, $\Delta f_{A}=0.56 \mathrm{MHz}$ and the required $\mathrm{RF} \mathrm{C} / \mathrm{N}_{\mathrm{O}}$ is $87.0 \mathrm{~dB}-\mathrm{Hz}$.

|  | NUMBER <br> OF RADIO <br> CHANNELS | $\begin{aligned} & \text { REQUIRED } \\ & \text { TOTAL } \\ & \text { RADIO } \\ & \text { C/No } \\ & \text { (dB) } \end{aligned}$ | $\begin{gathered} \Delta \mathrm{f}_{\mathrm{V}} \\ (\mathrm{MHz}) \end{gathered}$ | $\begin{gathered} \Delta f_{A} \\ (M H z) \end{gathered}$ | $\begin{gathered} \Delta \mathrm{f}_{\mathrm{R}} \\ (\mathrm{MHz}) \end{gathered}$ | $\begin{aligned} & \text { REQUIRED } \\ & \text { RF } \mathrm{C} / \mathrm{N}_{0} \\ & (\mathrm{~dB}) \end{aligned}$ | REQUIRED RF BANDWIDTH (MHz) |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
| NO <br> FEC CODING | 1 | 69.1 | 6.0 | 0.56 | 1.08 | 87.0 | 23.7 |
|  | 3 | 73.9 | 6.0 | 0.56 | 1.87 | 87.0 | 25.3 |
|  | 5 | 76.1 | 6.0 | 0.56 | 2.42 | 87.0 | 26.4 |
| RATE- $\frac{3}{4}$ <br> FEC <br> CODING | 1 | 65.8 | 6.0 | 0.56 | 0.74 | 87.0 | 23.0 |
|  | 3 | 70.6 | 6.0 | 0.56 | 1.28 | 87.0 | 24.1 |
|  | 5 | 72.8 | 6.0 | 0.56 | 1.65 | 87.0 | 24.8 |

Table 7.4: Peak deviations as functions of number of CATV quality radio channels for a fixed $R F$ power. $\Delta f_{V}, \Delta f_{A}$ and $\Delta f_{R}$ are the peak deviation due to video, audio subcarrier and radio subcarrier, respectively. Note that if there is no radio subcarrier then $\Delta f_{V}=6.0 \mathrm{MHz}$, $\Delta f_{A}=0.56 \mathrm{MHz}$ and the required $R F$ bandwidth is 21.5 MHz .

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This report investigates digital source encoding, modulation, multiplexing and error correction techniques appropriate to the distribution of radio programs by satellite. The candidate techniques were evaluated for two satellite scenarios - (i) dedicated transponder and (ii) shared transponder. In the first scenario, the whole transponder is dedicated to the transmission of only radio programs which can be transmitted using SCPC and/or TDM/FDMA techniques. In the second scenario, the radio programs and an FMTV signal are transmitted through the same transponder. There are two ways for the $T V$ and radio signals to share the transponder: (1) the $T V$ and radio signals are transmitted independently as SCPC carriers, and (2) the radio programs are time division multiplexed to modulate a subcarrier which is located above the baseband TV signal. Table 8.1 lists the candidate techniques considered in the report.

Adjacent channel and intermodulation interference affecting the TV and radio signals were analyzed in detail in the. report. Furthermore, system capacity (i.e. number of radio channels that can be supported), system flexibility and growth were also discussed.

The following conclusions are derived from the result of the study.

## Source Encoding:

- Pulse code modulation is appropriate to the encoding of 15 kHz radio program since the $T T / \mathrm{N}$ performance can be achieved by using PCM. With the use of $\mu=255$ law companding, the dynamic range of the codec can be increased to about 40 dB . In addition, both FEC coding and error concealment techniques can be used with PCM.

|  | Candidate | Aspects Considered for Study | Reference |
| :---: | :---: | :---: | :---: |
| SOURCE <br> ENCODING | Pulse code modulation (PCM) and delta modulation (DM) | (i) Bit rate and bit error rate performance to meet a required $T T / N$ performance | Section 2.1 |
| MULTIPLEXING | Continuous and packet multiplexing | (i) Bit rate efficiency <br> (ii) Error susceptibility <br> (iii)Ad:aptibility to different applications <br> (iv) SIfstem complexity (or cost) | Section 2.2 |
| MOD- <br> ULATION | Coherent and differential BPSK, QPSK MSK | (i) Modem performance <br> (ii) Power and bandwidth utilization efficiencies <br> (iii)Susceptibility to interference <br> (iv) System complexity (or cost) | Section 3.0 <br> Section 5.0 <br> Section 6.0 <br> Section 3.0 |
| ERROR CORRECTION | FEC coding and error concealment | (i) Redundancy and coding gain <br> (ii) System complexity | Section 4.0 |

Table 8.1 Candidate digital encoding, multiplexing, modulation and error correction techniques considered for radio program distribution.

- Delta modulation gives a $T T / N$ of less than 43 dB for $a$ bit rate of $400 \mathrm{kbit} / \mathrm{s}$ and a 5 kHz test tone, $T T / \mathrm{N}$ of less than 28.5 dB for the same bit rate is achieved for a 15 kHz test tone. These $T T / \mathrm{N}$ 's are below the required $T T / N$ for high quality level*. Furthermore, the simple but powerful error correction technique, namely, error concealment cannot be used with delta modulation. Thus, for radio program distribution, delta modulation is inferior to PCM.


## Multiplexing:

- Packet multiplexing provides more adaptibility to different applications than continuous multiplexing. But the latter is a suitable technique for a simple rigid structure (i.e. fixed data rate) receiver and for a higher bit rate efficiency. Note that there are existing systems for radio program distribution which use continuous multiplexing. These systems are NICAM-3 [8], Sony [10] and Scientific Atlanta [17].


## Modulation:

- The relative performance requirements of the modulation techniques are listed below in ascending order.
- Power requirement: Coherent BPSK, QPSK, MSK

Differential BPSK
Differential MSK
Differential QPSK

- Bandwidth requirement: QPSK

MSK
BPSK

[^23]\author{

- Susceptibility to interference: MSK BPSK <br> QPSK <br> - Phase noise requirement: Coherent BPSK Differential BPSK Differential MSK <br> Differential QPSK <br> Coherent MSK <br> Coherent opsk <br> - Modem complexity: Differential BPSK <br> Differential MSK <br> Differential QPSK <br> Coherent BPSK <br> Coherent QPSK <br> Coherent MSK
}
- Considering the relative performances listed above, the only drawback of (coherent and differential) BPSK is its bandwidth inefficiency. Differential QPSK requires the highest power and coherent QPSK puts the most stringent requirement on the local oscillator phase noise. On the other hand, the coherent MSK demodulator is the most complex one. Differential MSK is a compromise for these requirements. Differential MSK requires more power than coherent BPSK (about 1.1 dB was assumed in the report), its bandwidth utilization efficiency is slightly better than that of coherent BPSK, but its phase noise requirement is less stringent. Furthermore the DMSK demodulator is simple.
- Quality levels and satellite scenarios should also be considered in the choice of suitable modulation techniques. For CATV and high quality levels, which are intended for redistribution media and community reception, the receiver cost factor is not as important
as that for lower quality home reception. For the dedicated transponder scenario, there is more bandwidth available than required since the system capacity is power-limited. However, for the shared transponder scenario, there are limiations on both power and bandwidth (Note that the bandwidth is not limited in ANIK-C for FMTV/SCPC case).
- Consequently, it is seen that coherent and differential BPSK and differential MSK are the most promising modulation techniques for the dedicated transponder scenario and coherent OPSK is more suitable for the shared transponder scenario due to its bandwidth efficiency. Differential QPSK is not recommended because of its high power requirement which is not appropriate to power-limited systems.

A cost-benefit analysis is required to identify the specific modulation technique. A tradeoff exists between equipment (earth segment) and satellite utilization (space segment) costs.

## Error Correction:

- With the use of PCM for source encoding, error concealment should be used to improve the subjective performance of the radio program. In addition to error concealment, forward acting error correction (FEC) may be used on the most significant bits (MSB) of the PCM word to efficiently decrease the required power but still achieve the objective performance. It was found that if the bit error rate of the bit stream representing the radio signal at the input of the FEC decoder is less than $10^{-3}$ then only $n / 3$ (i.e. one-third
of the word) MSB's of an n-bit long PCM word need FEC coding in order to obtain the same performance as the case where all bits are coded.
- The rate $-3 / 4$ convolutional code with threshold decoding is suggested to be used for the radio program distribution. The suggested code provides a coding gain of about 3.6 dB (based on $\mathrm{BER}=10^{-7}$ and coherent PSK) and requires a relatively simple decoder. In the case where more coding gain is needed, a rate $-1 / 2$ convolutional code with Viterbi decoding is suggested. The code provides a coding gain of about 4.8 dB (based on $B E R=10^{-7}$ and coherent PSK), but it requires more bandwidth and a more complicated decoder. However, LSI implementations are expected to be commercially available in the next 2 years.


