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**The Effects of Non-Ideal Filtering and
Nonlinear Amplification
on the BER Performance of a
COFDM-based DAB System**

by

Daniel W. Burke

**A thesis submitted to the
Faculty of Graduate Studies and Research
in partial fulfilment of the requirements
for the degree of
Master of Engineering**

**Ottawa-Carleton Institute for Electrical Engineering
Faculty of Engineering
Department of Systems and Computer Engineering
Carleton University**

May 1994

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**"The Effects of Non-Ideal Filtering
and Nonlinear Amplification
on the BER Performance of a
COFDM-based DAB System"**

submitted by

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in partial fulfilment of the requirements
for the degree of Master of Engineering

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May 1994

Abstract

This document discusses an investigation into the effects of non-ideal filtering and nonlinear amplification on the performance of a digital audio broadcasting (DAB) system operating at 1.5 GHz while transmitting over both Rayleigh and Ricean fading channels. The experimental system was based on the coded orthogonal frequency division multiplexing (COFDM) transmission scheme and was similar to a system presently being considered by the Radiocommunications Bureau of the International Telecommunications Union for a world standard in DAB. The non-ideal filters were incorporated into both the transmit and receive equipment, while the nonlinear amplifier was only included in the transmitter.

An extensive filtering study was performed on both finite impulse response (FIR) and infinite impulse response (IIR) filter designs. The study concluded that nonlinear phase responses and passband ripples less than 1 dB in magnitude in higher order filters had little effect on the performance of the system. From these conclusions, it was recommended that of the filter designs tested, an 8th order Chebychev IIR filter with 1 dB ripples would be the most appropriate choice for such a DAB system.

Secondly, the system losses caused by nonlinear amplification were determined. The satellite amplifiers studied included a travelling wave tube amplifier (TWTA) and a L-Band solid state power amplifier built by SPAR. The terrestrial amplifier considered was a solid state UHF design manufactured by Larcan. The optimum operating point of each amplifier was determined.

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List of Abbreviations

AWGN	Additive White Gaussian Noise
AM	Amplitude Modulation
BER	Bit Error Rate
BPSK	Binary Phase Shift Keying
BU	Bad Urban
CCETT	Centre Commun d'Etudes de Télédiffusion et Télécommunications
CD	Compact Disc
CDF	Cumulative Distribution Function
COFDM	Coded Orthogonal Frequency Division Multiplexing
CRC	Communications Research Centre
DAB	Digital Audio Broadcasting
DFT	Discrete Fourier Transform
DQPSK	Differential Quadrature Phase Shift Keying
FDM	Frequency Division Multiplexing
FFT	Fast Fourier Transform
FIR	Finite Impulse Response
FM	Frequency Modulation
GSM	Groupe Speciale Mobile
HT	Hilly Terrain
IBO	Input Backoff
IDFT	Inverse Discrete Fourier Transform
IFFT	Inverse Fast Fourier Transform
IIR	Infinite Impulse Response
ISI	Inter-Symbol Interference
ISO	International Standards Organization
ITU-BR	International Telecommunications Union - Radiocommunications Bureau
L-Band	Radio Frequency Band from 1 GHz to 2 GHz
MSAT	Mobile Satellite
OFDM	Orthogonal Frequency Division Multiplexing
pdf	Probability Density Function
QAM	Quadrature Amplitude Modulation
QPSK	Quadrature Phase Shift Keying
SAT	Satellite
SNR	Signal to Noise Ratio
SSPA	Solid State Power Amplifier
RA	Rural Area
TU	Typical Urban
TWTA	Travelling Wave Tube Amplifier
UHF	Ultra High Frequency (Radio Frequency Band from 300 MHz to 3 GHz)
WARC	World Administrative Radio Conference
WFIR	Windowed Finite Impulse Response

List of Symbols

B_c	Coherence Bandwidth
B_d	Doppler Spread
B_i	Information Rate
B_n	Noise Bandwidth
$c_{i,k}$	QPSK Channel Symbol
$C_{i,k}$	DQPSK Channel Symbol
Δ	Guard Interval
E_b/N_0	Bit Energy to Noise Power Ratio
ε	Expected value
f_c	Carrier Frequency
K	Ricean Fading Channel Power Ratio
N_c	Number of Carriers
N_g	Number of Samples in the Guard Interval
T_c	Coherence Time
T_m	Multipath Spread
T_s	Useful Symbol Interval
T_s^*	Transmitted Symbol Interval
τ	Propagation Delay
λ	Doppler Frequency
λ_{dp}	Direct Path Doppler Frequency
λ_{max}	Maximum Doppler Frequency
$c(\tau;t)$	Channel Impulse Response Lowpass Equivalent Signal
$r(t)$	Received Lowpass Equivalent Signal
$s(t)$	Transmitted Bandpass Signal
$u(t)$	Transmitted Lowpass Equivalent Signal
$v(t)$	Complex Envelope of the Transmitted Bandpass Signal
$x(t)$	Received Bandpass Signal
$y(t)$	Complex Envelope of the Received Bandpass Signal

Chapter 1

INTRODUCTION

With the increasing demand for quality audio broadcasting and the crowding of both the AM and FM bands, an alternative to conventional analog audio broadcasting has become necessary. The Eureka 147 DAB system, a COFDM-based DAB system, is presently being proposed for world standardization as this alternative broadcasting system. Designed with the mobile receiver in mind, it promises "CD-like" sound quality because of its digital implementation and its extensive anti-fading techniques.

1.1 Thesis Objective

While much research has been completed on the performance of COFDM-based DAB systems like the Eureka 147 DAB system, little of the work completed has taken into account the very realistic impairments that exist in conventional broadcasting systems. These impairments can include non-ideal filtering, nonlinear amplification and synchronization problems in the receiver. It is the objective of this thesis to study the performance degradation suffered by a COFDM-based DAB system transmitting over Rayleigh and Ricean fading channels when it is exposed to non-ideal filtering and nonlinearities during amplification.

1.2 Thesis Organization

This document is divided into four primary chapters. Chapter 2 will provide the reader with most of the background required to understand the results presented in this work. It reviews the characterization of dispersive multipath channels, introduces the fundamental mathematical theory behind the orthogonal frequency division multiplexed (OFDM) signals used by the experimental DAB system as well as provides a literature survey of publications relevant to this thesis.

Chapter 3 contains the specifications of the DAB system. It provides a brief description of the distributed emission technique used, an explanation of the system's different transmission modes and an extensive discussion of the simulation model that was implemented. Included within this description of the simulation model are the procedures that were used to verify its operation.

Chapter 4 presents the results obtained during the research component of this thesis. It begins by describing the manner in which the DAB system's performance was measured and then proceeds to display and discuss the experimental results. These results are broken down into four principal sections, namely, ideal system results, non-ideal filtering results, nonlinear amplifier results, and non-ideal filtering and nonlinear amplification results.

Chapter 5 reviews the work that was completed and makes several conclusions based upon the outcome of the research. It also makes recommendations as to what further research would be worthwhile.

1.3 Thesis Contributions

This thesis performs a study on the effects of non-ideal filtering and nonlinear amplification on the performance of a COFDM-based DAB system operating at 1.5 GHz. A computer simulation program has been developed that measures the BER performance of a COFDM-based DAB system over both Rayleigh and Ricean fading channels. The DAB system examined can be treated as ideal, or it can incorporate non-ideal filters in the transmitter and receiver (either FIR or IIR), a nonlinear amplifier in the transmitter, or both. The contributions of this thesis can be summarized as follows:

- 1) A comprehensive study of the effects of both FIR and IIR filters on the performance of a COFDM-based DAB system was completed. It was learned that the nonlinear phase responses of the IIR filter designs considered had a negligible effect on the DAB system, indicating that comparable FIR designs would not be required. In addition, it was demonstrated that the impact of filter passband ripples could be quite significant but this impact was minimized if they were limited to 1 dB in magnitude and higher order filters were used. These observations led to the conclusion that of all of the filters investigated, an 8th order Chebychev IIR filter with 1 dB ripples would be the most appropriate choice for use with the experimental DAB system.
- 2) The losses experienced by a COFDM-based DAB system as a result of nonlinear amplification in the transmitter were determined for typical satellite and terrestrial

amplifiers. The optimum operating point of each of the three amplifiers examined (a TWTA, a SPAR L-Band SSPA and a Larcen UHF SSPA) was evaluated.

- 3) It was demonstrated that the losses introduced by non-ideal filtering and nonlinear amplification are not purely additive. In most of the cases examined, the loss introduced by both components was greater than the sum of the losses caused by each component individually.

A more detailed examination of these findings and recommendations as to what further research in this field would be worthwhile can be found in chapter 6.

Chapter 2

BACKGROUND

Section 2.1 is provided for those readers who may not be particularly familiar with the theory behind modelling multipath channels. It is basically a summary of section 7.1 from [1]. Section 2.2, most of which was taken from [2], introduces the OFDM signals used by the experimental DAB system.

2.1 Characterizing Multipath Channels

It is common in communication theory to treat most channels as a single path, AWGN channel. However, in the case of mobile communications, there is often more than one path between the transmitter and the receiver. This situation occurs as a result of the transmitted signal reflecting off various objects between the transmitter and the receiver. The received signal then becomes the sum of the transmitted signal along all of these paths. These "multipaths" can add either constructively, increasing the envelope of the received signal, or destructively, decreasing the envelope of the received signal. This process is referred to as signal fading.

When there is no line-of-sight path between the transmitter and the receiver, the received envelope level has a Rayleigh distribution and the channel is called a Rayleigh fading channel. However, when a line-of-sight path exists, the received envelope follows a Ricean distribution and the channel is referred to as a Ricean fading channel.

Let us consider a transmitted bandpass signal that can be represented by

$$s(t) = \text{Re} \{ u(t)e^{j2\pi f_c t} \} \quad (2.1.1)$$

where $u(t)$ is the lowpass equivalent signal of $s(t)$ and f_c is the carrier frequency used during transmission. Assuming that there is a discrete number of paths between the transmitter and the receiver, the received bandpass signal would be

$$x(t) = \sum_n \alpha_n(t) s[t - \tau_n(t)] \quad (2.1.2)$$

where $\alpha_n(t)$ and $\tau_n(t)$ are the attenuation and the propagation delay along the n^{th} path, respectively, at time t . Substituting for $s(t)$ yields,

$$x(t) = \text{Re} \{ (\sum_n \alpha_n(t) e^{j2\pi f_c \tau_n(t)} u[t - \tau_n(t)]) e^{j2\pi f_c t} \} \quad (2.1.3)$$

It can be seen from (2.1.3) that the received lowpass equivalent signal is given by the expression

$$r(t) = \sum_n \alpha_n(t) e^{j2\pi f_c \tau_n(t)} u[t - \tau_n(t)] \quad (2.1.4)$$

From comparing the expressions for $u(t)$, the transmitted signal, and $r(t)$, the received signal, the lowpass equivalent of the time-varying channel's impulse response can be identified as

$$c(\tau; t) = \sum_n \alpha_n(t) e^{j2\pi f_c \tau_n(t)} \delta[\tau - \tau_n(t)] \quad (2.1.5)$$

As well, if we assume that $c(\tau; t)$ is wide sense stationary, several important

characteristics of the multipath channel can be derived.

First, its autocorrelation function would be specified by

$$\phi_c(\tau_1, \tau_2; \Delta t) = \frac{1}{2} \varepsilon \{ c^*(\tau_1; t) c(\tau_2; t + \Delta t) \} \quad (2.1.6)$$

Under most circumstances, the attenuation and propagation delay along one path are independent of those along another path, so incorporating this into the expression of (2.1.6) we get,

$$\frac{1}{2} \varepsilon \{ c^*(\tau_1; t) c(\tau_2; t + \Delta t) \} = \phi_c(\tau_1; \Delta t) \delta(\tau_1 - \tau_2) \quad (2.1.7)$$

At $\Delta t = 0$, $\phi_c(\tau; 0)$ is commonly written as $\phi_c(\tau)$, which is the average power output of the channel as a function of propagation delay. As a result, $\phi_c(\tau)$ is called the *multipath intensity profile*, or the *delay power spectrum* of the channel. The *multipath spread* of the channel is the range of τ for which $\phi_c(\tau)$ is nonzero and is denoted by T_m .

A similar relationship can be found in the frequency domain. The Fourier transform of $c(\tau; t)$ with respect to the delay τ is

$$C(f; t) = \int_{-\infty}^{+\infty} c(\tau; t) e^{j2\pi f\tau} d\tau \quad (2.1.8)$$

Since the Fourier transform is a linear operation, $C(f; t)$ will have the same statistical properties as $c(\tau; t)$. Therefore, its autocorrelation function can be defined by

$$\phi_C(f_1, f_2; \Delta t) = \frac{1}{2} \varepsilon \{ C^*(f_1; t) C(f_2; t + \Delta t) \} \quad (2.1.9)$$

By substituting (2.1.8) into (2.1.9), the autocorrelation function can be simplified so that

it is a function of the frequency difference ($\Delta f = f_1 - f_2$) only. It is given by

$$\phi_c(f_1, f_2; \Delta t) = \int_{-\infty}^{+\infty} \phi_c(\tau; \Delta t) e^{j2\pi \Delta f \tau} d\tau \quad (2.1.10A)$$

where $\phi_c(f_1, f_2; \Delta t) = \phi_c(\Delta f; \Delta t)$ (2.1.10B)

The expression $\phi_c(\Delta f; \Delta t)$ is referred to as the *spaced-frequency, spaced-time correlation function* of the channel. This indicates both the time and frequency variations of the channel.

