Difficulties with NTSC Compatible Quadrature Digital Data Transmission

by

Paul Eric Beckmann

Submitted to the
Department of Electrical Engineering and Computer Science
in partial fulfillment of the requirements
for the degrees of
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and
Master of Science
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Difficulties associated with transmitting digital data in quadrature with an NTSC signal are presented. For such a system to be practical, the data must be backward compatible and rugged enough to allow recovery in a wide reception area. Two partial response signaling and one NRZ code were tested and none provided compatible operation. The main difficulty dealt with the injection level of the data. A high injection level is needed for robust data transmission, but such a high level caused unacceptable video interference. Audio and video demodulation is analyzed in detail and the effects of quadrature modulation are revealed. Software simulations and experimental hardware tests were used to study backward compatibility. Of all video detectors tested, quasi-synchronous detectors were most affected by quadrature modulation. Although these results do not rule out a compatible system, they do demonstrate the inherent difficulties associated with applying quadrature modulation to compatible television transmission.

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Chapter 1

Introduction

The purpose of this research is to study the difficulties associated with sending modulated digital data in quadrature with an NTSC\(^1\) signal. This quadrature signal must be backward compatible with current television receivers; not causing noticeable video or audio interference. Also, data transmission must be robust enough to allow reliable data recovery in a wide reception area. Through analysis, simulation and hardware tests, the difficulties associated with quadrature transmission applied to NTSC are exposed.

The ability to send digital data compatibly in an NTSC signal has a variety of practical applications. Digital stereo sound with quality surpassing the current stereo standard could be added. Future enhanced definition television systems could use the channel for sending image augmentation signals such as vertical motion detail. Also the channel would be applicable for high speed Teletext, with a data rate approximately 12 times higher than current implementations.

1.1 Standard Television Transmission

The conventional NTSC signal is vestigial side band amplitude modulated (VSB-AM) and has the frequency spectrum shown in Figure 1.1. Each channel is restricted to a 6 MHz bandwidth. The luminance signal which is amplitude modulated originally extended \(\pm 4.5\) MHz around the visual carrier. The lower 3.25 MHz of the luminance signal is removed leaving a vestigial

\(^1\)NTSC is the television transmission standard used in the United States.
side band. The chrominance signal has a baseband bandwidth of 1.5 MHz and is added to the luminance on a subcarrier 3.579545 MHz above the visual carrier. The chrominance is also VSB-AM, having part of its upper sideband removed. The proposed system would add a signal, with a bandwidth less than 3 MHz, centered around the visual carrier and 90 degrees out of phase with it.

Monophonic television sound was upgraded to an MTS (multichannel television sound) system by the addition of supersonic subcarriers carrying additional audio signals. This composite signal is called the aural baseband and is shown in Figure 1.2. These signals are transmitted on a carrier located 4.5 MHz above the visual carrier. The aural baseband consisting of 4 subchannels frequency modulates this carrier with a maximum frequency deviation of 100 KHz. It contains three AM and one FM subchannels, in addition to a stereo pilot tone.

This research focused upon testing several digital data modulation schemes and analyzing them for compatibility. The problem was addressed using a combination of analytical, simulation and experimental tools. Analysis, primarily in terms of signal levels and data channel error rates, was used to establish general feasibility regions. Once a general range of parameters was identified, simulation methods were used to examine the data transmission performance and compatibility issues in detail. The experimental work was used to verify the simulations and explore issues not yet addressed.

Any changes proposed to this standard must be compatible with the entire population of television receivers. This includes black and white, color, monophonic and stereo receivers. Receiver design must also be taken into account because the added signal will affect synchronous, quasi-synchronous and envelope detectors differently.
Compatibility will be judged by video and audio signal to noise ratios (SNRs). These provide a convenient method of comparison but cannot be used to make the final decision. A sine wave disturbance in an audio channel is much more noticeable than white noise of the same power, even though both may yield the same SNR. Visual masking techniques which take advantage of certain properties of the human eye cannot be fairly evaluated by SNRs alone. Therefore, SNRs are used only to compare similar types of interferences, while subjective testing provides the final basis for establishing compatibility.

The simulation tools used were developed on a special purpose VAX computer known as the David Sarnoff Research Center Digital Video Facility (DVF). This facility provided real time sampling and display of video along with off line processing facilities. Video, audio and data modulation, transmission and demodulation were simulated.

The software simulations were verified and added insight gained by hardware test bed experiments. The test bed consists of a transmitter, a channel simulator, and several different demodulators. The previous software simulations were repeated in hardware, testing compatibility with commercial television receivers.

This study will be presented in several sections. General compatibility issues will be addressed first. Next, quadrature modulation applied to tele-
vision transmission and its effects on different receivers will be analyzed. An overview of digital data transmission systems will be presented followed by a description of the system chosen. The details of the software simulation will be explained followed by a discussion of the results obtained. Then the experimental hardware tests and their findings will be described. The study will then conclude with a summary of results and specific problems with quadrature modulation and its application to compatible television transmission.
Chapter 2

Quadrature Modulation

2.1 Introduction

Quadrature modulation was chosen as a means for transmitting digital data for several reasons. It offers a data rate much higher than time division methods such as Teletext. Interference into video should theoretically not occur for synchronous demodulators. Also as will be shown, spectrum space is available in an NTSC signal.