## Interference:

- For the dedicated transponder case, the adjacent channel interference level can be made small and negligible by using sharp roll-off filters and large spacings between adjacent channels. It was found that with an IF receive filter having amplitude response resembling that of a 4 th 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 level. 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.
- For the shared transponder FMTV/SCPC case, interference to the FMTV signal is not a severe problem. There is no direct interference to the FMTV carrier from the radio SCPC carriers since the latter are located outside the

TV receiver $I F$ filter noise bandwidth. Interference to FMTV due to intermodulation degrades the FMTV carrier-to-noise density ratio by about 1.0 dB if there are 4 CATV quality radio channels (with no FEC coding) sharing the same transponder with the FMTV carrier and if the transponder is operated at 2 dB backoff. However, interference to the radio $S C P C$ carriers is significant because these carriers are low power. The following restrictions must be imposed on the SCPC carriers in order to avoid excessive interference on them:

- All SCPC carriers should be grouped to one side of the FMTV carriers to avoid comparably sized $2 f_{v}-f_{r}$ IM products.
- Any SCPC carriers should be located at least 13 MHz away from the centre of the FMTV spectrum to avoid direct interference from the FMTV signal.
- To avoid interference from cross-polarized staggered transponder, the FMTV signal in the staggered transponder should be located at least 9 MHz from any SCPC carriers in the wanted transponder.
- For the shared transponder subcarrier case, there is little interference to the subcarrier. Interference due to intermodulation between the colour and radio subcarriers may cause degradation to the subjective performance of the TV signal. Note that, in this case, the video associated audio subcarrier causes no significant $I M$ products because this subcarrier is low power.
- Interference from an adjacent DBS satellite was also investigated in the report. If the adjacent satellite serves the same coverage area as the wanted one, then the spacing between the satellites should be at least $2.5^{\circ}$ if the wanted and interfering signals are orthogonal (i.e. cross-polarized) and $15^{\circ}$ if the signal are co-polarized.


## System Capacity:

- For the dedicated transponder case, the system capacity (number of radio channels can be transmitted through the transponder) is power limited. Despite the fact that the EIRP from the ANIK $C$ satellite is smaller than that from the DBS satellite the former is capable of transmitting more radio channels if the earth station receive antenna $G / T$ is $16 \mathrm{~dB} / \mathrm{K}$ or larger (see Appendix B). This is because the larger transponder bandwidth of the ANIK $C$ satellite (twice as large as that of the DBS satellite) allows a greater staggering advantage (reduce the degradation due to IM interference).
- For the shared transponder FMTV/SCPC case, the system capacity is power limited and, for the DBS satellite only, the capacity is also limited by the available RF bandwidth. The number of radio channels that can be transmitted using the subcarrier technique is limited by the constrained baseband bandwidth within the TV associated audio frequency and the second harmonic of the colour subcarrier. Table 8.2 compares the earth station receive antenna $G / T$ 's required to receive from 2 to 5 radio channels transmitted using the FMTV/SCPC and subcarrier techniques. The results shown in the table assume 2 dB input backoff for the FMTV/SCPC case and no input backoff for the subcarrier case. It is seen from the table that, in the FMTV/SCPC case, if the total

|  | DBS Satellite |  |  | ANIK C Satellite |  |  |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: |
|  | FMTV/SCPC |  | Subcarrier | FMTV/SCPC |  |  |
| Number <br> of Radio Channels | Radio/TV <br> Power <br> Ratio $=-14 \mathrm{~dB}$ | Radio/TV <br> Power <br> Ratio $=-12 \mathrm{~dB}$ |  | Radio/TV <br> Power <br> Ratio $=-14 \mathrm{~dB}$ | Radio/TV <br> Power <br> Ratio $=-12 \mathrm{~dB}$ | Subcarrier |
| 2 | 14.8 | - | 13.1 | 19.2 | - | 17.6 |
| 3 | 16.6 | 14.4 | 13.5 | 21.0 | 18.7 | 18.0 |
| 4 | 18.0 | 15.7 | 13.8 | 22.2 | 19.9 | 18.3 |
| 5 | 18.9 | 16.6 | 14.0 | 23.1 | 20.8 | 18.5 |

Table 8.2: Earth station receive antenna $G / T$ in $d B / K$ required to support the TV signal and 2 to 5 radio channels for DBS and ANIK C satellites. 2 dB input backoff is assumed for the FMTV/SCPC case and the transponder is assumed to be operated at saturation for the subcarrier case.
radio-to-TV power ratio at the transponder input is increased by $2 d B_{r}$ the required $(G / T) E$ is decreased by more than that. This happens because of the small signal suppression phenomenon. The table also indicates that the subcarrier technique requires smaller (G/T) E than the $F M T V / S C P C$ technique for a given number of radio channels (only applicable for less than 5 channels) and that the ANIK $C$ system requires larger $(G / T) E$ than the the DBS satellite system.

System Flexibility and Growth:

- The use of the SCPC/FDMA technique for radio program distribution provides the most flexibility as compared to $T D M / F D M A$ and subcarrier techniques since the radio and TV signals can be transmitted and received independently. Furthermore, the radio SCPC carriers can be transmitted from different locations. TDM/FDMA and subcarrier techniques require that the radio programs associated with each RF carrier have to be transmitted from the same location and that the radio and $T V$ programs have to be received simultaneously in the subcarrier case. TDM is better suited to CATV applications, while subcarrier techniques represent a limited, but low cost implementation approach.

This report has identified and evaluated candidate digital source encoding, modulation, multiplexing and error correcting techniques which are appropriate to the distribution of radio programs by satellite. However, further work is proposed relating to the following areas:

- Subjective and objective performance improvements due to the use of error concealment on the digital radio signal in conjunction with FEC,
- The use of reduced bit rate source encoding techniques such as adaptive differential PCM (ADPCM),
- Frequency plans for both the shared transponder FMTV/SCPC and subcarrier cases were suggested (see Figures 5.16 and 6.20). However, these plans, especially for the subcarrier case, should be evaluated once the exact TV frequency deviation, TV receiver IF filter specifications and number of radio channels must be supported are known,
- As identified in Section 6.1.3.2 for the subcarrier case, the subjective performance of the video signal may be degraded by large intermodulation products. Hardware tests should be performed to determine the extent of the degradation.
- Further examination of the FMTV-based AFC technique described in Appendix D.


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## APPENDIX A

## NOISE DUE TO DIGITAL ERRORS IN PCM

CODECS WITH NO COMPANDING

APPENDIX A

NOISE DUE TO DIGITAL ERRORS IN PCM
CODECS WITH NO COMPANDING
A. 1 Natural Binary (NB) Code

For a natural binary code, an error in the LSB causes an error $q$ in the recovered sample (where $q$ is the quantizer step size). An error in the second LSB causes an error of $2 q$ in the recovered sample. An error in the third LSB causes an error of $2^{2} q$ in the recovered sample. Therefore, the averaged error noise power $\overline{\varepsilon_{n}^{2}}$ is given by:

$$
\begin{aligned}
\overline{\varepsilon_{n}^{2}} & =\frac{1}{n}\left[q^{2}+(2 q)^{2}+\left(2^{2} q\right)^{2}+\ldots\left(2^{n-1} q\right)^{2}\right] \\
& =\frac{1}{n} q^{2}\left[\frac{2^{2 n}-1}{2^{2}-1}\right] \cong \frac{2^{2 n} q}{3 n}=\frac{(2 A)^{2}}{2 n}=\frac{4}{3 n} A^{2}
\end{aligned}
$$

where

$$
\begin{aligned}
\mathrm{n}= & \text { number of bits per code word } \\
\mathrm{A}= & \text { the quantizer peak maximum input swing for no- } \\
& \text { clipping. }
\end{aligned}
$$

and

$$
q=\frac{A}{2^{n}} \text { is the step size. }
$$

## A. 2 Folded Binary (FB) Code

In a folded binary code, an error in the MSB bit, which is the sign bit, causes an r.m.s. recovered sample folding error of $\left(2 \sigma_{s}\right)^{2}$ where $\sigma_{s}$ the input signal r.m.s. voltage.