By setting $\Delta t = 0$ in (2.1.10A), then the relationship between $\phi_c(\Delta f; 0) = \phi_c(\Delta f)$ and $\phi_c(\tau; 0) = \phi_c(\tau)$, becomes the Fourier transform with respect to delay, namely

$$\phi_c(\Delta f) = \int_{-\infty}^{+\infty} \phi_c(\tau) e^{j2\pi \Delta f \tau} d\tau \quad (2.1.11)$$

As a result, the inverse of the multipath spread of the channel is called the *coherence bandwidth* of the channel, or

$$B_c = 1/T_m \quad (2.1.12)$$

This indicates that if two signals that have a frequency separation greater than B_c and are transmitted through the same channel simultaneously, each will be affected independently.

By now focusing on the time variations of the channel, its Doppler broadening can be characterized. The Doppler effects can be related to time variations using the Fourier transform of $\phi_c(\Delta f; \Delta t)$. We define the Fourier transform of $\phi_c(\Delta f; \Delta t)$ with respect to Δt as

$$S_c(\Delta f; \lambda) = \int_{-\infty}^{+\infty} \phi_c(\Delta f; \Delta t) e^{j2\pi\lambda\Delta t} d\Delta t \quad (2.1.13)$$

With $\Delta f = 0$ and $S_c(0; \lambda) = S_c(\lambda)$, (2.1.13) can be written as

$$S_c(\lambda) = \int_{-\infty}^{+\infty} \phi_c(\Delta t) e^{j2\pi\lambda\Delta t} d\Delta t \quad (2.1.14)$$

$S_c(\lambda)$ is a power spectrum that specifies the intensity of the signal as a function of the Doppler frequency λ . As a result, $S_c(\lambda)$ is called the *Doppler power spectrum* of the channel. The range of λ for which $S_c(\lambda)$ is nonzero is called the *Doppler spread*, B_d , of the channel.

Since $\phi_c(\Delta t)$ is the Fourier transform of $S_c(\lambda)$, the inverse of B_d is called the *coherence time* of the channel. That is,

$$T_c = 1/B_d \quad (2.1.15)$$

Now, any two signals that are transmitted through the same channel more than T_c seconds apart will be effected by the channel independently.

In order to obtain one expression that will provide a measure of the average power as a function of both propagation delay τ and Doppler frequency λ , a new function $S(\tau; \lambda)$ is defined by the double Fourier transform (i.e. over both Δt and Δf)

$$S(\tau; \lambda) = \int_{-\infty}^{+\infty} \int_{-\infty}^{+\infty} \phi_c(\Delta f; \Delta t) e^{j2\pi\lambda\Delta t} e^{j2\pi\tau\Delta f} d\Delta t d\Delta f \quad (2.1.16)$$

This new function is referred to as the *scattering function* of the channel and acts as a

three dimensional representation of the channel's power spectrum.

2.1.1 Types of Signal Fading

There can be several different manners in which a signal can fade, all of which depend upon the signal bandwidth as well as the multipath channel characteristics.

When the signal bandwidth is less than the coherence bandwidth of the channel, it is said to be a *frequency non-selective* channel. In this case, the signal is not severely distorted by the channel as the same channel response affects the entire signal. If the signal bandwidth is larger than the coherence bandwidth of the channel, it is said to be a *frequency selective* channel. This can cause severe signal distortion as the channel's effect on some sections of the signal is completely independent of its effect on other sections of the signal. However, this signal distortion leads to an inherent frequency diversity in the channel and often results in performance superior to that obtained over a frequency non-selective channel.

Similarly, if the transmitted symbol interval is much smaller than the coherence time of the channel, then the channel is said to be *slowly fading*. This means that the channel response is approximately the same for consecutive symbols. However, if the transmitted symbol interval is much larger than the channel's coherence time, then the channel is called a *fast fading* channel. In this case, one cannot assume that the channel response will be the same from symbol to symbol.

By convention, channels that are both frequency non-selective and slowly fading are referred to as *flat fading channels*.

2.2 OFDM Signals

With traditional FDM systems, data is multiplexed onto different channels which have non-overlapping spectra. The primary drawback of this technique is the excessive bandwidth that may be required in order to ensure that the spectra of adjacent channels do not overlap. The technique of OFDM is similar in that the transmitted information is frequency multiplexed onto several channels but rather than having non-overlapping spectra, the channels are only required to have spectra that are orthogonal to all of the other channels used. This greatly reduces the amount of bandwidth required by the system and can lead to much simpler modern implementations[3].

An advantage of using OFDM in mobile applications is its inherent resistance to the inter-symbol interference introduced by the delay spreading of the channel. With OFDM, the transmitted information sequence is carried by a large number of orthogonal, low bit rate subcarriers rather than a single, high bit rate carrier. The resulting spectral spreading can cause many mobile channels to become frequency selective in nature[2] which in turn improves the system's performance because of the frequency diversity inherent to the channel[1]. With many other modulation techniques, the gains attributed to the frequency selectivity of the channel can be offset by the ISI introduced into the received signal by this selectivity. However, the long symbol duration of the low bit rate subcarriers in an OFDM signal greatly reduces the effects of this ISI. As well, incorporating a guard interval (see section 2.2.2) into each subcarrier can often eliminate the ISI[2], making OFDM a robust and highly effective modulation technique for combatting the time varying characteristics of many mobile channels.

2.2.1 OFDM Modulation

A set of N_c frequencies is selected using

$$f_k = f_0 + k/T_s \quad ; k = 0 \text{ to } N_c-1 \quad (2.2.1)$$

where T_s represents the useful symbol duration and f_0 is the lowest frequency to be used. This allows N_c basis functions that are orthogonal over the interval $[0, T_s]$ to be defined by

$$\psi_{ik}(t) = g_k(t - iT_s) \quad ; k = 0 \text{ to } N_c-1, i = -\infty \text{ to } +\infty \quad (2.2.2)$$

$$g_k(t) = \begin{cases} \exp(j2\pi f_k t) & ; 0 \leq t < T_s \\ 0 & ; \text{otherwise} \end{cases} \quad (2.2.3)$$

Using these orthogonal basis functions, the transmitted bandpass OFDM signal can be written as

$$s(t) = \sum_{i=-\infty}^{\infty} \sum_{k=0}^{N_c-1} C_{ik} \psi_{ik}(t) \quad (2.2.4)$$

where $\{ C_{ik} \}$ are the channel symbols that are being transmitted during the i^{th} symbol interval on the k^{th} subcarrier.

2.2.2 The Guard Interval

When inter-symbol interference is introduced into the received signal by the channel (i.e. a frequency selective channel), the assumption that all of the individual subcarriers are orthogonal no longer holds. This often causes a degradation in the

performance of the OFDM system, but this degradation can be virtually eliminated through the use of a guard interval[2]. By extending the transmitted symbol duration, T_s' , along each subcarrier by a guard interval of length Δ , all of the ISI introduced by the channel will be absorbed during this interval if Δ is greater than the multipath spread of the channel. This leaves the remaining useful symbol interval, T_s ($T_s' = T_s + \Delta$), free of ISI. The receiver would then intentionally ignore any signal received during the guard interval and all of the negative effects of the ISI would be removed from the OFDM system.

Considering the use of a guard interval, the representation of the transmitted signal can be updated. The basis functions used by the receiver would then become

$$\psi_{ik}(t) = g_k(t - iT_s') \quad ; k = 0 \text{ to } N_c-1, i = -\infty \text{ to } +\infty \quad (2.2.5)$$

and the transmitted signal would have the basis functions

$$\psi_{ik}'(t) = g_k'(t - iT_s') \quad ; k = 0 \text{ to } N_c-1 \text{ and } i = -\infty \text{ to } +\infty \quad (2.2.6)$$

$$g_k'(t) = \begin{cases} \exp(j2\pi f_k t) & ; -\Delta \leq t < T_s \\ 0 & ; \text{otherwise} \end{cases} \quad (2.2.7)$$

allowing the transmitted bandpass signal to be represented as

$$s(t) = \sum_{i=-\infty}^{\infty} \sum_{k=0}^{N_c-1} C_{ik} \psi_{ik}'(t) \quad (2.2.8)$$

The basic structure of an OFDM signal including a guard interval can be seen in figure 2.1 while its power spectrum is shown in figure 2.2.

Figure 2.1: The basic features of an OFDM signal

Figure 2.2: The power spectrum of an OFDM signal

2.2.3 OFDM Modulation Using the IDFT

Consider that the signal $s(t)$ for any given useful symbol interval (i constant) can be written as

$$s(t) = \sum_{k=0}^{N_c-1} C_k \psi_k(t) \quad ; 0 \leq t < T_s \quad (2.2.9)$$

or
$$s(t) = \sum_{k=0}^{N_c-1} C_k \exp(j2\pi f_k t) \quad ; 0 \leq t < T_s \quad (2.2.10)$$

By substituting (2.2.1) in for f_k , the expression for $s(t)$ can be rewritten in terms of a carrier frequency, f_o , and its complex envelope, $v(t)$, as

$$s(t) = \exp(j2\pi f_o t) \sum_{k=0}^{N_c-1} C_k \exp(j2\pi k t / T_s) \quad ; 0 \leq t < T_s \quad (2.2.11)$$

where
$$v(t) = \sum_{k=0}^{N_c-1} C_k \exp(j2\pi k t / T_s) \quad ; 0 \leq t < T_s \quad (2.2.12)$$

Next, by considering the complex envelope of $s(t)$ as a discrete sequence sampled at intervals of $T = T_s / N_c$, it can be expressed as

$$v(nT) = \sum_{k=0}^{N_c-1} C_k \exp(j2\pi k n / N_c) \quad ; 0 \leq n < N_c \quad (2.2.13)$$

It can be seen from (2.2.11) and (2.2.13) that OFDM modulation can be performed using an N_c point inverse discrete Fourier transform (IDFT) followed by a complex multiplier.

To insert the guard interval when an IDFT is being used in the transmitter, the last

N_g ($N_g = \Delta/T$) samples of the N_c point IDFT output sequence must be transmitted before the entire N_c samples of the output sequence. Therefore, a total of $N_c + N_g$ samples would be required for each transmitted symbol interval.

2.2.4 OFDM Demodulation

When interpreting the received OFDM signal, it is important to remember that in multipath channels only the interval $[0, T_s]$ should be utilized in the decision making process. Providing that $T_m < \Delta$, we can treat the channel as memoryless as all of the ISI will have been removed from the signal. The received signal, $x(t)$, can then be written for any given symbol interval as:

$$x(t) = \sum_{k=0}^{N_c-1} H_k C_k \psi_k(t) \quad ; 0 \leq t < T_s \quad (2.2.14)$$

$$\text{or} \quad x(t) = \sum_{k=0}^{N_c-1} H_k C_k \exp(j2\pi f_k t) \quad ; 0 \leq t < T_s \quad (2.2.15)$$

assuming that during the given symbol interval, the complex channel response on the k^{th} subcarrier can be represented as $H_k = \rho_k \exp(j\phi_k)$.

By rewriting equation (2.2.15) in terms of a carrier frequency, f_c , and the complex envelope of the received signal, $y(t)$, it becomes

$$x(t) = \exp(j2\pi f_c t) \sum_{k=0}^{N_c-1} H_k C_k \exp(j2\pi k t/T_s) \quad ; 0 \leq t < T_s \quad (2.2.16)$$

where
$$y(t) = \sum_{k=0}^{N_c-1} H_k C_k \exp(j2\pi kt/T_c) \quad ; 0 \leq t < T_c \quad (2.2.17)$$

By mixing $x(t)$ down to baseband so that only $y(t)$ remains, $y(t)$ can be sampled at a rate of $T = T_c/N_c$ yielding

$$y(nT) = \sum_{k=0}^{N_c-1} H_k C_k \exp(j2\pi kn/N_c) \quad ; 0 \leq n < N_c \quad (2.2.18)$$

By observing that the relationship between $y(nT)$ and $H_k C_k$ in (2.2.18) is the IDFT, it can be concluded that the received value on each subcarrier, $H_k C_k$, can be recovered by taking the discrete Fourier transform (DFT) of the sampled complex envelope of the received signal, as demonstrated in equation (2.2.19).

$$H_k C_k = (1/N_c) \sum_{n=0}^{N_c-1} y(nT) \exp(-j2\pi kn/N_c) \quad ; 0 \leq k < N_c \quad (2.2.19)$$

This now allows the recovery of the transmitted data with a maximum likelihood sequence estimator or a simple threshold element.

2.3 Literature Survey

One of the first papers to discuss the use of the DFT as a means of modulating FDM signals was the work by Weinstein and Ebert in 1971[3]. Although it only dealt with transmission in the AWGN environment, it produced a much simpler FDM data

modem that no longer required banks of subcarrier oscillators and demodulators. While their modulation technique was not explicitly referred to as OFDM, it was the start of much research that eventually lead to the OFDM systems of present.