An error in each of the less significant bits has the same effect as in the NB code. Therefore the average error noise power is given by

$$
\begin{aligned}
\overline{\varepsilon_{n}^{2}}= & \frac{1}{n}\left[4 \sigma_{s}^{2}+q^{2}(2 q)^{2}+\ldots+\left(2^{n-2} q\right)^{2}\right] \\
= & \left.\frac{1}{n} 4 s+q\left[\frac{2^{2 n}-1}{2-1}-2^{2 n-2}\right]\right\}= \\
& \frac{1}{n} 4_{s}+q \frac{2^{2 n}-3 \cdot 2^{2 n-2}}{3}= \\
= & \frac{1}{n} 4_{s}+\frac{q 2^{2 n}}{4 \cdot 3}=\frac{1}{n} 4 s+\frac{1}{3} A
\end{aligned}
$$

(Note that for $F B$ code $q=\frac{A}{2^{n-1}}$ ).

## A. 3 <br> Gray (G) Code

When the Gray code is used, noise due to digital error depends on the pdf and level of the input signal.

If the pdf of the input signal is $p(x)$, the noise due to digital errors is given by:

$$
\begin{aligned}
\overline{\varepsilon_{n}^{Z}}= & {\left[4 \sigma_{S}^{2}+2 \int_{0}^{A}(2 x-A)^{2} P(x) d x+\right.} \\
& \left.2\left\{\int_{0}^{A / 2}\left(2 x+\frac{A}{2}\right)^{2} P(x) d x+\int_{A / 2}^{A}\left(2 x-\frac{3}{2} A\right)^{2} P(x) d x\right\}+\ldots\right] .
\end{aligned}
$$

$$
\text { A. } 4
$$

Codes Performance Comparison

The signal power to errors induced noise power ratios for the three codes are given for $\frac{A}{\sigma_{S}}=12.7 \mathrm{~dB}$ by:

$$
\begin{aligned}
& \frac{\sigma_{S}^{2}}{\varepsilon_{n}^{2}}(N B)=-13.9-10 \operatorname{lgP}_{e} d B \\
& \frac{\sigma_{S}^{2}}{\varepsilon_{n}^{2}}(F B)=-10.1-101 g P_{e} d B \\
& \frac{\sigma_{S}^{2}}{\varepsilon_{n}^{2}}(G)=-11.7-10 \operatorname{lgP}_{e} d B(P(x) \text { is gaussian })
\end{aligned}
$$

It is observed that the folded binary code is affected by digital errors less than the other two codes.

APPENDIX B

SYSTEM CAPACITY AS A FUNCTION

OF THE EARTH STATION RECEIVER
$G / T$

## APPENDIX B

## SYSTEM CAPACITY AS A FUNCTION OF THE EARTH STATION RECEIVER G/T

This Appendix gives the maximum number of radio channels that can be transmitted through a dedicated or shared transponder as a function of the earth station receive antenna G/T. The results presented here are for DBS and ANIK C satellites whose saturating flux densities, EIRP's, transponder bandwidths and satellite receive antenna. G/T's are given in Table 6.1 in Section 6.1.l.2.
B. 1 System Capacity for Dedicated Transponder Case

The available radio carrier-to-noise density ratio is given by

$$
\left(\frac{C}{N_{0}}\right)_{\text {avail }}=\left[\left(\frac{C}{N_{0}}\right)^{-1}+\left(\frac{C}{N_{0}}\right)_{\text {down }}^{-1}+\left(\frac{C}{I_{0}}\right)^{-1}\right]^{-1}
$$

where

$$
\begin{aligned}
& \left(\frac{C}{N_{O}}\right)_{\text {up }}=\phi_{S}-\log \left(\frac{4 \pi}{\lambda^{2}}\right)-B_{i}-k+\left(\frac{G}{T}\right)_{S} d B \\
& \left(\frac{C}{N_{0}}\right)_{\text {down }}=E I R P-L_{d}-B_{o}-k+\left(\frac{G}{T}\right)_{E} d B \\
& \left(\frac{C}{I_{0}}\right)=\left(\frac{C}{I}\right)_{C C}+\left\lfloor 10 \log \frac{B W_{R F}}{N}-1.5\right\rfloor \mathrm{dB}
\end{aligned}
$$

with

$$
\begin{aligned}
& \phi_{S}=\text { saturating flux density } \\
& \lambda=\text { unlink wavelength } \\
& B_{i}=\text { satellite input backoff } \\
& \mathrm{k}=\text { Boltzmann's constant } \\
& \left(\frac{G}{T}\right)_{s}=\text { satellite receive antenna } G / T \\
& \text { EIRP = satellite transmitted EIRP } \\
& L_{d}=\text { downlink free space path loss } \\
& B_{i}=\text { satellite output backoff } \\
& \left(\frac{C}{I}\right)_{c c}=\text { centre channel carrier-to-IM noise for equal } \\
& \text { and equally spaced carriers (see Figure 6.4). } \\
& \mathrm{BW}_{\mathrm{RF}}=\text { satellite useable transponder bandwidth } \\
& {\left[\log \frac{\mathrm{BW}_{\mathrm{RF}}}{\mathrm{~N}}-1.5\right]=\mathrm{IM} \text { interference improvement }} \\
& \text { factor due to channel } \\
& \text { staggering }
\end{aligned}
$$

If the required carrier-to-noise density ratio per radio channel is ( $\left.\mathrm{C} / \mathrm{N}_{\mathrm{o}}\right)_{\text {req }} \mathrm{I}^{\prime}$ ' then the number of radio channel supported by the system is:

$$
N=\frac{\left(\frac{C}{N_{0}}\right) \text { avail }}{\left(\frac{C}{N_{0}}\right) \text { req'd }} \quad \text { channels }
$$

|  |  |  |  |  |  |  |  |  |  |  |  |  |  |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
|  |  |  |  |  | $\cdots$ |  |  | - |  |  |  |  |  |
|  |  |  |  |  |  |  | , |  |  |  |  |  |  |
|  |  | $\cdots$ |  |  |  |  |  |  |  |  |  |  |  |
|  |  |  |  | andwuld | dth 73 | ftefor | BPSSK |  |  |  |  |  |  |
|  |  |  |  |  |  |  |  |  |  |  | (a) | ATV | , |
| $\cdots$ | $\underline{\square}$ | $\underline{+}$ | $\underline{\sim}$ | $\bigcirc$ | $\pm$ |  | - | $\pm$ | $\cdots$ |  |  |  |  |
|  |  | $\bigcirc$ | 1 | - | - |  | $\cdots$ | + |  | - |  | Ha-u. | 2 |
| 20. | $\underline{+}$ | $\pm$ |  | $+$ | + | - | $\underline{+}$ | GPSK | CMSK | $\square$ |  |  |  |
| +1 | +1+1 | $\square$ | + | + + | + | - |  |  |  |  |  |  |  |
|  | ,1, |  |  |  |  |  |  | Leb |  |  |  |  |  |
| + | + |  |  |  |  | - | 1-30 | SMSK. |  |  |  |  |  |
|  |  |  |  |  |  |  |  |  |  |  |  |  |  |
|  |  |  |  |  |  |  |  | DQPSK |  |  |  |  |  |
| 10. |  |  |  |  |  |  |  |  | $1 \cdot 1$ |  |  | 1 |  |
|  |  |  |  |  | 2 |  | - | - | - |  | , |  |  |
| $\cdots$ |  |  |  |  |  |  |  |  |  |  | $\cdots$ | - |  |
|  |  |  |  |  |  |  |  |  |  |  |  |  |  |
| 5. |  |  |  | $\cdots$ |  |  | $\underline{ }$ |  |  |  |  | - | B, K |
|  | - |  |  |  | - | - |  |  |  |  |  | C | cide |
|  | $\underline{\underline{I N}}$ |  |  |  |  |  |  |  |  |  |  |  |  |



Figure B.1: Number of radio channels plotted against earth station receive antenna G7T for DBS satellite (dedicated transponder case)
(1)


|  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
|  |  |  | $\bigcirc$ |  |  |  |  |  |  |  |  | - | T1 |  |  |
| 100 |  |  |  |  |  |  |  |  | - |  |  |  |  |  |  |
| , |  | Bandi | ath 1 | $t$ for |  | , |  |  |  |  |  |  |  |  |  |
|  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |
|  |  |  |  |  |  |  | $\bigcirc$ |  | 1 | , |  |  |  |  |  |
|  |  |  |  |  |  | Dec: | $\ldots$ |  | . | , |  |  |  |  |  |
| $50$ |  |  |  | Prex | 7 | MSt |  |  |  |  |  |  |  |  |  |
|  |  |  |  |  |  | - | - |  | - |  |  |  |  |  |  |
|  |  |  |  | , |  |  | , |  |  |  |  |  |  | Qual |  |
|  |  |  |  |  |  | d, | $\cdots$ |  | , |  |  |  |  |  |  |
| $\cdots$ | $\underline{3}$ | $\underline{51}$ | $\cdots$ | - |  |  |  | , |  |  |  |  |  |  |  |
|  | $\cdots$ | $\underline{+}$ | $\underline{-1}$ | 1 | + |  |  | $\underline{\square}$ |  |  |  |  | - | $\pm$ |  |
|  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |
| 20 | T | $\square$ | - | 1 | T | $\square$ | - | $\square+$ | 1 |  |  |  | $\underline{-}$ |  |  |
|  |  | + |  | - | - | - | - | - | + | - |  | - |  |  |  |
|  |  | +1 |  | $1+$ | $\div$ |  |  | +1 | - |  |  |  |  | $\cdots$ |  |
|  |  | +: |  |  |  |  |  |  |  |  |  |  |  |  | $(G) 1)^{2}$ |
| 16 | 6 | +1: |  | - 20 | 0 |  |  | 2 | 4 |  |  |  | $\bigcirc$ | , 2 |  |
|  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |

[^24]The value of $N$ is plotted against the earth station receive antenna $G / T$ in Figures $B .1$ and $B .2$ for three quality levels and for DBS and ANIK $C$ satellites. The system capacity limited by the transponder bandwidth is also plotted for BPSK modulation which is the least bandwidth efficient. It is seen from Figure B. 1 that the system capacity for DBS satellite is severely power limited. The figure suggests that higher $(G / T)_{E}$ value should be used for optimum power and bandwidth utilization of the transponder. Figure B. 2 shows that the ANIK C capacity is also power limited except for BPSK modulation with $(G / T)_{E}$ of about $26 \mathrm{~dB} / \mathrm{K}$ or more. Furthermore; the figures show that coherent detection provides higher system capacity than differential detection. This is because differential detection requires more power. For example, DQPSK requires 2.3 dB more power than CQPSK, thus systems which use CQPSK can support about $70 \%$ more radio channel than those which use DQPSK.

Note that the results plotted in Figures B. 1 and B. 2 are for radio program without FEC coding. If FEC coding is used, the power limited system capacity will increase by the amount equal to the coding gain. However, the bandwidth limited capacity will decrease by the same proportion with the bandwidth expansion of the code. Nevertheless, FEC coding can be useful for this case since the system capacity is power limited.

## B. 2 <br> System Capacity for Shared Transponder FMTV/SCPC Case

Calculation of the system capacity in this case is similar to that described above. But, in this case, since the radio carriers share the same transponder with the FMTV carrier then only a small number of radio carriers can be supported.

Figures B. 3 and B. 4 plot the number of CATV quality radio carriers that can be transmitted along with an FMTV signal


Figure B. 3: Number of CATV quality radio channels supported by DBS satellite transponder as a function of $(G / T)_{F}$. The minimum $(G / T)_{E}$ required to obtain FMTV $C / N o=87 \mathrm{~dB}-\mathrm{Hz}$ is 13.14 dB and 13.64 dB fo F total input radio-to-TV power ratio of -14 dB and -12 dB , respectively. (Shared transponder FMTV/SCPC case)


Figure B.4: Number of CATV quality radio channels supported by ANIKCsatellite transponder as a function of $(G / T)_{E}$. The minimum $(G / T)_{E}$ required to obtain FMTV $C / \mathrm{No}=87 \mathrm{~dB}-\mathrm{Hz}$ is $17.13 \mathrm{~dB} / \mathrm{K}$ and $17.51 \mathrm{~dB} / \mathrm{K}$ for total input radio-to-TV power ratio of -14 dB and -12 dB , respectively. (Shared transponder FMTV/SCPC case)
from the transponder of a DBS or ANIK C satellite against the earth station receive antenna $G / T$ for uplink radio-to-TV power ratios of $-12 d B$ and $-14 d B$ and for satellite input backoff of 2 dB . The bandwidth limited capacity is also shown for the DBS satellite*. The figures also show the minimum $(G / T)_{E}$ which is required to obtain an FMTV carrier-to-noise density of $87 \mathrm{~dB} \cdot \mathrm{~Hz}$. Note that the results in the figures are for CATV quality radio programs with no FEC coding. Furthermore, PSK and MSK modulations with coherent detection were assumed. System capacity with differential PSK and MSK can be found easily by dividing the power limited capacity given in Figures B. 3 and B. 4 by the excess power required for differential decoding. With the use of FEC coding, the capacity will increase by the amount equal to the coding gain.

It is seen from the figures that the larger the radio-to-TV ratio is, the higher the number of radio programs can be transmitted. However, a higher radio-to-TV power ratio results in lesser $T V$ power transmitted from the satellite. As a result, higher earth station receive antenna $G / T$ is required to maintain the received $T V$ carrier power level. Furthermore, if more radio programs are transmitted, the transponder has to be backed-off to reduce IM interference.

## B. 3 <br> System Capacity for Shared Transponder Subcarrier Case

The FM equations in this case are:

[^25]\[

$$
\begin{aligned}
& \left(\frac{S_{V}}{N_{0}}\right)=6\left(\frac{C}{N_{0}}\right) \frac{\Delta f_{V}^{2}}{f_{v}^{2}} \\
& \left(\frac{C_{A}}{N_{0}}\right)=\frac{1}{2}\left(\frac{C}{N_{0}}\right)_{R F} \frac{\Delta f_{A}^{2}}{f_{A}^{2}} \\
& \left(\frac{C_{R}}{N_{0}}\right)=\frac{1}{2}\left(\frac{C}{N_{0}}\right)_{R F} \frac{\Delta f_{R}^{2}}{f_{R}^{2}} \\
& B W_{T}=2\left(\Delta f_{V}+\Delta f_{A}+\Delta f_{R}+f_{v}\right)
\end{aligned}
$$
\]

where

$$
\begin{aligned}
\frac{S_{v}}{N_{O}} & =\text { video signal-to-noise density ratio } \\
\frac{C_{A}}{N_{O}} & =\text { audio subcarrier-to-noise density ratio } \\
\frac{C_{R}}{N_{O}} & =\text { radio subcarrier-to-noise density ratio } \\
\left(\frac{C}{N_{O}}\right)_{R F} & =\text { received RF carrier-to-noise density ratio } \\
\Delta f_{v}= & \text { peak frequency deviation due to video signal } \\
\Delta f_{A}= & \text { peak frequency deviation due to audio } \\
\Delta f_{R}= & \text { peak carrier frequency deviation due to radio } \\
& \text { subcarrier }
\end{aligned}
$$

$$
\begin{aligned}
& \mathrm{f}_{\mathrm{V}}=\text { peak frequency of the video signal }=4.2 \mathrm{MHz} \\
& \mathrm{f}_{\mathrm{A}}=\text { audio subcarrier frequency }=5 \mathrm{MHz} \\
& \mathrm{f}_{\mathrm{R}}=\text { radio subcarrier frequency }=6 \mathrm{MHz} \\
& \mathrm{BW}_{\mathrm{T}}=\text { Carson's rule bandwidth of the } \mathrm{FM} \text { signal. }
\end{aligned}
$$

Solving the four equations above for $\left(\frac{C}{N_{O}}\right)_{R F}$, we obtain:


With $\left(\frac{S^{V}}{N_{0}}\right)=97.88 \mathrm{~dB}-\mathrm{Hz},\left(\frac{\mathrm{C}_{\mathrm{A}}}{\mathrm{N}_{\mathrm{O}}}\right)=65 \mathrm{~dB}-\mathrm{Hz}$ and $\left(\frac{\mathrm{C}_{\mathrm{R}}}{\mathrm{N}_{\mathrm{O}}}\right)=$ required $\mathrm{C} / \mathrm{N}_{\mathrm{o}}$ per radio channel $+\log ($ number of radio channels), the required $F M$ signal carrier-to-noise density ratio $\left(\frac{C}{N_{0}}\right){ }_{R F}$ is then computed as a function of the number of radio channels. Finally, the earth station receive antenna required to obtain the $\left(\frac{C}{N_{0}}\right)_{R F}$ values can be calculated by using the link equations described in Section Bul. Note that since there is only one FM carrier which carries the composite (TV plus radio) signal, then the transponder can be operated at saturation in this case. The required earth station antenna $G / T$ required to receive the saturated FMTV signal alone (ie. TV signal with no radio subcarrier) is ll. $25 \mathrm{~dB} / \mathrm{K}$ and $15.81 \mathrm{~dB} / \mathrm{K}$ for DBS and ANIK $C$, respectively.