In the early 1980's, Hirosaki studied an orthogonally multiplexed QAM modem for data communication in several papers [4]-[6]. His earlier studies emphasized its use in an AWGN channel, but some of his later work incorporated impulsive noise disturbances as well. Other papers investigating various OFDM data communication systems can be found in Cimini's 1985 study of an OFDM system in a mobile radio channel[7], and the work of Casas, Leung and Warner regarding a hybrid transmission technique, OFDM/FM, in [8] and [9].

OFDM's use in digital audio broadcasting applications became apparent in 1987 with the publication of the CCETT's paper "Principles of modulation and channel coding for digital broadcasting for mobile receivers"[2]. In this paper, the authors, Lassalle and Alard, outlined the OFDM-based system that was being developed by the CCETT for use in DAB. With this paper and several to follow [10]-[13], the CCETT developed a transmission scheme, COFDM, for the purpose of broadcasting compressed audio signals to fixed, portable and vehicular receivers. This transmission scheme incorporated channel coding techniques with an OFDM modulator and it has been combined with the ISO Layer II audio source coding system to yield a complete DAB system known as the Eureka 147 DAB system[14]. This system has since been proposed to the ITU-BR as a world standard for DAB which has resulted in the ITU-BR making two recommendations (Rec. 774 and 789) describing the requirements to be satisfied in a world standard for

DAB. The Eureka 147 DAB system is currently the only field tested system to fulfil these requirements[14].

In 1990, the Canadian government, in collaboration with the Canadian broadcast industry, initiated a program to assess the performance of the Eureka 147 DAB system. As part of this assessment, the staff at the Communications Research Centre began to investigate a COFDM-based system that was very similar to the Eureka system. Some preliminary results concerning the time interleaving depth, the channel code constraint length, the number of receiver quantization levels and the useful symbol duration length, were presented at the 47th Annual Broadcast Engineering Conference (NAB93) in April 1993[15]. These results as well as others discussing the system's performance over a satellite channel were submitted as a technical paper to Study Group 10 of the ITU-BR[16]. Another paper submitted to the same study group[17], provided an introduction to the distributed emission concept proposed for use with a COFDM-based DAB system. In addition, an extensive system description and some of its performance results were contained in an internal CRC report[18].

Researchers of the German Aerospace Research Establishment have also contributed to the pool of literature regarding OFDM applications in DAB. In 1991, Hoehner et al proposed the use of a rate-compatible punctured convolutional (RCPC) channel coding scheme[19], while a paper by Rapp[20] discussed the effects of nonlinear amplification on OFDM signals over AWGN channels. In Rapp's paper, only the nonlinear characteristics of a TWTA and theoretical approximations of solid state power amplifiers were considered.

Other miscellaneous papers relevant to OFDM research would be Bingham's paper on multicarrier modulation[21], Tüsel and Kammeyer's carrier recovery scheme[22] and Feig's study of FDM system power limitations[23].

2.4 Problem Statement and Definition

All of the research outlined in section 2.3 regarding both the Eureka 147 DAB system and CRC's COFDM-based DAB system has concentrated on the theoretical aspects of implementing either system. As a result, all of the published performance curves indicate a best case performance limit as they ignore all of the conventional hardware related impairments that an operating broadcast system would be subjected to. While the effect of nonlinear amplification on OFDM signals was originally investigated by Rapp in [20], he only considered the performance of a generic OFDM system over an AWGN channel. Since his study did not incorporate channel coding techniques or time varying channel models, it is not conclusive as to whether or not his results will apply to a COFDM-based DAB system serving vehicular or portable receivers. In addition, he made assumptions about the phase distortion characteristics of SSPAs which specifications from product manufacturers[24][25] would indicate to be false.

Therefore, it is the objective of this thesis to determine the degradation in performance that a COFDM-based DAB system experiences when it is exposed to non-ideal filtering and nonlinear amplification. The investigation will encompass a system transmitting over both Rayleigh and Ricean fading channels, and the amplifier models utilized will consider both power and phase distortion according to the specifications

provided by amplifier manufacturers. Initially, these two topics will be investigated independently but later study will consider the joint effects of both non-ideal filtering and nonlinear amplification. This work will be divided up as follows:

- 1) **Non-ideal Filtering** - Typical FIR and IIR filter designs will be introduced into both the transmitter and the receiver of the simulated DAB system. The effects of a nonlinear phase response, passband ripples, and the steepness of the transition band on the system's performance will be quantified. As well, a recommendation regarding which of the filters examined would be the most appropriate choice for use with the DAB system will be made.
- 2) **Nonlinear Amplification** - The nonlinear characteristics of typical satellite and terrestrial high power amplifiers will be added to the transmitter of the DAB system simulation. The tradeoffs between performance degradation and the required input backoff for each of the amplifier models will be determined. From these results, the optimum operating point of each amplifier will be evaluated.
- 3) **Non-ideal Filtering and Nonlinear Amplification** - The experimental DAB system will be subjected to joint effects of non-ideal filtering and nonlinear amplification. The filter recommended in study 1 will be used in

conjunction with the amplifiers of study 2. Measurements will be made to determine if the losses introduced by these components are independent of each other and whether or not the optimum operating point of each amplifier evaluated here is consistent with those evaluated in 2.

The motivations for this research were two fold. First, the ITU has yet to make a final decision on a world standard for DAB and although previous research has indicated that a COFDM-based system would be an appropriate choice[2][10]-[20], this work may provide insight into performance limitations that have yet to be considered. Secondly, should a COFDM-based system become the world standard for DAB, this information would allow future system designers to have more realistic performance estimates that might otherwise have not been available to them. The results of this work could be incorporated into more accurate estimates of broadcast coverage and transmission link budgets.

Chapter 3

THE COFDM-BASED DAB SYSTEM

The purpose of chapter 3 is to introduce the reader to the experimental COFDM-based DAB system. Section 1 provides a summary of the system specifications while section 2 discusses the model of the system which was used during the computer simulation.

3.1 System Specifications

3.1.1 DAB Spectrum Allocation

During the World Administrative Radio Conference in February of 1992 (WARC '92), 40 MHz of spectrum at 1.5 GHz (1.452 GHz - 1.492 GHz) was allocated worldwide for use in satellite DAB and complementary terrestrial services[16]. As a result, all of the research included in this thesis has concentrated on a system operating in that frequency range.

3.1.2 System Operating Parameters

The Eureka 147 DAB system description provided in [14] allowed for operation in three distinctly different transmission modes (I, II and III). Each mode used the same transmission scheme (COFDM), but differed in the length of the useful symbol duration,

the length of the guard interval and the recommended operating frequency range. According to [14], only modes II and III were suitable for use at 1.5 GHz, so operation in mode I was not investigated. The system operating parameters for modes II and III are summarized below in table 3.1.

Parameter		Value
System Bandwidth		1.536 MHz
Guard Interval Duration (Δ)	mode II mode III	62.5 μ s 31.25 μ s
Useful Symbol Duration (T_s)	mode II mode III	250 μ s 125 μ s
Number of Subcarriers	mode II mode III	384 192
Subcarrier Spacing	mode II mode III	4 kHz 8 kHz
Modulation		DQPSK
Demodulation		differential
Convolutional Code	rate (r) constraint length (L) decoder soft decision levels (Q)	variable 7 Viterbi 16

Table 3.1: The Eureka 147 DAB system operating parameters

3.1.3 Broadcast Coverage - Distributed Emission

The requirement for acceptable system coverage has been proposed to be providing good reception in at least 99% of the locations for at least 99% of the time[26]. The difficulty associated with using only a single transmitter for broadcasting COFDM signals is achieving this requirement[26][27]. With conventional analog broadcasting, reception in heavily shadowed areas is poor, but because of the signal's analog structure, the received signal can generally be deciphered by the listener[17]. However, when the reception of a COFDM signal is attempted in a heavily shadowed area, many bit errors occur at the receiver, preventing it from successfully reconstructing the transmitted audio signal. As a result, the system "collapses" leaving no decipherable signal for the listener. It is because of this sudden drop in performance that an alternative coverage technique was required for the experimental DAB system[17].

It has been suggested that in order to meet the proposed coverage requirements, the COFDM-based DAB system should utilize a transmission concept referred to as *distributed emission*[17][27]. With this concept, a network of multiple transmitters operating on the same frequency would be used to broadcast the COFDM signal rather than a single transmitter. It is the presence of the guard interval in the COFDM signal that permits this use of multiple transmitters. As long as the delay associated with receiving all of the multiple transmissions is less than the length of the guard interval, no inter-symbol interference will be introduced into the received signal and the receiver's performance will be based on the sum of the power contained within all of the multiple transmissions[27].

An example of this technique would be a broadcasting network consisting of a central transmitter (either terrestrial or satellite) and a series on-channel, terrestrial repeaters. The repeaters are used to create "active" echoes by re-broadcasting the signal of the central transmitter. This allows them to enhance reception in areas that are severely shadowed from the central transmitter or to extend the total coverage area by supplementing the initial transmission[27]. Figure 3.1 demonstrates a simple implementation of this type of network.

3.1.4 Selecting a Transmission Mode

For operation at 1.5 GHz, only two of the system's operating modes are appropriate. Mode II uses 384 subcarriers spaced 4 kHz apart while mode III uses 192 subcarriers with a spacing of 8 kHz[14]. When deciding which mode should be selected, the designer must consider two issues.

First, the maximum allowable distance between the multiple transmitters is proportional to the length of the guard interval. If the transmitters are close together, the delay between receiving the multiple broadcasts will be quite small. As the distance between them increases, so does the potential delay associated with receiving their transmissions. Should this delay begin to exceed the length of the guard interval, their transmissions will introduce inter-symbol interference into the received signal, thus degrading the performance of the DAB system. This would indicate that mode II is the more attractive implementation as its guard interval is longer than mode III's. As a result, mode II would require fewer transmitters to provide the same coverage as mode

Figure 3.1: An example of the distributed emission concept

III.

However, since differential encoding/decoding is used, the receiver requires that the channel response be approximately the same from one symbol interval to the next. If this condition is not met (i.e. in the case of a fast fading channel), the differential decoder will introduce a large number of bit errors at the receiver. Therefore, from a sheer performance perspective, mode III becomes more attractive than mode II since its symbol interval is shorter than that of mode II. This would make mode III less susceptible to the effects of high Doppler frequencies.

When considering the above, it becomes apparent that neither transmission mode appears to be significantly better than the other. For this reason, the work done in this document will focus on both transmission modes.

3.1.5 Transmission Format

3.1.5.1 Frame Size

The COFDM-based DAB system transmits its broadcast information as a sequence of frames that are 24 ms in length. This frame length was selected to be compatible with the ISO Layer II source coding system[11][14] being used to compress the transmitted audio signals. In mode II, 24 ms corresponds to 77 symbol intervals and to 154 symbols in mode III. Of these, the first two symbols of each frame are reserved for frame synchronization purposes (see section 3.1.5.2), while the remaining symbols hold the transmitted data[12][18].

3.1.5.2 Synchronization Symbols

The first symbol of each frame is referred to as the *blank symbol*. During this symbol, all of the subcarriers transmit no information ($C_{ik} = 0$). This allows the receiver to detect the start of a frame and to estimate the noise-plus-interference power at each frequency[18].

The *sweep symbol* is used as the second symbol of every frame. This symbol has a fixed magnitude and a known phase on each subcarrier that are given by $|C_{ik}| = 1$ and $\angle C_{ik} = \pi k^2/N_c$. The addition of this symbol allows the receiver to estimate the impulse response of the channel and provides a phase reference to demodulate the first data symbol[18].

3.2 The Simulation Model

This section serves to introduce the model of the COFDM-based DAB system that was used during the computer simulation. All signals and system components were represented by their baseband equivalents, with the block diagram of figure 3.2 demonstrating their interconnection. The following three subsections discuss the components of the system's transmitter and receiver as well as the experimental channel models. The final subsection outlines the test procedures that were used to verify the validity of the simulation software. The simulation program utilized a sampling rate of 4.096 MHz in both the transmitter and receiver and assumed perfect synchronization.

The software was written entirely in C and compiled using the gcc compiler of the SunOs 4.1.1 operating system. All simulations were run on a SUN SPARC-2 workstation.

Figure 3.2: The COFDM-based DAB system

3.2.1 COFDM Transmitter

3.2.1.1 Convolutional Encoder

While the particular channel coding scheme to be used in any COFDM-based DAB system has been the topic of much research [2][15][19], a conventional convolutional code was implemented in the system under study. The code parameters were as follows:

$k = 1$
 $n = 2$
constraint length (L) = 7
generator polynomials = 0133, 0171

The generator polynomials listed above are represented in octal and are considered an optimum code for maximizing free distance for a constraint length of 7[28]. They can be found listed in the generator polynomial tables of [28] or [29].

3.2.1.2 Interleaver

Interleaving is a process that is used to combat the fading characteristics of a channel. The fundamental concept is to use an error correcting code with relatively short codewords (i.e. a convolutional code) and then to scramble the coded bits so that they are not transmitted in the order in which they are to be interpreted. This allows the burst errors that occur when the signal fades to be dispersed as random bit errors throughout the descrambled bit stream at the receiver. The error correcting code is then used to correct these randomized bit errors. A more detailed discussion of various interleaving techniques can be found in [1] or [30].