Using the approach described above, the FM signal carrier power are calculated for 1 to 5 CATV quality radio channels. The results without and with FEC coding are plotted in Figures B. 5 and B.6, respectively. Comparing Figures B. 3 and B. 4 with Figure B.5, it is seen that the subcarrier technique requires less power than the FMTV/SCPC technique: It is worthwhile to note that in the subcarrier case, the $R F$ bandwidth can be expanded (i.e. use larger. frequency deviation) to reduce the carrier power required. For example, Figure B. 5 shows that if the total RF bandwidth is expanded to 27 MHz from 23 MHz (i.e. increase the composite peak deviation to 9.3 MHz from 7.3 MHz$)$, the required $(G / T)_{E}$ drops by an amount of 2.1 dB .


Figure B.5: Number of CATV quality radio channels plotted against power in excess of $87 \mathrm{~dB}-\mathrm{Hz}$. No FEC coding is used.


Figure B.6: Number of CATV quality radio channels plotted against power in excess of $87 \mathrm{~dB}-\mathrm{Hz}$. Rate $-3 / 4$ FEC code is used.

## APPENDIX C

## PROGRAM TO EVALUATE DISTORTION

IN FM SYSTEMS

## APPENDIX C

## PROGRAM TO EVALUATE DISTORTION IN FM SYSTEM

This Fortran IV program simulates transmission of a waveform over a multiple hop (e.g. satellite or terrestrial radio) FM system consisting of the following elements:

Transmitter

- preemphasis filter
- ideal FM modulator

Link (repeated $n$ times)

- bandpass filter
- nonlinear amplifier

Receiver

- ideal FM demodulator
- deemphasis filter
in order to determine the resulting distortion. For FDM message signals, this appears as intermodulation (IM) noise; for video signals, as distortion of luminance and chrominance components, and interference to and from audio sub-carrier(s).

Equally spaced complex time or frequency samples are used to represent the signal at each point in the $R F$ channel. Manipulation of these samples consists of applying simple complex Fortran arithmetic and the fast Fourier Transform
(FFT). To simulate ideal FM modulation, successive phase samples from a complex array are equated to successive samples of the integrated (real) baseband waveform multiplied by a constant depending on rms deviation. Ideal FM demodulation consists of differentiating the phase function of a channel distorted version of this same complex array. Transformation of the array to the frequency domain is performed to more easily handle the effects of filtering.

The message signal is represented by a sum of one hundred equally spaced tones whose amplitudes and phases are arbitrary. The distortion of video test patterns or other periodic waveforms may be evaluated on the basis of change to their Fourier coefficients following transmission through the FM system. Representing white noise as a set of constant amplitude, random phase tones, with some amplitudes set to zero corresponding to the process of removing slots, permits the NPR (noise power ratio) test to be simulated as described*. The resulting picowatts of IM noise may then be added to the thermal noise contribution. To achieve the necessary accuracy, typically $\pm 0.5 \mathrm{~dB}$, this simulated NPR test must be repeated and results averaged for several independent choices of the phases of the baseband tones.

The program is presently configured to perform one of the following distortion evaluations as specified by the user:

- NPR test for $F D M / F M$ system
- differential gain, differential phase and harmonic levels for video system with audio sub-carrier(s)
- plotted cycle of transmitted and received versions of a periodic waveform and computed signal-to-distortion power ratio

[^26]While initially designed to analyze the effects of channel filtering only, the program now handles additive RF interferers which are also simulated. This capability has been applied in optimizing the bandwidth utilization of FDM/FM carriers in adjacent satellite RF channels. More generally, the simulation allows the modelling of even more complex channels that could for example include additive thermal noise, phase noise (due to frequency translation), AGC, AFC, modem nonlinearity, threshold extension demodulation, companding, etc.

As a systems tool, the program permits transmission parameters such as FM deviation, equalizer coefficients, and receiver noise bandwidth to be optimized, and degradations to be quantified. Note that unlike analytical models, the simulation places no restrictions on either modulation index or types of channel filters.

The program has been thoroughly tested against measurements and computed results obtained by other methods. In particular, computed NPR's and harmonic levels for single pole filters, ideal cut-off filters, and linear/quadratic group delay, with and without limiter AM/PM conversion, show excellent agreement with published results, as do comparisons on computed and measured NPR's for 1800 channel terrestrial radio (low index) and 960 channel satellite (moderate index) FDM transmission, and three tone television tests over satellite links at a variety of deviations.


#### Abstract

APPENDIX D

AUTOMATIC FREQUENCY CONTROL


## APPENDIX D

## AUTOMATIC FREQUENCY CONTROL (AFC)

A problem with receiving low rate $\operatorname{SCPC}$ is a large frequency offset relative to the SCPC bandwidth. Requirements on frequency uncertainty and phase noise at the demodulator are more stringent for radio program than for TV because of the narrower bandwidths involved. The normal solution for this problem is to use a pilot tone which is transmitted continuously with which a circuit can perform spectrum centering. This approach is economically feasible only for the case where the pilot tone does not represent a significant loss in terms of power available for SCPC signals. However, for the shared transponder - SCPC case, the TV signal may take up almost all of the transponder allowing perhaps 5 SCPC carriers. Use of a pilot tone is costly both in terms of available satellite capacity and single channel (tunable) receiver cost. Performing AFC on the TV carrier (wideband FM) is a more economical alternative.

The proposed circuit that can be used to recover a narowband carrier centred from a wideband $F M$ signal is shown in Figure D.l. The circuit consists of a zero-crossing detector which provides a squared-up version of the wideband FM signal; $a \div N$ frequency divider which divides the frequency and phase deviation of the signal; and a phase-locked loop. Computer simulation has been done for voice signal as the modulating signal* and for $\mathrm{N}=18$. The simulated loop performance results are shown along with theoretical results in Figure D.2. It is seen from the figure that for small RMS phase jitter, the closed loop response has to be very narrowband. Further studies such as loop performance simulation for modulating video signal (instead of voice signal) and hardware tests are necessary to optimize the performance and hardware implementation of the circuit.

[^27]Wideband
FM Signal In


Figure D.1: FM coherent carrier recovery system

## APPENDIX E

## CHANNEL CAPACITY AND INTERFERENCE ANALYSIS USING 23 MHz DBS TRANSPONDER

## APPENDIX E

CHANNEL CAPACITY AND INTERFERENCE ANALYSIS USING 23 MHz DBS TRANSPONDER

In Section 6.0 and Appendix B, the system capacity and interference analysis were carried out for both DBS ( 27 MHz transponder bandwidth) and ANIK C (54 MHz transponder bandwidth) cases. This appendix presents results for the DBS case where a 23 MHz transponder bandwidth is used. Except for the bandwidth differential, satellite characteristics are assumed to be the same in both cases. The transponder bandwidth has a direct impact on the following:

- system capacity
- adjacent channel interference
- intermodulation interference
E.1 System Capacity Analysis
E.l. 1 System Capacity for dedicated Transponder

As shown in Appendix $B$ the system capacity is power-limited for both DBS ( 27 MHz ) and ANIK C cases. With the transponder bandwidth reduced to 23 MHz , the system capacity is also power limited. Furthermore, fewer radio channels can be transmitted through a 23 MHz transponder than through a 27 MHz one. Because there is less bandwidth in which the radio carriers can be staggered to avoid interference from intermodulation products. Thus, the transponder has to be operated at a larger input backoff.

Figure E.l plots the optimum transponder capacity versus the earth station receive antenna $G / T$ for $C A T V$, high and low

quality levels. The bandwidth-limited capacity is also indicated in the Figure for BPSK modulation which is the least bandwidth efficient. It is seen from the figure that the capacity is power limited in each case.
E.1.2 System Capacity for Shared Transponder FMTV/SCPC Case

Figure E. 2 plots the number of radio SCPC carriers that can be transmitted with an FMTV signal in a transponder versus the earth station receive antenna $G / T$ for a total radio-toTV input power ratio of -14 dB and for a composite satellite input backoff of 2 dB . The following assumptions apply to the results shown in the figure:
(i) Radio carriers are placed at least 1 MHz from the transponder edge to avoid signal disstorion due to high group delay near the transponder edge. The carriers are also placed at least 9 MHz away from the center of the transponder to avoid direct interference to the FMTV signal as described in Section 6.1.2.1. As a result, the bandwidth available to the radio carriers is only 1.5 MHz .
(ii) Both direct interference and intermodulation interference are not considered.

For the -14 dB radio-to-TV power ratio considered, a receiver antenna $G / T$ of at least 13.14 dB is required to support one TV signal and one CATV quality radio signal (except when DOPSK is used). Note that the Carson's rule bandwidth of the TV signal is about 22 MHz * which is almost as large as the transponder bandwidth. Thus, there is little RF bandwidth available for radio SCPC traffic. As a

[^28]
#### Abstract

result, it is expected that the radio SCPC carriers may be severely degraded by interference from the FMTV signal. The interference problems for this case will be addressed in a later section.


E.1.3 System Capacity for Shared Transponder Subcarrier Case

The system capacity for the subcarrier case was evaluated in Appendix B, Section B. 3 for transponder bandwidths of 23 MHz and 27 MHz . It is shown in Appendix $B$ that to transmit 5 CATV quality CPSK (or CMSK) radio channels through a 23 MHz transponder using the subcarrier technique, requires approximately 1.8 dB additional power relative to the videoonly requirement. Only about $l d B$ more power (relative to video-only case) is required if rate $-3 / 4$ FEC coding is employed for the radio signals.
E. 2 Interference and Transponder Sharing Analysis

Section 6.0 of the report evaluates interference effects in detail for a DBS satellite using a transponder bandwidth of 27 MHz . . This section uses the analysis to present results for the 23 MHz case.
E.2.1 Internal Interference
E.2.1.1 Dedicated Transponder Case

Adjacent channel interference in this case may not be a severe interference problem since the system capacity is power limited, i.e. there is more bandwidth available than required so that the radio carriers can be spaced widely apart to avoid adjacent channel interference.


Figure E.2: Number of radio channels versus earth station receive antenna $G / T$ for DBS satellite. The input radio-to-TV power ratio is -14 dB (23 MHz shared transponder FMTV/SCPC case)

The carrier-to-noise ratio degradation due to intermodulation interference for CATV quality reception is shown in Table E.l. It is seen from the table that the $\mathrm{C} / \mathrm{N}$ degradation due to $I M$ interference increases with increasing $(G / T)_{E}$ despite the fact that the input backoff is increased. This is because increasing $(G / T)_{E}$ permits the number of carriers to increase resulting in more IM products. The table also shows that with a $\operatorname{lm}$ receiving antenna and $360^{\circ} \mathrm{K}$ noise temperature ( $G / T \simeq 13.8 \mathrm{~dB} / \mathrm{K}$ ), 16 CATV quality level radio channels can be transmitted through a 23 MHz transponder*.
E.2.1.2 Shared Transponder - FMTV/SCPC Case

A possible frequency plan for the FMTV/SCPC case is shown in Figure E.3. The Carson's rule bandwidth of the FMTV signal is 22 MHz which is almost equal to the usable transponder bandwidth of 23 MHz . Thus it is almost impossible to avoid overlapping between the FMTV and radio SCPC spectra. As a result, the $F M T V$ signal may cause strong direct interference to the radio signal(s). As estimated in Figure 6.8 if the radio SCPC carrier is located at 11 MHz away from the center of the FMTV spectrum, a (CATV quality, CBPSK) radio carrier-to-noise TV interference ratio of 17 dB at $\mathrm{BER}=10^{-7}$ by an amount of about 1.1 dB [58].

Another interfering source which could severely degrade the radio signals occuring during third-order $I M$ product $f_{V}+$ ${ }^{f} r_{1}-f_{r_{2}}$ light video loading. To avoid products, a guard band equal to at least the total bandwidth occupied by the radio carriers is needed between the bandwidths occupied by the FMTV and radio signals. This guard band is obviously not feasible in this case because the FMTV already virtually occupies the whole transponder. An alternative solution is to transmit only one radio channel so that there is no such IM product.

[^29]| $\begin{aligned} & (\mathrm{G} / \mathrm{T})_{E} \\ & (\mathrm{~dB} / \mathrm{K}) \end{aligned}$ | Optimum <br> Input Back- <br> off (dB) | Output Backoff ( dB ) | C/N Degradation due to $I M$ (dB) | Transponder <br> Capacity <br> (Carriers) |
| :---: | :---: | :---: | :---: | :---: |
| 4 | 4.5 | 1.9 | 1.61 | 5 |
| 6 | 5.2 | 2.1 | 1.93 | 7 |
| 8 | 6.7 | 2.5 | 2.10 | 9 |
| 10 | 7.4 | 2.8 | 2.25 | 12 |
| 12 | 7.7 | 2.9 | 2.54 | 14 |
| 14 | 8.2 | 3.1 | 2.80 | 16 |

Table E.I:
C/N degradation due to intermodulation interference and transponder capacity for CATV quality level with CPSK or CMSK quality level with CPSK or CMSK modulation (for required $\mathrm{C} / \mathrm{N}_{\mathrm{o}}=69.1 \mathrm{~dB}-\mathrm{Hz}$ ).


Figure E.3: Frequency plan for FMTV and radio SCPC carriers.

So far, only interference from FMTV to radio SCPC carriers is discussed. However, interference from radio carriers to the FMTV signal is small and can be neglected because the radio carriers are located outside the TV receiver noise bandwidth and its power level is small compared to the FMTV signal power level.
E.2.1.3 Shared Transponder - Subcarrier Case

The interference anlaysis for this transponder category is the same as that given in Section 6.l.3. Only the results of the anlaysis are presented here.

It has been shown that a subcarrier carrying TDM'd radio channels located at 6 MHz at baseband has small direct interference on the video signal. A sharp roll-off bandpass filter may be needed for the audio subcarrier which is located at 5 MHz to reject direct interference from the radio carrier.

When intermodulation interference is considered for an FM system with a 4 th-order demodulator input Chebyshev bandpass filter and a hard-limter with $5^{\circ} / \mathrm{AB} A M / P M$ conversion, the weighted video $S / N$ is degraded by 0.75 dB if the radio subcarrier carries one radio channel (see Section 6.1.3.2). If the radio subcarrier carries 4 TDM'd radio channels, the degradation is 1.87 dB . It is also found that both the audio and radio subcarriers are not affected by IM interference since no IM products fall within the neighbourhood of their frequencies ( 5 MHz and 6 MHz respectively).
E.2.2 External Interference - Intrasystem Interference

This section discusses interference from signals in crosspolarized staggered transponders. The transponder frequency arrangement in this case is shown in Figure E.4. The


Figure E.4: DBS 23 MHz transponder frequency plan


Figure E.5: SCPC frequency plan to avoid interference from signals in cross-polarized staggered transponder
figure shows that the co-polarized channel spacing is 26 MHz and the cross-polarized channel spacing is 13 MHz .

## E.2.2.1 Dedicated Transponder Case

It has been shown in Section E.l.l that the system capacity in this transponder category is power limited. Thus, there is more bandwidth available than needed. Therefore, to avoid interference from SCPC signals in the cross-polarized staggered channel, the SCPC carriers can be arranged in such a way that the carriers in the staggered transponder fall into unused frequency slots in the wanted one. This is illustrated in Figure E.5.

Where the traffic in the staggered transponder is wideband such as FMTV, mutual interference between the signals can be avoided by assigning the staggered transponder to a different spot beam.
E.2.2.2 Shared Transponder - FMTV/SCPC Case

A suggested frequency plan for this transponder category is shown in Figure E.6. The figure shows that the FMTV signal is located at the center of the transponder. The radio SCPC carrier is located 1 MHz from the transponder edge (i.e. 10.5 MHz from the center of the transponder). Only one SCPC carrier is shown in the figure since the radio signal may be severely degraded by direct and IM interference from the FMTV'signal if an FMTV carrier and two or more radio carriers are transmitted through the same transponder (see Section E.2.1.2).