For this system, both time and frequency interleaving were utilized. The purpose of the time interleaving was to separate consecutive bits by a period of time comparable to or exceeding the coherence time of the channel. Similarly, the purpose of the frequency interleaving was to separate consecutive bits by a frequency spacing comparable to or exceeding the coherence bandwidth of the channel. Results in [15] indicated that time interleaving with a depth of 384 ms (16 data frames) combined with frequency interleaving¹ was sufficient for combatting the fading characteristics of the channels examined.

3.2.1.3 Differential Encoder

This encoder accepted a binary input that was immediately converted into QPSK symbols according to the symbol mapping shown in figure 3.3. These symbol values, denoted by c_{ik} had complex values of $\pm 1 \pm j$. They were then differentially encoded along each subcarrier using

$$C_{ik} = \frac{c_{ik}C_{i-1,k}}{(1+j)} \quad (3.2.1)$$

The resulting symbol mapping of C_{ik} can be seen in figure 3.4. Once encoded, the symbols were forwarded to the OFDM modulator.

¹ Exact time and frequency interleaving algorithms are the proprietary knowledge of CRC and cannot be presented here.

Figure 3.3: The QPSK constellation

Figure 3.4: The DQPSK constellation

3.2.1.4 OFDM Modulator

This block accepted N_c differentially encoded channel symbols and multiplexed them onto the subcarriers that were being used by the DAB system. This multiplexing was accomplished through the use of an IDFT, which was implemented with the radix-2 inverse fast Fourier transform (IFFT) algorithm[31].

Since the radix-2 IFFT required an integer power of 2 inputs, an N point IFFT was performed where $N = 2^m$ and m was the smallest integer larger than $\log_2(N_c)$. The first and last $N_c/2$ input symbols of the IFFT processor were utilized by the OFDM modulator for transmitting the input channel symbols, while the remaining $(N-N_c)$ IFFT symbols were set to zero. This ensured that all of the input channel symbols were multiplexed onto subcarriers within the system's allocated bandwidth.

After the IFFT operation had been completed, the guard interval was inserted and the resulting sequence was prepared to be transmitted.

3.2.1.5 Transmitter Filters

This block contained the filter design that was under study. Both FIR and IIR filters were considered, but much of the work focused on the IIR designs. The FIR designs were implemented as a tap delay line, while the IIR designs consisted of a cascade of second order biquads. All of the filters incorporated into the simulation program were designed with the software package DFDP3/plus manufactured by Atlanta Signal Processors Inc.(ASPI).

It should be noted that transmitter filtering was only included in simulations that

were intended to study the effects of filtering on the performance of the system and unless otherwise specified, it was not included in those that studied the effects of nonlinear amplification.

3.2.1.6 Transmitter Power Amplifier

Prior to passing the signal through the transmission channel, it was subjected to the nonlinear characteristics of a typical broadcast amplifier. Each sample of the transmitted signal was distorted in both amplitude and phase according to measurements made on the amplifier under investigation[24][25][32]. The exact distortions appropriate for each input sample were determined by linearly interpolating between entries in a look up table.

Similar to the transmitter filters, the amplifier models were only included in the simulations that evaluated the effects of nonlinear amplification and not the simulations that solely studied the effects of filtering.

3.2.2 COFDM Receiver

3.2.2.1 Receiver Filters

The filters present in the receiver had a frequency response identical to those used in the transmitter. These filters were intended to suppress any high frequency noise components in the received signal as well as to minimize the effects of any adjacent channel interference. However, evaluating the performance of the DAB system in the presence of adjacent channel interference was beyond the scope of this work.

As with the transmitter filters, these filters were only included in those simulations which directly investigated the effects of non-ideal filtering unless it is otherwise specified.

3.2.2.2 OFDM Demodulator

The OFDM demodulator discarded the first N_g samples received during each transmitted symbol interval (i.e. the signal received during the guard interval) and then proceeded to demultiplex the transmitted channel symbols from the remaining N samples (i.e. the useful symbol interval) using the DFT. As before, the radix-2 fast Fourier transform (FFT) algorithm was used to implement the N point DFT. After demultiplexing the subcarriers within the received signal, the demodulator expected the received channel symbols to be located on the first and last $N_c/2$ output symbols of the N point FFT processor. These recovered symbols were then forwarded to the differential decoder.

3.2.2.3 Differential Decoder

The purpose of the differential decoder was to reverse the differential encoding performed at the transmitter. This was accomplished using

$$c_{ik}' = (1+j)Y_{ik}Y_{i-1,k}^* \quad (3.2.2)$$

where Y_{ik} was the value received during the i^{th} symbol interval on the k^{th} subcarrier. The resulting value, c_{ik}' , corresponded to an estimate of the original QPSK symbol that was constructed from the binary output of the transmitter's interleaver. These QPSK values

were then quantized along the real and imaginary axis using a 16 level, uniform quantizer and fed to the receiver's de-interleaver.

3.2.2.4 De-Interleaver

The de-interleaver was used to de-scramble the transmitted bit sequence and re-assemble it in the order in which it is intended to be interpreted. Its purpose was to perform the exact inverse of the transmitter's interleaver. It should be noted that the received samples were quantized before they were de-interleaved. This order is reverse to what would traditionally be done but for the proposed system, this ordering of the operations had no effect on the received BER and it resulted in the simulation program requiring 75% less memory to store the received samples than if de-interleaving was done first.

3.2.2.5 Viterbi Decoder

The Viterbi decoder was a realization of a maximum likelihood decoder for convolutional codes. The decoder implemented used a trellis structure to decode the received binary data stream in order to recover the original information bit sequence and it was capable of performing both hard and soft decision decoding. However, due to superior performance, soft decision decoding was used exclusively in the test system. In an AWGN channel, 3 bit (8 level) uniform quantization has been determined to give near optimum results[16][29] while for various fading channels, it has been demonstrated that 4 bit (16 level) uniform quantization is required for near optimum performance[15].

In order for the soft decision decoding to be successfully implemented, a means of assigning quantization metrics was necessary. While it is traditional to assign these metrics based upon the statistical properties of the channel, this can be a tedious and daunting task. An alternative technique proposed by Clark in [29], involves assigning a linear scaling to these metric values. He points out that it is the difference between the various metric values that allows successful decoding, not the actual metric values themselves. He proceeded to demonstrate that using his linear technique, the results were virtually identical to those obtained using statistical metrics. In addition, Clark's technique did not require metric recalculation for different received SNRs.

For the Viterbi decoder implemented, a 16 level uniform quantizer was used in conjunction with Clark's linear metric structure. This combination was selected because of the results of previous research[15][18] and its exceptional ease of implementation. As well, the path memory of the decoder was set to 64 bits.

3.2.3 Experimental Channel Models

Since it has been proposed that the COFDM-based DAB system be implemented as network of multiple transmitters (see section 3.1.3), it is possible that these systems could be terrestrial-based, satellite-based, or a mix of both. As a result, it was necessary to study its performance over both Rayleigh fading terrestrial channels and Ricean fading satellite channels. This section serves to introduce the simulated channel models and define the rationale behind choosing them.

3.2.3.1 Terrestrial Channels

The terrestrial channel models selected were those proposed by the Groupe Speciale Mobile (GSM) in a report regarding the modelling of UHF radio channels[33] and later used in DAB research by the CCETT[10]-[13][34]. Of the four models presented, only the typical urban (TU) and the rural area (RA) were utilized. In [15], it was demonstrated that the performance in the other two models, bad urban (BU) and hilly terrain (HT), was virtually identical to that of the TU channel, so they were not considered. The multipath intensity profiles of both the TU and RA mobile radio channels are shown in figures 3.5 and 3.6. Recent terrestrial channel measurements taken at 1.5 GHz[35] confirm that these multipath intensity profiles are representative of those found in typical urban and rural areas throughout Canada.

A classical U-shaped Doppler power spectrum (see figure 3.7) was incorporated into both of the terrestrial channel models[10]-[13][34] and was defined by:

$$S_c(\lambda) = \begin{cases} 1/\pi (\lambda_{\max}^2 - \lambda^2)^{-1/2} & ; \lambda < \lambda_{\max} \\ 0 & ; \text{otherwise} \end{cases} \quad (3.2.3)$$

3.2.3.2 Satellite Channels

The satellite channel model utilized was similar to that employed in the study of MSAT, a mobile satellite communication system operating just above 1.5 GHz[36][37]. This model consisted of a direct path with a fixed Doppler shift and a Rayleigh fading dispersive component[36]. This emulated a flat fading channel which can be considered the worst case for satellite broadcasting[16].

The dispersive component of the simulated channel was generated using Jakes' sum of nine sinusoids technique and the resulting Doppler power spectrum was given by [38]

$$S_c(\lambda) = \begin{cases} (1 - (\lambda/\lambda_{\max})^2)^{-1/4} & ; \lambda < \lambda_{\max} \\ 0 & ; \text{otherwise} \end{cases} \quad (3.2.4)$$

This corresponded to a U-shaped Doppler power spectrum, see figure 3.8, similar to that used in the terrestrial channel models. A channel parameter, K , was specified to represent the ratio of the power in the dispersive component of the received signal to that in the direct path.

While it could be argued that some frequency selectivity would exist in a satellite channel for the DAB system under study, measurements made in [39] at 1.5 GHz stipulate that the coherence bandwidth of such a satellite channel falls in the range of 0.8 to 4.0 MHz. This range suggests that while a satellite channel could appear to be moderately frequency selective to the experimental DAB system, the flat fading case examined would not be uncommon either.

Figure 3.5: The typical urban (TU) multipath intensity profile

Figure 3.6: The rural area (RA) multipath intensity profile

Figure 3.7: The Doppler power spectrum of the terrestrial channels

Figure 3.8: The Doppler power spectrum of the satellite channel's dispersive component

3.2.4 Software Verification

In order to have confidence in the simulation results that were obtained, the software developed was incrementally tested in order to verify its proper operation. These tests consisted of:

- 1) Convolutional Encoding/Decoding: A single carrier BPSK system was simulated using convolutional codes with a code rate of $1/2$ and a constraint length of either 3, 5 or 7. Results were generated using both hard and soft decision Viterbi decoding over an AWGN channel. The decoder had a path memory of 32 bits and the soft decision decoder utilized an 8 level uniform quantizer. These results were then compared to the traditional performance benchmark of Heller and Jacobs [28] (see appendix 2) to demonstrate the validity of the software. For the convolutional code (see section 3.2.1.1) used in the experimental system, the difference between the E_b/N_0 required for a BER of $1E-4$ in the implemented decoder and Heller and Jacobs was approximately 0.25 dB and 0.10 dB for hard decision decoding and soft decision decoding respectively.
- 2) The Channel Models: A single time impulse was inserted into the terrestrial channel models and the corresponding multipath intensity profile was evaluated. These profiles agreed with those illustrated in [33] and reproduced in figures 3.5 and 3.6. Since the satellite channel was a flat fading channel, its multipath intensity profile could not be measured as it was essentially an impulse. The

Doppler power spectrum of each of the three channel models was also evaluated and they agreed with what was stipulated in [10],[34] and [38]. The final test of the channel model software occurred during test #3 when the system performance was determined over them and compared to previously reported results.

- 3) **Ideal COFDM DAB System:** The COFDM-based system outlined in sections 3.2.1 and 3.2.2 was implemented excluding the filters and nonlinear amplifiers. Its performance was measured over an AWGN channel and the three experimental channel models for various channel parameters ($\lambda_{\max}$, K and λ_{dp}) and compared to previously published results. The results of the simulation program were essentially the same as those in [15] and [16]. They are included in section 4.3.2 of this document.
- 4) **Filter Implementations:** The impulse response of various FIR and IIR filters were measured and compared to that predicted by DFDP3/plus, the digital filter design package produced by ASPI. For all the filters tested, both the magnitude and phase responses matched the results of DFDP3/plus.
- 5) **Nonlinear Amplifier Modelling:** A constant envelope signal was fed into the nonlinear amplifier simulation code and the amplifier's power and phase characteristics were measured. These simulation measurements were then compared to manufacturer specifications[24][25][32] and found to be virtually

identical. As well, simulation of the entire DAB system (excluding filtering) was performed with a very large input backoff (approximately -20 dB for a TWTA) to ensure that the amplifier was operating in its linear region with minimal phase distortion. At this input backoff setting, the results measured were equivalent to those that assumed a perfectly linear amplifier (see test #3 or section 4.3.2).

After successfully completing the five tests listed above, it was determined that the results generated by the computer simulation program would be reliable.

Chapter 4

EXPERIMENTAL RESULTS

The purpose of this chapter is to present the results that were obtained through the use of the computer simulation program outlined in chapter 3. The techniques used in measuring the performance of the DAB system are introduced in sections 4.1 and 4.2, while the experimental results are displayed and discussed in section 4.3.

4.1 BER Measurement of the COFDM Simulations

The average BER is a common performance measure for digital broadcasting systems[10]-[13][15][16]. However, since most of the experimental measurements were made in a time-varying channel, the simulation had to be run for a sufficient length of time in order for the measured BER to be an accurate estimate for the average BER.

To evaluate this length of time, simulations were performed that recorded the measured BER versus the number of Doppler cycles that had passed. Figure 4.1 contains such a plot for a mode III COFDM signal transmitted through a typical urban channel with $\lambda_{max} = 100$ Hz at an E_b/N_o of 12.0 dB. It can be seen in this plot that the measured BER converges to the published result of $6.0E-5$ [15] in slightly under 50 Doppler cycles. As a result, all simulations that incorporated fading channels ran for at least 50 Doppler cycles. Also, each resultant BER value consisted of at least 100 error events.

Figure 4.1: Plotting BER versus the number of doppler cycles

4.2 Calculating E_b/N_0 from the Measured SNR

For the purpose of presenting the simulation results, it was necessary to normalize the SNR measured at the COFDM receiver to the standard measurement of E_b/N_0 . This would allow a true comparison to other broadcast systems since the value of the SNR depends on the transmitted signal's format, while E_b/N_0 is based solely on the throughput of the system. The relationship between the two measures is given by,

$$E_b/N_0 = \text{SNR} (B_n/B_i) \quad (4.2.1)$$

The noise bandwidth B_n was equal to the sampling rate used in the receiver, while B_i was equal to the transmitted net information rate[18]. The net information rate had to take various system parameters into consideration. These included the guard interval, the code rate, the number of bits transmitted per channel symbol and the synchronization symbols inserted by the transmitter.

4.3 Presentation of Results

The following pages present the results obtained during the simulation of the experimental DAB system. The results were obtained using four different maximum Doppler frequencies in the three channel models of section 3.2.3. The frequencies used were 25 Hz, 100 Hz, 200 Hz and 275 Hz. At 1.5 GHz, they correspond to vehicle speeds of 18 km/h, 72 km/h, 144 km/h and 198 km/h respectively. This range of Doppler frequencies allows the results to emulate a slow moving vehicle, fast city driving, fast highway driving or a high-speed train.

In the case of the satellite channel, the channel parameter K was defined by

$$K = 10 \log_{10} (\text{average multipath power} / \text{direct path power}) \quad (4.3.1)$$

and had a value of either -5 dB or -10 dB. As well, the Doppler frequency of the direct path was set to one of 0, $0.5\lambda_{\max}$ or $0.87\lambda_{\max}$ (angle of arrival of 90° , 60° or 30°).

The most significant point on each performance curve was the E_b/N_0 value which had a BER of $1E-4$. From listening tests conducted in [40], it was demonstrated that a listener began to notice the presence of noise in the received signal when the input to audio source decoder had this BER. The value of $1E-4$ was also stipulated to be the performance benchmark by working group 10B of the ITU-BR[14].

4.3.1 Accuracy of the Simulation Results

To determine the accuracy of the results presented, one simulation was run repeatedly using different seeds for the pseudo-random number generators used to create the transmitted bit sequence and the gaussian noise. It was found that after 15 runs, the deviation in the E_b/N_0 measurements at a BER of $1E-4$ was approximately ± 0.2 dB.

4.3.2 Ideal System Results

Figures 4.2 through 4.9 indicate the performance of the COFDM-based DAB system over the experimental mobile channels. This ideal system assumes perfect synchronization in the receiver, no filtering and a completely linear amplifier.

4.3.2.1 Terrestrial Channel Results

Simulation in the terrestrial channels was performed at all four of the aforementioned maximum Doppler frequencies so that the effects of varying the channel's maximum Doppler frequency could be noted. From figures 4.2 through 4.5, it can be concluded that as the maximum Doppler frequency was increased, the system's performance degraded. This behaviour resulted from the time variations of the channel. As the channel's maximum Doppler rate increased, the likelihood of the channel response remaining constant over consecutive symbol intervals along each subcarrier decreased. In the receiver, each subcarrier's differential decoder assumed that the channel response would remain constant over successful symbol intervals and when this assumption was invalidated, the decoder introduced errors into the received bit stream. Because mode II had a longer symbol duration, it was affected more by this occurrence than mode III for a fixed value of $\lambda_{\max}$ except at very low Doppler rates. In an attempt to offset the extreme effects of high Doppler frequencies demonstrated here, a study presently underway at CRC is investigating the use of decision feedback loops to improve the performance of the differential detection in the receiver[41].

The difference in performance between the typical urban and rural area channels was a result of the frequency diversity that existed in the typical urban channel. Because the multipath spread of the rural area channel was $0.7 \mu\text{s}$ ($B_c = 1.43 \text{ MHz}$), it was essentially a flat fading channel with respect to the COFDM signal transmitted by the DAB system. Conversely, the typical urban's multipath spread of $7 \mu\text{s}$ ($B_c = 143 \text{ kHz}$) made the channel frequency selective with respect to the transmitted signal.

Figure 4.2: The performance of mode II over the TU channel

Figure 4.3: The performance of mode II over the RA channel

Figure 4.4: The performance of mode III over the TU channel

Figure 4.5: The performance of mode III over the RA channel

4.3.2.2 Satellite Channel Results

For the satellite channel model, the maximum Doppler frequency of the channel remained fixed at a typical value of 100 Hz. Since the terrestrial channel results indicated the effects of increasing the channel's maximum Doppler frequency, the satellite channel results concentrated on the effects of varying the channel's power ratio K and the Doppler frequency of the direct path.

The results of varying these channel parameters are shown in figures 4.6 through 4.9. From these plots, it can be observed that by increasing the power ratio K , the system's performance degraded. This behaviour is reasonable when it is considered that a channel with $K = -\infty$ would be equivalent to an AWGN channel while if K was increased to $+\infty$, it would essentially be a Rayleigh, flat fading channel. This would indicate that for K values of -10 dB and -5 dB, the system performance should be worse than an AWGN channel yet better than a Rayleigh, flat fading channel with an equivalent Doppler power spectrum. When these results are compared to the AWGN channel results in [2] and the results from the typical rural channel of section 4.3.2.1, this behaviour can be verified.

The performance at $K = -10$ dB exceeded that at $K = -5$ dB due to the stronger direct path component in the received signal. For the satellite channel model employed, signal fades were caused when the received multipath component essentially cancelled out the received direct path component. The stronger the direct path component was, the less this situation occurred.

Figure 4.6: The performance of mode II over the SAT channel with $\lambda_{\max} = 100$ Hz and $K = -10$ dB

Figure 4.7: The performance of mode II over the SAT channel with $\lambda_{\max} = 100$ Hz and $K = -5$ dB

Figure 4.8: The performance of mode III over the SAT channel with $\lambda_{\max} = 100$ Hz and $K = -10$ dB

Figure 4.9: The performance of mode III over the SAT channel with $\lambda_{\max} = 100$ Hz and $K = -5$ dB

The dependence of the system's performance on the direct path's Doppler frequency was not so easily explained. Figure 4.10 indicates the approximate E_b/N_0 value that was required to achieve a BER of $1E-4$ with mode III for various λ_{dp} values. Since there appears to be no distinctive trend, it is difficult to conclude as to what is causing this unusual dependency. It was found in [41], that when the received signal consisted of a single path with a fixed Doppler shift, the system performance degraded as the Doppler shift was increased. As well, section 4.3.2.1 confirmed a similar performance relationship in a flat, Rayleigh fading channel (i.e. the rural area channel). Therefore, it can be stated that this unusual dependence was due to the interaction between the Doppler shift of the direct path and multipath components of the received signal. In order to obtain a definitive explanation for this occurrence, an extensive analysis of the COFDM signal and the channel model would be required. This investigation was not directly related to the objective of this thesis, but it should be noted that an alternative study by Le[16] made similar observations, so the results would appear to be valid.

4.3.3 Non-Ideal Filter Results

The results presented in this section investigate the effects of non-ideal filters in the transmitter and receiver on the DAB system's performance. As mentioned previously, the filter characteristics that were investigated were the effects of a non-linear phase response, the steepness of the transition band's roll-off, and the presence of ripples in the filter's passband. For all of the results presented, perfect synchronization and linear amplification was assumed.

Figure 4.10: The approximate E_b/N_0 at a BER $\approx 1E-4$ for mode III over the SAT channel with $\lambda_{\max} = 100$ Hz, $K = -10$ dB and various direct path Doppler frequencies

In order to avoid unnecessary simulations, it was decided that a case study would be performed involving the use of transmission mode III over a typical urban channel with $\lambda_{\max} = 275$ Hz. During this study, many filter designs were tested and from these results, a smaller set of filters was selected to be simulated in other channel environments.

4.3.3.1 A Non-Ideal Filtering Case Study

The study performed investigated the use of both FIR and IIR filter designs. In the FIR class, only windowed designs were examined while for the IIR class, only the Butterworth and Chebychev approximations were used. These limits forced the study to have a reasonable scope, yet allowed all of the desired filter characteristics to be tested.

By considering the filters examined as the lowpass equivalent of the IF stage bandpass filters, the passband of the filters had to be half of the bandwidth of the COFDM signal. Therefore, the design criteria for all of the experimental filters was set at having a passband corner frequency (-3 dB frequency) of approximately 768 kHz (± 2 kHz) and a minimum stopband attenuation of at least 40 dB.

During this study, it was observed that the loss introduced by many of the filter designs was comparable. This made the resulting BER plots difficult to read as many of the performance curves overlapped with each other. As a result, for each filter class (windowed FIR, Butterworth or Chebychev), only the performance curve of the design which introduced the greatest loss is plotted with the ideal case in figures 4.11 through 4.13. However, the most significant results for all of the filters examined are summarized in tables 4.1 through 4.3.

4.3.3.1.1 Windowed FIR Filters

The rationale behind selecting this class of filters was that they exhibited linear phase characteristics. Designs with rectangular windows were not tested as they did not meet the stopband attenuation criteria. Due to its popularity[31], the Hamming window was used exclusively for all of the windowed FIR filters, allowing them to meet, or exceed the stopband attenuation design criteria while maintaining a reasonably flat passband response. By altering the number of taps in the filter design, the sharpness of the transition band could be changed and any impact on system performance noted. As well, since a sampling frequency of 4.096 MHz was used in the receiver, filters with an extraordinary large number of taps were not tested as they would be difficult to realize in an actual real-time system. Accordingly, the tap length of the filters examined was set to either 20, 40 or 60.

The results generated are presented in figure 4.11 and table 4.1. These results indicate that as the number of taps was increased, the losses caused by the filter decreased. However, this reduction in loss was not necessarily caused by the increase in the steepness of the transition band. By comparing the frequency responses of these filters (see appendix 3), it can be seen that the passband response of the filter was also altered when the tap length of the filter was changed. The response became flatter over a larger range of the passband as the length of the filter was increased. It is more likely that the losses seen here were the result of the spectral shaping (i.e. the attenuation of some of the subcarriers) caused by the passband response rather than the steepness of the transition band. Therefore in order to better evaluate the effects of increasing the

Figure 4.11: The effects of windowed FIR filters on mode III over the TU channel with $\lambda_{\max} = 275$ Hz

steepness of the transition band of the filter, a study should be conducted in the presence of adjacent channel interference. However, this study was not intended to consider the presence of adjacent channel interference.

WFIR Filter Length	Loss at BER = 1E-4 (dB)	-40 dB Frequency (kHz)
20 Taps	0.80	1176
40 Taps	0.50	966
60 Taps	0.45	901

Table 4.1: Summarizing the performance of the windowed FIR filters

4.3.3.1.2 Butterworth IIR Filters

The Butterworth, or maximally flat, IIR filter approximation was selected because it had a flat passband response (similar to the FIR designs tested) yet did not exhibit a linear phase response. This would allow a demonstration of the impact of a nonlinear phase response on the DAB system's performance.

Figure 4.12 and table 4.2 contain the simulation results for Butterworth filters of order 8, 10 and 12. From these results, it can be concluded that the nonlinear phase characteristics of the Butterworth IIR designs had very little effect on the system's performance. This result can be justified by considering that the wideband COFDM signal consisted of a large number of narrowband DQPSK subcarriers that were frequency

Figure 4.12: The effects of Butterworth IIR Filters on mode III over the TU channel with $\lambda_{\max} = 275$ Hz

multiplexed together. As long as the phase response of the filter remained reasonably linear over the narrow bandwidth of each subcarrier, the nonlinear phase behaviour of the filter should have had little effect. The small losses observed can be primarily attributed to the spectral shaping caused by the filter around its -3 dB corner frequency rather than the nonlinear phase characteristic. The significance of this result is that it demonstrates that the use of a group delay equalizer would not be necessary should the use of a filter with a nonlinear phase response be required.

The loss measured for the 10th order filter would appear to be an anomaly but such a small discrepancy could have resulted from the 10th order filter having a -3 dB frequency slightly lower than those in the 8th and 12th order designs.

Butterworth Filter Order	Loss at BER = 1E-4 (dB)	-40 dB Frequency (kHz)
8th	0.15	1135
10th	0.30	1061
12th	0.15	1028

Table 4.2: Summarizing the performance of the Butterworth IIR filters

4.3.3.1.3 Chebychev IIR Filters

The Chebychev approximation IIR filter designs contained both passband ripples and a nonlinear phase response. Because the results of the previous section indicated that a nonlinear phase response had little effect on the system's performance, any significant loss present in these simulations could be attributed to the presence of ripples in the passband response of the filter and not its nonlinear phase response.