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| $\cdots \cdots$ | $\square$ | $\pm$ | $\cdots$ | 0 | － | $\cdots$ |  |  |  |  | $\cdots$ |  |  | $\cdots$ |  | $\square$ | $=$ | $\pm$ |  | $\cdots$ | $\square 7$ | － |  | ＝ |  |  |  | － | $\underline{\square}$ | ＋ |  | $\pm$ | － |  |
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| －3－0． | $\underline{-1}$ | $\cdots$ | $\square$ | $1 \times$ | 0 | 1－ |  |  |  | － |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  | － |  |  |
| $\cdots$ | $\square$ | $\cdots$ | $\square$ | $\cdots$ | $\square$ | － | $\cdots$ |  |  | － |  |  |  | $\underline{\square}$ | $\square$ |  |  |  |  |  | － |  |  |  |  |  |  |  |  |  |  |  | $\bigcirc$ |  |
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|  |  |  |  | $\cdots$ |  | $\cdots$ |  | $\underline{=}$ |  |  | $\cdots$ | ： | $\cdots$ | － |  | $\pm$ | －a | － |  | $\cdots$ | $\square$ |  |  | $\cdots$ | － |  |  |  |  |  |  |  |  |  |
|  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  | － | ． |  |  |  |  |  |  |  |  |  |  |  |  |  |  |
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| － |  | － |  | － |  |  | 1 | ＋ |  |  |  |  |  | ¢ |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |
|  |  | － |  |  |  | $\square$ |  |  |  |  |  |  |  |  | － |  |  | － | $\triangle$ | 7 |  |  |  |  | $\square$ |  |  |  |  |  |  |  |  |  |
|  |  |  |  |  |  | － | －－ | － | － | － | － | $\cdots$ | － |  |  | － |  |  |  | $=$ |  |  |  |  |  |  |  |  |  |  |  |  |  |  |
|  |  | $\ldots$ |  |  |  | $1 \times$ | $\cdots$ |  | $\square$ | － |  |  |  | － |  | － | $\pm$ |  |  | － |  |  | － |  |  |  |  |  |  |  |  |  |  |  |
| E | $\cdots$ | $\cdots$ | $\cdots$ |  |  |  |  | $\bigcirc$ | $\square$ | － |  | － |  |  |  | $\cdots$ | $\square$ |  |  |  | －－ | $\cdots$ |  |  |  |  |  |  |  |  |  |  |  |  |
| ：－ | － | － | ＝ | $\cdots$ | … | $=$ |  | $\square$ | $\square$ | 7： | $\leq$ | 7 | $\square$ | $\square$ | $\pm$ | － | 1 | － | $\pm$ | $\cdots$ | $=$ | $\square$ |  |  |  |  |  |  |  |  |  |  |  |  |
| －－1 |  |  | $\cdots$ |  |  |  | ， |  | $\bigcirc$ | － | － | $\square$ |  | － |  |  |  |  | － |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |
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| －$-\square$ | $\cdots$ | － | $\cdots$ |  |  |  |  | $\cdots$ | $\pm$ | I： |  | $\leq$ | $\pm$ | － |  | － | $\cdots$ | \％： | － | $\square$ | － | $\cdots$ | $\underline{\square}$ |  | $\square$ |  |  | － | $\cdots$ | $\underline{\square}$ | $\because$ | $\pm$ | $\cdots$ |  |
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| $\square$ | $\square$ | \％ | $\square$ | $\square$ |  |  | $\cdots$ |  | － |  | $\cdots$ | $\cdots$ |  |  | － | － | － | F | $\cdots$ | 1. |  |  |  | － | － | $\cdots$ | － | $\cdots$ |  | － |  |  |  |  |
| $\underline{1}$ | －1 | 0 | $\pm$ | ．．． | $\cdots$ | B | $\square$ | I－ |  | $\underline{\square}$ |  | $\cdots$ | E： | 3 | ： | $\underline{\square}$ | － | － | $\square$ | $=$ | $\cdots$ | － | $\pm 1$ | $\square$ | E＝： | $\pm$ | － | $\cdots$ | $\cdots$ | $\square$ | － |  |  |  |
| －F\％： | － | ：$:$ | $\square$ | － | $\cdots$ | － | ： | $=$ |  | － |  | － | F－： | －－ | － | \％ | 4 | 0 | $z=$ | $\square$ | － | － | $\cdots$ | $\square$ | $\pm$ | － | $\square$ | ＋ | $\square$ | －－ | － |  |  |  |
| ． | － | － |  | $\cdots$ |  | －1－1 |  | $\square$ |  |  |  | $\cdots$ | $=$ | － |  | $\square$ | $\square$ | $\cdots$ | $\square$ | $\square$ | ＋ |  | $\cdots$ |  | $\square$ |  | $\square$ | 1：－ | $\square$ | $\cdots$ | $=$ | $\because$ |  |  |
| －－＋ | －： | $\cdots$ | －：－ |  |  |  |  | $=$ |  | ＋－ |  |  |  | － |  |  | $\cdots$ |  | \％ |  | $\square$ |  | $\cdots$ | － |  |  | $\cdots$ |  |  |  |  |  |  |  |
| $\ldots$ | 3 |  |  |  |  |  |  |  | $\because$ | $\cdots$ |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |

The following interference problems are evaluated for the suggested frequency plan: interference from FMTV to FMTV, interference from FMTV to SCPC and interference from SCPC to FMTV. Note that there is no mutual interference between the radio carriers since these carriers do not overlap each other. Also note that in the analysis to follow, only signals falling within the noise bandwidth of the wanted receiver are considered as potential interferors and that the interfering signals serve the same coverage area as the wanted one.

Interference from FMTV to FMTV:

Assume that the FMTV power spectrum is approximately given by:

$$
S(f)=\frac{C_{V}}{14.7} S_{n}(f)
$$

where $C_{V}$ is the total power of the FMTV signal

$$
S_{n}(f)=\left[\begin{array}{ll}
400 e^{.95133 f} ; & -11.5 \leqslant \mathrm{f} \leqslant-6.3 \mathrm{MHz} \\
1.0 & ; \quad|\mathrm{f}| \leqslant 6.3 \mathrm{MHz} \\
400 \mathrm{e}^{-.95133 \mathrm{f}} ; & 6.3 \leqslant \mathrm{f} \leqslant 11.5 \mathrm{MHz} \\
0.0 & ; \text { otherwise }
\end{array}\right.
$$

The TV carrier-to-(TV) interference ratio is then given by:

$$
\frac{C_{V}}{I_{V}}=-10 \log \frac{I}{14.7}\left[\int_{4.0}^{11.5} S_{n}(f) d f+\int_{-11.5}^{-4.0} S_{n}(f) d f\right]-I_{N} d B
$$

where

$$
I_{N}=r+10 \log \left(10^{-\frac{X P D}{10}}+10^{\frac{G_{X}}{10}}\right) d B
$$

with
$r$ is the ratio of the interfering signal power to the wanted signal power at' the receiver input,

XPD is the rain induced cross-polar discrimination, $\mathrm{G}_{\mathrm{X}}$. is the relative cross-polar gain $=-25 \mathrm{~dB}$.

Substituting value of $S_{n}(f)$ into the above equation, $C_{V} / I_{V}$ becomes

$$
\frac{C_{V}}{I_{V}}=3.42-I_{N} d B
$$

The clear weather (i.e. $X P D=\infty d B$ ) $C_{V} / I_{V}$ is 28.42 dB . If converted to equivalent thermal noise, this interference level degrades the FMTV carrier-to-noise density ratio (nominal value $=87 \mathrm{~dB}-\mathrm{Hz}$ ) by 0.17 dB which should be acceptable.

Interference from FMTV to SCPC:

The total FMTV interference power received is given by

$$
I_{V}=\frac{I_{N} C_{R}}{14.7} \int_{D_{f}-\frac{B}{2}}^{D_{f}+\frac{B}{2}} S_{n}(f) d f
$$

where
$I_{N}$ is the interfering signal power normalized to the radio signal power,
$C_{R}$ is the radio signal power,
$D_{f} \quad$ is the frequency spacing between the $T V$ signal (in the staggered transponder) and radio signal center frequencies $=2.5 \mathrm{MHz}$,
$B$ is the noise bandwidth of the radio receiver, $S_{n}(f)$ is the normalized TV power spectrum.

The radio carrier-to-(TV) interference ratio is

$$
\frac{C_{R}}{I_{V}}=-I_{N}-10 \log \frac{1}{14.7}\left[\phi\left(D_{f}+\frac{B}{2}\right)-\phi\left(D_{f}-\frac{B}{2}\right)\right] d B
$$

where

$$
\phi(f)=\int_{-\infty}^{f} S_{n}(f) d f
$$

Thus, with $D_{f}=2.5 \mathrm{MHz}$, the clear weather $C_{R} / I_{V}$ for CATV quality level radio channel with coherent BPSK is 22.9 dB . By converting this interference power to equivalent thermal
noise power, the radio signal $\mathrm{E}_{\mathrm{b}} / \mathrm{N}_{\mathrm{O}}$ is degraded by about 0.45 dB . This degradation is relatively high but can be compensated by increasing the transmitted radio power.

Interference from SCPC to FMTV:

Since the radio carrier falls within the $T V$ receiver noise bandwidth, the $T V$ carrier-to-(radio) interference ratio in clear weather is given by

$$
\frac{C_{V}}{I_{R}}=-I_{N}=-r-G_{x}
$$

where

$$
\begin{aligned}
r= & \text { radio-to-TV power ratio (note that only one } \\
& \text { radio carrier is considered) } \\
G_{X}= & -25 \mathrm{~dB}
\end{aligned}
$$

For example, if the radio signal is for a CATV quality level and if coherent detection is used then $r=-17.9 \mathrm{~dB}$. As a result, $C_{V} / I_{R}=42.9 \mathrm{~dB}$. This low power interferor would cause negligible interference to the FMTV signal.
E.2.2.3 Shared Transponder - Subcarrier Case

For this transponder category, assume that both wanted and interfering signals are of the same type i.e. an FM signal carrying combined TV and radio signals. The interference analysis for this case is exactly the same as that for interference from FMTV to FMTV in the previous section.

Analysis for interference from co-channel signals originated from an adjacent satellite is given in Section 6.2.2. It was found that, for a 1 m antenna (1.80 3 dB beamwidth), two satellites located $2.5^{\circ}$ apart can serve the same service area without causing more than -30 dB of interference power into each other under the clear weather conditions if the satellites transmit orthogonal (cross-polarized) signals. If the satellites transmit signals of the same polarization to the same service area, a spacing between the satellites of about $15^{\circ}$ is needed to achieve the same interference level.

NGUYEN, NAM
Study of digital modulation and multiplexing techniques appropxiate...



[^0]:    *"Study of Radio Program Distribution By Direct Broadcast Satellite Final Report", prepared by Miller Communications Systems Ltd. under DSS Contract No. OST80-00159, October 1981.

[^1]:    *The 625-line European TV system was used.

[^2]:    Table 3.1(b): Issues of Modem Implementation (continued next page).

[^3]:    *Note that for non-bandlimited signals, performance is affected by $\Delta \omega_{0} T$ and therefore frequency offset and detector delay time errors are equivalent.

[^4]:    *These operating points apply for wideband TDMA signals. Section 6.l.l.2 evaluates the optimum operating point of the TWT for the FDMA case.

[^5]:    *FEC coding applies to the information bits. The effect of digital errors on the framing bits was treated in detail in Section 2.2.3.

[^6]:    *Compare this with $33 \%$ transmission rate increase if all the bits are coded.

[^7]:    *Despite the fact that only $m=n / 3$ most significant bits are coded, a code rate of $1 / 2$ increases the transmission rate by $33 \%$ which is still considered too large in the case of a shared transponder where the system capacity may be limited by bandwidth.
    ** For example, a table look-up decoder for the (127, 112) B.C.H. code requires $14 \times 2^{14}$ bits ( 224 kbits ) of memory [37].

[^8]:    *The coding gains shown in this table are obtained using the theoretical BER curves. As a result, CPSK and DQPSK have the same coding gain since the BER curves for these modulations are parallel at high $E_{b} / N_{o}$ values. Measured BER curves should be used in order to get more accurate coding gain values.
    **In Section 4.1, it was stated that the rate $7 / 8$ code represents the near optimum trade-off between bandwidth utilization and codec complexity. However, this is not applicable in this section since the FEC code is applied only to the most significant bits.

[^9]:    *Figures 4.6 and 4.7 show that to obtain maximum $S / N$, the higher the uncoded BER, the larger the number of MSB's that require FEC coding. Since the coded BER is fixed for a given quality level, then the use of a code with large $\mathrm{E}_{\mathrm{b}} / \mathrm{N}_{\mathrm{O}}$ gains implies the required input BER is higher. As a result, more MSB's need to be coded.

[^10]:    *An increase in overall transmission rate due to overhead bits in the multiplexing process is expected. The increase, however, is small so that its effect on the power and bandwidth calculations can be neglected.

[^11]:    *An audio sub-carrier will be considered along with the video signal and radio sub-carrier in Section 5.2.3.

[^12]:    * Note that the same adjacent channel spacing was used in Section 5.0 for SCPC signals for the calculation of bandwidth requirements.

[^13]:    * This is the input backoff for which the satellite transponder can carry a maximum number of SCPC carriers each meets the required $\mathrm{C} / \mathrm{N}_{\mathrm{O}}$.

[^14]:    * IM interference reduction due to random arrangements have been accounted for in the link calculation by the last term in (6.3).

[^15]:    Fote that the discussion applies to only a small number of radio SCPC carriers. Table 6.7 shows that for radio-to-TV power ratio of -8 dB , the optimum operating input backoff is 4 dB . The available radio $\mathrm{C} / \mathrm{N}_{\mathrm{O}}$ in this case is 78 dB which can support about 7 CATV-quality radio (CPSK or CMSK) carriers.

[^16]:    * As calculated in Section 6.1.3.1, the video signal-tounweighted noise ratio is 31.6 dB . With a de-emphasis and weighting advantage of 13 dB , the weighted $\mathrm{S} / \mathrm{N}$ is 44.6 dB .

[^17]:    *Anik-C has not been addressed because the frequency plan was not specified.

[^18]:    * A spacing of about 13 MHz is required to reduce direct interference from FMTV to SCPC as suggested in Section 6.1.2.1.

[^19]:    * The received $T V C / N_{o}$ is calculated as $(87-A) d B-H z$ where $A$ is the rain attenuation. The $\mathrm{C} / \mathrm{N}_{0}$ degradation is given by $\log \left(1+\frac{1}{18 \mathrm{MHz}} \frac{\mathrm{C} / \mathrm{N}_{\mathrm{O}}}{\mathrm{C}_{\mathrm{V}} / \mathrm{I}_{\mathrm{V}}}\right)$

[^20]:    * Clear weather $C_{R} / I_{V}$ is calculated by using (6.24) and (6.22). In the second equation $r=-\beta$ (in $d B$ ) where $\beta$ is the SCPC-to-TV power ratio which is listed along with the radio receiver noise bandwidth $B$ in Table 6.4.

[^21]:    * Note that the external interference analysis was performed with the transponder frequency plan as shown in Figure 6.l3 (DBS case). Thus the remarks to follow should apply to this case only.

[^22]:    *including 2 dB modem implementation margin for BPSK (2.5 dB for. DMSK) and FEC coding gain.

    Table 7.1: Radio subsystem specifications for dedicated transponder - SCPC and TDM carriers

[^23]:    *Recall that the required (non-weighted) $T T / N$ 's are 50 dB , 47 dB and 36 dB for CATV quality, high quality and low quality, respectively.

[^24]:    Figure B.2: Number of radio channels plotted against earth station receive antenna G/T for ANIK C satellite (dedicated transponder case)

[^25]:    * It is assumed that the bandwidth available for the radio carriers in the DBS case is 3 MHz (see Figure 6.20). In the ANIK C case, there is plenty of $R F$ bandwidth for radio traffic since the FMTV signal occupies only 22 MHz of the 54 MHz transponder bandwidth.

[^26]:    * R.G. Medhurst and J.H. Roberts, "Evaluation of Distortion in FM Trunk Radio Systems by a Monte Carlo Method", Proc. IEE (G.B.), April 1966, pp. 570-580.

[^27]:    * Reference: "Coherent FM Carrier Recovery", MCS File No. 8301-2

[^28]:    * Carson's rule bandwidth $=2(6 \mathrm{MHz}+5 \mathrm{MHz})=22 \mathrm{MHz}$ where 6 MHz is the peak frequency deviation and 5 MHz is the TV associated audio subcarrier frequency.

[^29]:    * With a 27 MHz transponder, 18 radio channels can be transmitted.