Two sets of simulations were performed using Chebychev IIR filter designs of order 4, 6 and 8. The first, see figure 4.13 and table 4.3, tested filters that contained 1 dB passband ripples, while figure 4.14 and table 4.4 examined designs with 2 dB passband ripples.

From these results, it can be concluded that passband ripples can have a significant impact on the performance of the COFDM-based DAB system. The cause of this loss was again the spectral shaping of the wideband COFDM signal that resulted from introducing filters into the DAB system. The passband ripples caused the attenuation and amplification of the many subcarriers within the COFDM signal. This resulted in an unexpected increase or absence of energy along each subcarrier in the received signal. If this increase or absence was small, then its effect on the performance of the system was minimal. However, when it became large, its effect became significant. This loss of performance resulted from non-optimal metric assignments in the receiver's Viterbi decoder. The assignments were initially made on the assumption that the average received energy along all of the subcarriers was constant. However, the presence of passband ripples made this assumption invalid which degraded the performance of the

Figure 4.13: The effects of Chebychev IIR filters with 1 dB ripples on mode III over the TU channel with $\lambda_{\max} = 275$ Hz

Figure 4.14: The effects of Chebychev IIR filters with 2 dB ripples on mode III over the TU channel with $\lambda_{\max} = 275$ Hz

Viterbi decoder and in turn, the entire DAB system.

Chebyshev Filter Order	Loss at BER = $1E-4$ (dB)	-40 dB Frequency (kHz)
4th Order	0.25	1257
6th Order	0.55	1012
8th Order	0.20	920

Table 4.3: Summarizing the performance of Chebyshev IIR filters with 1 dB ripples

Chebyshev Filter Order	Loss at BER = $1E-4$ (dB)	-40 dB Frequency (kHz)
4th Order	1.65	1200
6th Order	1.20	990
8th Order	0.35	906

Table 4.4: Summarizing the performance of Chebyshev IIR filters with 2 dB ripples

The contents of tables 4.3 and 4.4 indicate that for all three of the filter orders examined, the design with 1 dB ripples introduced less loss into the system than the corresponding 2 dB design. However, the amount of additional loss caused by the larger ripples reduced as the filter order was increased. From examining the frequency

responses of the six filters (see appendix 3), it can be seen that the ripple size specified to the filter design package (either 1 dB or 2 dB) was only a nominal measurement and not the actual magnitude of the passband ripples in the final design. The following table presents the actual maximum/minimum passband ripple magnitude for the filters studied.

Filter Order	1 dB Filter Designs	2 dB Filter Designs
4th	0.76 dB / -0.83 dB	1.97 dB / -2.56 dB
6th	0.88 dB / -0.98 dB	1.54 dB / -1.88 dB
8th	0.52 dB / -0.54 dB	1.06 dB / -1.20 dB

Table 4.5: The actual measured maximum/minimum passband ripple magnitude in the Chebychev IIR filters

If the results of tables 4.3 and 4.4 were considered independently, it could be concluded from table 4.5 that the larger the filter's passband ripples were, the more loss the filter introduced into the system. With the 1 dB ripple designs, the 6th order design had the largest passband ripples, and it introduced the most loss. Conversely, the 8th order design had the smallest passband ripples and it introduced the least amount of loss. Similarly, with the 2 dB designs, the 4th order filter had the largest ripples and the largest loss while the 8th order filter had the smallest ripples and the smallest loss.

However, when both sets of results are considered together, this statement does not hold entirely true. This is noticed when the 6th order, 1 dB ripple design and the 8th

order, 2 dB ripple design are compared. The measurements of table 4.3 stipulate that the ripples of the 8th order, 2 dB design were actually larger than those in the 6th order, 1 dB design, but according to tables 4.3 and 4.4, it introduced less loss than the 6th order, 1 dB design. This would indicate that the loss introduced by the filter was dependent on both the magnitude of the passband ripples and the order of the filter.

The influence of the filter order on the loss experienced can be confirmed by considering the frequency response plots of appendix 3. These plots demonstrate that as the filter order is increased, the number of passband ripples also increased. As well, as the number of ripples in the passband increased, their width decreased and the number of subcarriers that each ripple affected also decreased. Measurements made on the frequency responses of both these two filters determined that both had gains less than 0 dB for about 55% of the passband, so both designs would have attenuated and/or amplified approximately the same number of subcarriers. The -3 dB frequency of the 8th order, 2 dB ripple design was also measured to be marginally smaller (766.5 kHz) than the 6th, 1 dB ripple design (767.9 kHz), so this could not have been responsible for the superior performance of the 8th order, 2 dB ripple design. This leaves only the width and location of the passband ripples as the cause of the performance difference, and this characteristic is solely dependent on the order of the filter and the design algorithm selected.

From these results, it can be concluded that the best Chebychev design would be that which offered the best balance between a high filter order and small passband ripples.

4.3.3.2 The Use of Non-Ideal Filters

The purpose of this section was to quantify the losses introduced by some of the better filter designs determined in the case study of section 4.3.3.1 over various transmission environments. The filters selected were those that caused the least amount of performance degradation in each of sections 4.3.3.1.1 through 4.3.3.1.3 while providing a relatively steep transition between the passband and the stopband. These three filters were:

1. 60 tap windowed FIR filter with a Hamming window
2. 12th order Butterworth IIR filter
3. 8th order Chebychev IIR filter with 1 dB ripples

In order to keep the presentation of results brief, the use of these three filters was simulated in both transmission modes over the typical urban, rural area and a satellite channel with $K = -10$ dB and $\lambda_{dp} = 0$ Hz. The value of λ_{max} was fixed at a typical value of 100 Hz for all of these situations. While parts of this study demonstrated that the filtering loss increased as λ_{max} was increased (i.e. the TU channel), it was found that the loss incurred at the greatest λ_{max} considered, 275 Hz, was consistently within 0.2 dB of the loss suffered at $\lambda_{max} = 100$ Hz. It was also learned that the K parameter and the direct path's Doppler frequency of the satellite channel had no significant impact on the losses introduced by the filters.

As was the case in section 4.3.3.1, the losses introduced by these three filters were comparable to each other for any given channel. This made the resulting BER plots difficult to read as their respective performance curves commonly overlapped.

Accordingly, only the performance curve of the filter which introduced the greatest amount of loss over the given channel is shown in figures 4.15 through 4.20 along with the ideal case. Tables 4.6 and 4.7 summarize the losses introduced by all of the filter designs over all of the channels for mode II and mode III respectively.

Filter Design	TU with $\lambda_{\max} = 100$ Hz	RA with $\lambda_{\max} = 100$ Hz	SAT with $\lambda_{\max} = 100$ Hz, $K = -10$ dB and $\lambda_{dp} = 0$ Hz
60 tap WFIR	0.10 dB	0.00 dB	0.20 dB
12th order Butterworth	0.15 dB	0.00 dB	0.10 dB
8th order Cheby w/ 1 dB ripples	0.10 dB	0.00 dB	0.15 dB

Table 4.6: The losses experienced by mode II due to non-ideal filtering over selected channels at a BER of 1E-4

Filter Design	TU with $\lambda_{\max} = 100$ Hz	RA with $\lambda_{\max} = 100$ Hz	SAT with $\lambda_{\max} = 100$ Hz, $K = -10$ dB and $\lambda_{dp} = 0$ Hz
60 tap WFIR	0.25 dB	0.25 dB	0.15 dB
12th order Butterworth	0.15 dB	0.05 dB	0.10 dB
8th order Cheby w/ 1dB ripples	0.15 dB	0.00 dB	0.00 dB

Table 4.7: The losses experienced by mode III due to non-ideal filtering over selected channels at a BER of 1E-4

Figure 4.15: The effects of selected filters on mode II over the TU channel with $\lambda_{\max} = 100$ Hz

Figure 4.16: The effects of selected filters on mode II over the RA channel with $\lambda_{\max} = 100$ Hz

Figure 4.17: The effects of selected filters on mode II over the SAT channel with $\lambda_{\max} = 100$ Hz, $K = -10$ dB and $\lambda_{\text{dp}} = 0$ Hz

Figure 4.18: The effects of selected filters on mode III over the TU channel with $\lambda_{\max} = 100$ Hz

Figure 4.19: The effects of selected filters on mode III over the RA channel with $\lambda_{\max} = 100$ Hz

Figure 4.20: The effects of selected filters on mode III over the SAT channel with $\lambda_{\max} = 100$ Hz, $K = -10$ dB and $\lambda_{dp} = 0$ Hz

4.3.3.3 Comparing Filter Designs

While all of the three filter designs examined in the previous section would be suitable for use with the COFDM-based DAB system, at some point a decision would have to be made as to which design would be most appropriate. This section will provide a brief comparison between the three filter designs considered in section 4.3.3.2.

In order to implement any FIR filter design as a tap delay line, a processor would be required to execute one addition, one multiplication and one data move per tap to generate each real output sample from a real input sample. Similarly, for any IIR filter design implemented as a cascade of second order biquads (assuming direct form II[42]), a processor would be required to execute 4 additions, 5 multiplications and 2 data moves per second order biquad section to generate a real output sample from a real input sample. In the case of complex input and output samples, the processor would be required to execute twice the number of instructions listed above for both FIR and IIR designs.

The number of instructions required to generate each complex output sample by the three filters being examined is summarized in table 4.8 on the following page. It should be noted that if the filtering were done in the frequency domain, the number of instructions would be dependent only on the sampling frequency and not the individual filter design. However, the presence of the guard interval in the COFDM signals used by the experimental DAB system made this implementation cumbersome as the number of samples in each transmitted symbol duration was not an integer power of two. This condition would be required to achieve significant computational gain over the time domain case[31]. Accordingly, only time domain filtering is considered here.

Filter Design	# of Additions	# of Multiplies	# of Data Moves
60 Tap windowed FIR	120	120	120
12th order Butterworth IIR	48	60	24
8th order Chebychev I IIR	32	40	16

Table 4.8: Comparing the complexity of selected filter designs

From the values contained in table 4.8, it would appear that the Chebychev design would be the simplest filter to implement, as it required the least number of instructions to be executed per input sample. Assuming that the processor being used could execute each of the three instructions in a single clock cycle (ie. TMS320C40), the Chebychev filter would run about 1.5 times faster than the Butterworth design and over 4 times faster than the windowed FIR design.

The FIR design appears to be the least appropriate choice for the COFDM-based DAB system. Earlier results indicated that the nonlinear phase responses of the IIR filters did not have any significant impact on the system performance and while the FIR design did offer the lowest -40 dB frequency, it was only 2% less than that of the Chebychev design. In addition, the FIR design generally introduced the greatest system losses when used in conjunction with the COFDM-based DAB system. It is doubtful that any system designer could justify the choice of the FIR design considering the above factors and its increased computational complexity.

From this study, it would appear that the Chebychev IIR filter would be the most appropriate choice for the experimental DAB system. It offers the least complexity and a -40 dB frequency that is over 10% less than that of the Butterworth IIR filter considered. As well, the losses incurred by the DAB system were comparable in all of the situations simulated for both the Chebychev and the Butterworth designs.

4.3.4 Nonlinear Amplifier Results

Presented in this section are the results that were generated when the effects of nonlinear amplification were considered. The intent of this section was to determine the tradeoffs between the input backoff setting of a broadcast amplifier and the degradation in performance the system suffers from the amplifier nonlinearities present at that input backoff setting. Statistical analysis performed by M. Le (reproduced in appendix 4) demonstrated that the CDFs of the instantaneous power for the COFDM signal in both modes II and III were virtually identical. Since the impact of the amplifier's nonlinearities is dependent on the statistical nature of the input signal's instantaneous power, it can be concluded that the losses introduced by nonlinear amplification would be comparable for either mode. This conclusion was also reached by Rapp in [20], so only the performance a system operating in mode III is presented here.

The results are broken down into two groups, namely those that characterized the use of typical terrestrial amplifiers over various terrestrial channels and those that resulted from using typical satellite amplifiers over various satellite channels. While other simulations could have been done, their results would not have been relevant to any

expected system implementation (i.e. using a satellite amplifier with a terrestrial channel). As well, perfect synchronization and no filtering were assumed for all of these simulations.

4.3.4.1 Satellite Amplifiers

The satellite amplifiers selected were a TWTA and a L-Band SSPA manufactured by SPAR. The simulation results indicated that the K parameter of the channel (-5 dB or -10 dB) and the Doppler shift of the line of sight component had no significant effect on the losses due to nonlinearities, so only the results for $\lambda_{\max} = 100$ Hz, $K = -10$ dB and $\lambda_{dp} = 0$ Hz are included in this section.

The TWTA is a very common amplifier found in many satellite transponders[43]. Its nonlinear characteristics, both power and phase, are shown in figure 4.21[32], while its effects on the performance of the COFDM-based DAB system can be seen in figure 4.22. The second satellite amplifier, a SPAR L-Band SSPA, has the power and phase characteristics of figure 4.23[24]. Figure 4.24 contains the performance curves of the DAB system when the SPAR amplifier was incorporated into the simulated transmitter.

The performance trends in both figures 4.22 and 4.24 demonstrate that as the input backoff (IP0) was reduced, the losses introduced by the amplifier nonlinearities increased. This reflected the greater signal distortion caused by the severe nonlinearities present at the smaller input backoffs. However, the optimum operating point of the amplifier was not determined by the point which introduced the least amount of loss, but rather the point that offered the best trade-off between output power and performance degradation.

This trade-off was quantified by subtracting the nonlinearity loss at any input backoff setting from its corresponding output backoff. The use of this value can be best demonstrated through an example.

For a TWTA, an input backoff of -7.0 dB yields an output backoff of -2.20 dB and a nonlinearity loss of 0.95 dB. With an input backoff of -5.0 dB, the output backoff was -1.10 and the nonlinearity loss 1.60 dB. Therefore, by increasing the input power 2 dB, the output power increased 1.10 dB while the losses only increased 0.65 dB. This would result in a net gain of 0.45 dB along the transmission link. As a result, operating with an input backoff of -5.0 dB would be better than operating at -7.0 dB even though the losses incurred at -5.0 dB are greater.

Now, these same results can be calculated from comparing the difference between the output backoff and the nonlinearity losses at each point. At an input backoff of -7.0 dB, the output backoff minus the nonlinearity loss is -3.15 dB, and at -5.0 dB, this value is -2.70 dB. This indicates that an input backoff of -5.0 dB is a more appropriate choice than -7.0 dB since $-2.70 \text{ dB} > -3.15 \text{ dB}$. As well, $(-2.70) - (-3.15)$ determines the 0.45 dB gain experienced along the transmission link from reducing the input backoff setting.

Tables 4.9 and 4.10 summarize the performance of both amplifiers at various input backoff settings. The fourth column of these tables contain the output backoff minus the loss at a BER of $1\text{E-}4$ for the stated input backoff. These values have been plotted against their corresponding input backoff setting in figures 4.25 and 4.26. From this information, it can be determined that the optimum input backoff setting would be approximately -5.0 dB for a TWTA and -3.0 dB for the SPAR L-Band SSPA.

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ANSI/ISO TEST CHART No. 2

Figure 4.21: The nonlinear characteristics of a TWTA

Figure 4.22: The effects of a TWTA on mode III over a SAT channel with $\lambda_{\max} = 100$ Hz, $K = -10$ dB and $\lambda_{dp} = 0$ Hz

Figure 4.23: The nonlinear characteristics of a SPAR L-Band SSPA

Figure 4.24: The effects of a SPAR L-Band SSPA on mode III over a SAT channel with $\lambda_{\max} = 100$ Hz, $K = -10$ dB and $\lambda_{dp} = 0$ Hz

Input Backoff (dB)	Output Backoff (dB)	Loss at BER=1E-4 (dB)	Output Backoff - Loss (dB)
-3.00	-0.45	2.25	-2.70
-4.00	-0.75	1.65	-2.40
-5.00	-1.10	1.30	-2.40
-6.00	-1.60	1.10	-2.70
-7.00	-2.20	0.85	-3.05

Table 4.9: Summarizing the losses resulting from nonlinear amplification by a TWTA over a SAT channel with $\lambda_{\max} = 100$ Hz, $K = -10$ dB and $\lambda_{dp} = 0$ Hz

Input Backoff (dB)	Output Backoff (dB)	Loss at BER=1E-4 (dB)	Output Backoff - Loss (dB)
-1.00	-0.20	1.95	-2.15
-2.00	-0.40	1.50	-1.90
-3.00	-0.50	1.20	-1.70
-4.00	-1.00	0.80	-1.80
-5.00	-1.60	0.60	-2.20

Table 4.10: Summarizing the losses resulting from nonlinear amplification by a SPAR L-Band SSPA over a SAT channel with $\lambda_{\max} = 100$ Hz, $K = -10$ dB and $\lambda_{dp} = 0$ Hz

Figure 4.25: Determining the optimum operating point of a TWTA

Figure 4.26: Determining the optimum operating point of a SPAR L-Band SSPA

4.3.4.2 Terrestrial Amplifiers

The terrestrial broadcast amplifier examined was a UHF solid state power amplifier manufactured in Canada by Larcac Inc. During the course of this study, it was observed that the losses introduced by this nonlinear amplifier were different in the TU channel than in the RA channel for a fixed maximum Doppler frequency. Accordingly, the results for both of these channel models with $\lambda_{\max} = 100$ Hz are presented here.

The nonlinear characteristics of the Larcac amplifier are contained in figure 4.27. It should be noted that the phase characteristic had to be extrapolated over the input backoff range of -5.0 to 2.0 dB because of incomplete specifications provided by Larcac[25]. However, the finalized model for the amplifier was sent to Larcac and approved. Its effect on the performance of the experimental DAB system over a TU channel with $\lambda_{\max} = 100$ Hz can be seen in figure 4.28 and is summarized in table 4.11. Figure 4.29 and table 4.12 contain the results when a RA channel with $\lambda_{\max} = 100$ Hz was considered.

As was the case with the satellite amplifier results, the performance degradation caused by the amplifier nonlinearities became smaller as the input backoff of the amplifier was increased. This reduction in performance loss resulted from the more linear behaviour of the amplifier at larger input backoffs.

The difference between the losses introduced by the amplifier over the two channel models can be attributed to the presence of intermodulation products[43] in the amplifier's output signal. We must remember that when the distorted COFDM signal was passed through the terrestrial channels, it was essentially a completely different signal than the

undistorted COFDM signal studied in section 4.3.2. As a result, both are going to be affected by the fading characteristics of the two channels differently, so it would be inappropriate to expect the relative performance between the TU and RA channels to be the same for both signals.

By comparing the results of tables 4.11 and 4.12, it can be observed that the discrepancy between the losses in the two channel models is greatest at small input backoffs, where the intermodulation products are the strongest[43]. As the input backoff setting is decreased, the intermodulation products caused by the nonlinear amplifier decrease and so does the difference between the losses introduced to the system over the two channel models. This would appear to indicate that if the amplifier were operating in a highly linear region where the intermodulation products were minimized, the relative performance between the TU and RA channels would be consistent with that measured in section 4.3.2.

Using the technique that was discussed in the previous section, the optimum operating point of this terrestrial amplifier was determined. Figure 4.30 contains the plot of output backoff minus loss versus input backoff for the Larcan SSPA for the TU channel and figure 4.31 is the equivalent plot for the RA channel. From these plots, it can be stated that regardless of the channel model, the optimum operating point of the Larcan amplifier was an input backoff of -7.0 dB.

Figure 4.27: The nonlinear characteristics of a Larcam UHF SSPA

Figure 4.28: The effects of a Larcam UHF SSPA on mode III over a TU channel with $\gamma_{\max} = 100\text{ Hz}$

Figure 4.29: The effects of the Larcas UHF SSPA on mode III over the RA channel with $\lambda_{\max} = 100$ Hz

Input Backoff (dB)	Output Backoff (dB)	Loss at BER=1E-4 (dB)	Output Backoff - Loss (dB)
-5.0	-0.67	2.50	-3.17
-6.0	-0.95	2.05	-3.00
-7.0	-1.20	1.50	-2.70
-8.0	-1.55	1.20	-2.75
-9.0	-2.05	0.80	-2.85

Table 4.11: Summarizing the losses resulting from nonlinear amplification by a Larcen UHF SSPA over a TU channel with $\lambda_{\max} = 100$ Hz

Input Backoff (dB)	Output Backoff (dB)	Loss at BER=1E-4 (dB)	Output Backoff - Loss (dB)
-5.0	-0.67	2.00	-2.67
-6.0	-0.95	1.60	-2.55
-7.0	-1.20	1.10	-2.30
-8.0	-1.55	0.80	-2.35
-9.0	-2.03	0.55	-2.58

Table 4.12: Summarizing the losses resulting from nonlinear amplification by a Larcen UHF SSPA over a RA channel with $\lambda_{\max} = 100$ Hz

Figure 4.30: Determining the optimum operating point for a Larcen UHF SSPA over a TU channel with $\lambda_{\max} = 100$ Hz

Figure 4.31: Determining the optimum operating point for a Larcen UHF SSPA over a RA channel with $\lambda_{\max} = 100$ Hz

4.3.5 Non-Ideal Filters and Nonlinear Amplification Results

This section is included to examine the effects of incorporating both non-ideal filters and a nonlinear amplifier into the simulation of the COFDM-based DAB system. If the losses introduced by these components are independent of each other, then their total loss should be the sum of their individual losses measured in sections 4.3.3 and 4.3.4. Should their total loss not be equal to the sum of their individual losses, it can be concluded that some dependency between the filtering and amplification exists. The intent here is to determine whether there is such a dependency and to observe what, if any, effect it has on the optimum operating point of each amplifier recommended in section 4.3.4.

For these simulations, the only filter used was an 8th order Chebychev I IIR design with 1 dB ripples. Section 4.3.3 concluded that of all of the filters examined, this design was the most appropriate choice for use with the experimental DAB system. In addition, simulation was only performed in mode III at three input backoff settings of each of the amplifiers studied. These settings included the optimum operating point evaluated in section 4.3.4 as well as settings 1 dB above and 1 dB below it.

In order to gain some insight into the potential non-ideal filter/nonlinear amplifier interaction, a brief literature survey found a few publications related to the effects of nonlinear amplifiers. However, these papers[20][44][45] either included filtering in all, or none, of the simulations performed so it appeared that no standard reference for the comparisons being made here was readily available. To create such a reference, simulations were performed over an AWGN channel that incrementally included the

effects of non-ideal filtering, nonlinear amplification and finally both. The results of these simulations are summarized in section 4.3.5.1 and are intended to act only as a reference for sections 4.3.5.2 and 4.3.5.3.

4.3.5.1 AWGN Results

This section is used to summarize the performance of the DAB system over an AWGN channel. The bit error rate plots for all of these simulations are included in appendix 5 of this document but the relevant results are presented in tables 4.13 through 4.15. These tables allow the sum of the individual non-ideal filtering and nonlinear amplification losses to be compared to the total loss measured when both the Chebychev filter and a nonlinear amplifier were included in the system simulation. From them, it can be concluded that for the range of input backoffs considered, the combined loss is always greater than, or comparable to, the sum of the individual losses. While this behaviour does not seem unreasonable, it is difficult to identify its exact cause. At this point, any explanation would be purely speculative so none will be provided. However, these results clearly demonstrate that losses introduced by non-ideal filtering and nonlinear amplification are not additive.

In addition, the results which measured the combined effects of the non-ideal filtering and nonlinear amplification were used to evaluate the optimum operating point of each of the three amplifiers. For each of them, the optimum operating point determined here was consistent with those of section 4.3.4.

Input Backoff (dB)	Filter + Amp. Loss at BER=1E-4 (dB)	Combined Loss at BER=1E-4 (dB)	Output Backoff - Combined Loss at BER=1E-4 (dB)
-4.0	1.60	1.80	-2.55
-5.0	1.35	1.35	-2.45
-6.0	1.15	1.15	-2.75

Table 4.13: Summarizing the losses resulting from combining a TWTA and an 8th order Chebychev IIR filter with 1 dB ripples on mode III over an AWGN channel

Input Backoff (dB)	Filter + Amp. Loss at BER=1E-4 (dB)	Combined Loss at BER=1E-4	Output Backoff - Combined Loss at BER=1E-4 (dB)
-2.0	1.35	1.45	-1.85
-3.0	1.05	1.10	-1.60
-4.0	0.75	0.85	-1.85

Table 4.14: Summarizing the losses resulting from combining a SPAR L-Band SSPA and an 8th order Chebychev IIR filter with 1 dB ripples on mode III over an AWGN channel

Input Backoff (dB)	Filter + Amp. Loss at BER=1E-4 (dB)	Combined Loss at BER=1E-4 (dB)	Output Backoff - Combined Loss at BER=1E-4 (dB)
-6.0	1.40	1.65	-2.60
-7.0	1.20	1.30	-2.50
-8.0	0.90	1.10	-2.65

Table 4.15: Summarizing the losses resulting from combining a Larcas UHF SSPA and an 8th order Chebychev IIR filter with 1 dB ripples on mode III over an AWGN channel

4.3.5.2 Satellite Results

This section considers the joint effect of an 8th order Chebychev IIR filter with 1 dB ripples and either a TWTA or a SPAR L-Band SSPA on mode III over a satellite channel. To allow a direct comparison between these results and those of sections 4.3.3 and 4.3.4, the channel had $\lambda_{\max} = 100$ Hz, $K = -10$ dB and $\lambda_{dp} = 0$ Hz.

The curves of figure 4.32 considered the use of a TWTA while figure 4.33 incorporated a SPAR L-Band SSPA. These results are summarized in tables 4.16 and 4.17 and a behaviour similar to that observed over an AWGN channel can be detected. In both cases, the combined loss of the non-ideal filtering and nonlinear amplification was comparable to, or greater than the sum of the losses caused by them individually. Also of significance is that the optimum operating points of both amplifiers indicated here are consistent with those evaluated in section 4.3.4.

Figure 4.32: Combining the effects of an 8th order Chebychev IIR filter with 1 dB ripples and a TWTA on mode III over a SAT channel with $\lambda_{\max} = 100$ Hz, $K = -10$ dB and $\lambda_{dp} = 0$ Hz

Figure 4.33: Combining the effects of an 8th order Chebychev IIR filter with 1 dB ripples and a SPAR L-Band SSPA on mode III over a SAT channel with $\lambda_{\max} = 100$ Hz, $K = -10$ dB and $\lambda_{dp} = 0$ Hz

Input Backoff (dB)	Amp. + Filter Loss at BER=1E-4 (dB)	Combined Loss at BER=1E-4 (dB)	Output Backoff - Combined Loss at BER=1E-4 (dB)
-4.0	1.65	2.05	-2.80
-5.0	1.30	1.55	-2.65
-6.0	1.10	1.10	-2.75

Table 4.16: Summarizing the losses resulting from combining a TWTA with an 8th order Chebychev IIR filter with 1 dB ripples on mode III over a SAT channel with $\lambda_{\max} = 100$ Hz, $K = -10$ dB and $\lambda_{dp} = 0$ Hz

Input Backoff (dB)	Amp. + Filter Loss at BER=1E-4 (dB)	Combined Loss at BER=1E-4 (dB)	Output Backoff - Combined Loss at BER=1E-4 (dB)
-2.0	1.50	1.50	-1.90
-3.0	1.20	1.20	-1.70
-4.0	0.80	0.90	-1.90

Table 4.17: Summarizing the losses resulting from combining a SPAR L-Band SSPA with an 8th order Chebychev IIR filter with 1 dB ripples on mode III over a SAT channel with $\lambda_{\max} = 100$ Hz, $K = -10$ dB and $\lambda_{dp} = 0$ Hz

4.3.5.2 Terrestrial Results

The joint effects of a Larcan UHF SSPA and an 8th order Chebychev IIR filter with 1 dB ripples are presented here. Performance over both the TU and RA channels with $\lambda_{\max} = 100$ Hz was simulated with the results being shown in figures 4.34 and 4.35 and summarized in tables 4.18 and 4.19.

The results over the TU channel were somewhat surprising. As opposed to what was observed with the AWGN and satellite channels, the combined loss of filtering and amplification was actually less than the sum of the individual losses. Conversely, the RA channel simulations followed the same trend indicated by the AWGN and satellite results in that the combined loss is greater than the sum of the individual losses.

While these results do not aid in the identification of the cause of the non-ideal filter/nonlinear amplifier dependency, they do reinforce the conclusion that the effects of non-ideal filtering and nonlinear amplification are not purely additive. As was the case with the other two amplifiers, the presence of non-ideal filtering did not affect the optimum operating point of the Larcan amplifier evaluated in section 4.3.4.

Figure 4.34: Combining the effects of an 8th order Chebychev IIR filter with 1 dB ripples and a Larcan UHF SSPA on mode III over a TU channel with $\lambda_{\max} = 100$ Hz

Figure 4.35: Combining the effects of an 8th order Chebychev IIR filter with 1 dB ripples and a Larcan UHF SSPA on mode III over a RA channel with $\lambda_{\max} = 100$ Hz

Input Backoff (dB)	Filter + Amp. Loss at BER=1E-4 (dB)	Combined Loss at BER=1E-4 (dB)	Output Backoff - Combined Loss at BER=1E-4 (dB)
-6.0	2.15	1.75	-2.70
-7.0	1.60	1.45	-2.65
-8.0	1.30	1.15	-2.70

Table 4.18: Summarizing the losses resulting from combining a Larcan UHF SSPA with an 8th order Chebychev IIR filter with 1 dB ripples on mode III over a TU channel with $\lambda_{\max} = 100$ Hz

Input Backoff (dB)	Filter + Amp. Loss at BER=1E-4 (dB)	Combined Loss at BER=1E-4 (dB)	Output Backoff - Combined Loss at BER=1E-4 (dB)
-6.0	1.60	1.85	-2.80
-7.0	1.10	1.40	-2.60
-8.0	0.80	1.25	-2.80

Table 4.19: Summarizing the losses resulting from combining a Larcan UHF SSPA with an 8th order Chebychev IIR filter with 1 dB ripples on mode III over a RA channel with $\lambda_{\max} = 100$ Hz

Chapter 5

CONCLUSIONS AND RECOMMENDATIONS

5.1 Thesis Summary

The intent of this thesis was to investigate the effects of non-ideal filtering and nonlinear amplification on a COFDM-based DAB system. During the course of this study, a computer simulation program was developed that allowed the user to measure the performance of a COFDM-based DAB system over Rayleigh or Ricean fading channels. This DAB system could be simulated as either ideal or non-ideal. The ideal system assumed no filtering and a perfectly linear amplifier while the non-ideal system allowed the user to incorporate either filters into the transmitter and receiver, a nonlinear amplifier in the transmitter, or both.

Using this simulation tool, an extensive study of the effects of various FIR and IIR filter designs on the BER performance of the system was performed. A second study measured the losses introduced into the system by a nonlinear amplifier. Three separate amplifier models were examined and the optimum operating point of each was evaluated. The final investigation determined the combined effects of both non-ideal filtering and nonlinear amplification.

Upon completion of these studies, the following conclusions can be made:

- 1) The COFDM-based DAB system did not suffer any significant loss in performance from the nonlinear phase responses of the transmit and receive filters examined.
- 2) It can be extrapolated from 1 that IIR filters would be the most appropriate choice for use in the COFDM-based DAB system. Since a linear phase response is not required, the added computational complexity of an equivalent FIR design is not necessary.
- 3) Ripples found in the passband of Chebychev IIR filters can have a significant effect on the performance of the system if they are large. However, it was shown that these losses can be minimized if higher order filters with ripples limited to 1 dB in magnitude are used.
- 4) Of all of the filters examined, the 8th order Chebychev IIR design with 1 dB ripples provided the best balance between complexity, performance and losses introduced.
- 5) It was demonstrated that the nonlinear distortion present in most typical broadcasting amplifiers can have a substantial impact on the performance of the COFDM-based DAB system. The results of the simulations performed were tabulated to identify the tradeoff between performance loss

and the required input backoff of the three amplifiers studied. The optimum operating point of each amplifier was evaluated and they corresponded to input backoff settings of -5.0 dB for the TWTA (an output backoff of -1.10 dB), -3.0 dB for the SPAR L-Band SSPA (an output backoff of -0.50 dB) and -7.0 dB for the Larcan UHF SSPA (an output backoff of -1.20 dB).

- 6) The joint effects of both non-ideal filtering and nonlinear amplification were measured. It was observed that the effects of non-ideal filtering and nonlinear amplification are not additive. In most of the cases examined, the combined loss was greater than the sum of the individual losses. However, the optimum operating points recommended in 5 did not appear to be affected by the presence of non-ideal filters in the system.

5.2 Recommendations for Future Work

After considering the conclusions made by this research, it would appear that the following topics would be worth investigating.

- 1) The performance of various filter designs should be examined in the presence of adjacent channel interference. While the results included in this document provide insight into the effects of passband ripples and nonlinear phase responses, the importance of the steepness of the filter's

transition band cannot be fully determined until adjacent channel interference is incorporated into the study.

- 2) During this investigation, it was observed that the performance of the COFDM-based DAB system over satellite channels had an unusual dependence on the Doppler shift of the direct path. While performance measurements were made, no exact cause for this behaviour could be identified. It would be useful to perform a project that would determine the exact cause of this dependence and to study if any means of counteracting it exists.
- 3) While this thesis demonstrated that the effects of non-ideal filtering and nonlinear amplification are not purely additive, it was unable to identify the exact cause for this behaviour. Determining what is responsible for this occurrence may identify specific combinations of non-ideal filter and nonlinear amplifier characteristics that should be avoided because of the introduction of excessive performance loss.

Appendix 1

SIMULATING GAUSSIAN NOISE

For the work done in this thesis, the gaussian noise was generated using the Box-Muller method presented in [46]. It states that if U_1 and U_2 are two independent variables uniformly distributed in the interval $[0,1]$, then two independent gaussian random variables can be created using the transform:

$$X = [-2\ln(U_1)]^{1/2}\cos(2\pi U_2) \quad (\text{A1.1})$$

$$Y = [-2\ln(U_1)]^{1/2}\sin(2\pi U_2) \quad (\text{A1.2})$$

Both X and Y will be normalized to have a mean of zero and a variance of one.

In the noise generator software, both U_1 and U_2 were produced using a recursive algorithm also presented in [46]. This algorithm, known as the power residue method, is defined by

$$X(k) = [\alpha X(k-1)] \text{ modulo } M, \quad k = 1,2,3,\dots \quad (\text{A1.3})$$

$$\text{and } U(k) = X(k) / M \quad (\text{A1.4})$$

It can be seen that $X(k)$ is distributed uniformly in the interval $[0,M]$ and $U(k)$ is uniformly distributed in the interval $[0,1]$. It was found that good gaussian statistical properties were obtained when $\alpha = 16807$ and $M = 2^{31} - 1$. Plots of the resulting gaussian pdfs are shown in figures A1.1 and A1.2.

Figure A1.1: The real gaussian noise pdf

Figure A1.2: The imaginary gaussian noise pdf

Appendix 2

VITERBI DECODING

In order to verify the proper operation of the convolutional encoder and the Viterbi decoder, several tests were run using a convolutionally encoded BPSK signal transmitted over an AWGN channel. The convolutional codes used in the tests all had $k=1$, $n=2$ and a constraint length of either 3, 5, or 7. Both hard and soft decision decoders were tested, and all had a path memory of 32. The soft decision decoder used an 8 level linear quantizer (optimum for AWGN[29]) and a linear metric assignment as discussed by Clark in [29]. These system parameters allowed the comparison of the test results to those published by Heller and Jacobs. Their results were reproduced in [28] by Lin and Costello, and are compared to the test results in figures A2.1 and A2.2. The legends of these two figures use the abbreviations of "IVD" for the implemented Viterbi decoder and "H & J" for Heller and Jacobs.

Figure A2.1: Hard decision Viterbi decoding of BPSK

Figure A2.2: Soft decision Viterbi decoding of BPSK

Appendix 3

FILTER FREQUENCY RESPONSES

Enclosed in this appendix are plots of the frequency responses of the twelve filter designs considered during the study of the effects of non-ideal filtering on the performance of the COFDM-based DAB system. They are presented in the order that they were studied in the non-ideal filtering case study.

Figure A3.1: The frequency response of the 20 tap WFIR filter

Figure A3.2: The frequency response of the 40 tap WFIR filter

Figure A3.3: The frequency response of the 60 tap WFIR filter

Figure A3.4: The frequency response of the 8th order Butterworth IIR filter

Figure A3.5: The frequency response of the 10th order Butterworth IIR filter

Figure A3.6: The frequency response of the 12th order Butterworth IIR filter

Figure A3.7: The frequency response of the 4th order Chebychev IIR filter with 1dB ripples

Figure A3.8: The frequency response of the 6th order Chebychev IIR filter with 1 dB ripples

Figure A3.9: The frequency response of the 8th order Chebyshev IIR filter with 1 dB ripples

Figure A3.10: The frequency response of the 4th order Chebyshev IIR filter with 2 dB ripples

Figure A3.11: The frequency response of the 6th order Chebychev IIR filter with 2 dB ripples

Figure A3.12: The frequency response of the 8th order Chebychev IIR filter with 2 dB ripples

Appendix 4

COFDM SIGNAL STATISTICS

This appendix contains a reproduction of the statistical analysis that was performed by M. Le on a COFDM signal over one data frame (24 ms). The CDFs of the instantaneous power were calculated for both operating modes and were found to be virtually identical. These CDFs are plotted on the following page in figure A4.1.

Figure A4.1: The CDFs of the instantaneous power of modes II and III

Appendix 5

AWGN CHANNEL SIMULATION RESULTS

This appendix holds the BER performance plots related to the COFDM-based DAB system operating in mode III over a AWGN channel. Figure A5.1 demonstrates the effect of introducing 8th order Chebychev IIR filters with 1 dB ripples into the system. Figures A5.2 through A5.4 indicate the losses resulting from using the three nonlinear amplifiers examined while figures A5.5 through A5.7 present the combined effects of non-ideal filtering and nonlinear amplification. A summary of these results can be found in section 4.3.5.1 of this thesis.

Figure A5.1: The effects of an 8th order Chebychev IIR filter with 1 dB ripples on mode III over an AWGN channel

Figure A5.2: The effects of a TWTA on mode III over an AWGN channel

Figure A5.3: The effects of a SPAR L-Band SSPA on mode III over an AWGN channel

Figure A5.4: The effects of a Larcen UHF SSPA on mode III over an AWGN channel

Figure A5.5: Combining the effects of an 8th order Chebychev IIR filter with 1 dB ripples and a TWTA on mode III over an AWGN channel

Figure A5.6: Combining the effects of an 8th order Chebychev IIR filter with 1 dB ripples and a SPAR L-Band SSPA on mode III over an AWGN channel

Figure A5.7: Combining the effects of an 8th order Chebychev IIR filter with 1 dB ripples and a Larcen UHF SSPA on mode III over an AWGN channel

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