Quadrature modulation uses the symmetry of real signals to transmit two bandlimited signals in the same frequency bandwidth. The Fourier transform of a real signal has an even real part, and an odd imaginary part. When a real signal $x(t)$ is modulated by $\cos \omega t$, this symmetry is preserved not only around the origin (it must since it is a real signal), but also around the carrier. When a different real bandlimited signal $y(t)$ is modulated by $\sin \omega t$, this symmetry again holds around the origin. However, a different symmetry occurs around the modulating carrier. The real part of the Fourier transform is odd, and the imaginary part is even around the carrier. This symmetry is shown in Figure 2.1.

An arbitrary function can be uniquely divided into an even and odd component. Using this fact, $x(t) \cos \omega t$ and $y(t) \sin \omega t$ may be combined,

$$z(t) = x(t) \cos \omega t + y(t) \sin \omega t$$

and then correctly recovered. To recover them, take the even real part around the carrier as the real part of $x(t)$, and the imaginary odd part as the imaginary part of $x(t)$. Similarly, the odd real part around the carrier is the
Figure 2.1: Symmetry of Fourier transform of modulated real signals.
imaginary part of \( y(t) \) and the even imaginary part is the real part of \( y(t) \).
This demodulation process is achieved by multiplying \( x(t) \) by \( \cos \omega t \) and then bandlimiting to obtain \( \hat{x}(t) \). \( \hat{y}(t) \) is recovered by multiplying by \( \sin \omega t \) and then bandlimiting. \( x(t) \) is referred to as the inphase component, and \( y(t) \) as the quadrature component. The complete modulation/demodulation process is shown in Figure 2.2.

The symmetry properties of quadrature modulation show that double sideband transmission is necessary, otherwise the received signal cannot properly be demodulated. Looking at Figure 1.1, it is clear that the quadrature signal can only extend \( \pm 1.25 \) MHz around the picture carrier.

The design of television receivers further limits the bandwidth of the quadrature signal. Television receivers contain a Nyquist filter, show in Figure 2.3, followed by the demodulator. This filter aids in the recovery of the transmitted video by forming a linear roll off. When recovered by a synchronous demodulator, no distortion occurs. The two linear sections add to form a flat transfer characteristic. The Nyquist filter however destroys the symmetry needed for quadrature transmission. After passing through the filter, the data signal is no longer in quadrature with the video, and interference
Figure 2.4: Inverse Nyquist filter.

into the picture will occur.

This interference is overcome by the addition of an inverse Nyquist filter at the transmitter. The inverse Nyquist filter was developed by Matsushita for use in their compatible quadrature transmission system. They utilize the quadrature channel for sending time expanded side panels of an NTSC compatible wide aspect ratio television system.\(^1\) Its slope is opposite that of the Nyquist filter and is shown in Figure 2.4. A cascade of the inverse Nyquist and Nyquist filters produces a symmetric, parabolic shaped transfer characteristic around the picture carrier. This keeps the data in quadrature with the video, and when demodulated with a truly synchronous receiver, would produce no interference.

The inverse Nyquist filter also enables correct recovery of the data. At the receiver a synchronous quadrature demodulator would recover the data before the Nyquist filter. When demodulating with \(\sin(wt)\) the linear slopes of inverse Nyquist filter combine to form a flat transfer characteristic. The inverse Nyquist filter produces inphase and quadrature components, and only the quadrature component is used in recovery.

In addition to the inverse Nyquist filter at the transmitter, the quadrature system requires a bandpass data extraction filter at the receiver. This would eliminate interference due to the quadrature portion of the video signal and limit the amount of demodulated channel noise. The complete NTSC compatible quadrature transmission system is shown in Figure 2.5. The vestigial sideband filter (VSB) has no effect on the data; it has a flat transfer characteristic over the interval where data is located.

2.2 Previous work

Augmenting NTSC with a quadrature signal was been previously explored by the David Sarnoff Research Center (DSRC), Matsushita and Hitachi Ltd. This research took some of their results and applied them to the problem of sending digital data in quadrature.

Norm Hurst of DSRC carried out experiments to assess the degree to which modulation of the quadrature carrier interferes with reception of sound. He injected sine waves in quadrature and measured the resulting audio SNRs and stereo separation. It was concluded that the BTSC system is very sensitive; for frequencies that map into the BTSC spectrum (0-100 KHz) the quadrature carrier powers must be kept very low, -37 dB (2.3 IRE pp) at most.²

This work followed his recommendation of minimizing low frequency energy. Also, by using a broad spectrum modulation code instead of single frequency sine waves, it was thought that the audio interference could be reduced and higher injection levels used.

Matsushita has successfully used the quadrature channel to transmit high frequency luminance information³, and time expanded side panels⁴ compat-

---

²Hurst, Norm, Interference of Quadrature Picture Carrier Modulation with Multichannel Television Sound: Experimental Results, internal DSRC memo, Sept. 29, 1987.
ibly. In their experiments, the quadrature component extends ±1.25 MHz around the carrier and an inverse Nyquist filter matched to this bandwidth is employed. The quadrature component is not transmitted during horizontal and vertical blanking intervals.

Matsushita states that the performance with respect to video compatibility depends heavily upon matching the slope of the receiver Nyquist filter with the inverse Nyquist filter. Video crosstalk was -20 dB for envelope detectors and -35 to -40 dB for quasi-synchronous and synchronous receivers. Since almost all receivers 14 inches or over, and 90% of the receivers in Japanese homes use either quasi-synchronous or synchronous detectors, such a system would be compatible.\(^6\) In order to remain invisible\(^6\), the quadrature injection level had to be between -10 and -30 dB for quasi-synchronous and synchronous receivers, and -30 to -50 dB for envelope detectors. Compatibility with respect to audio reception was not discussed by Matsushita and it is likely that their system would cause unacceptable interference.

Hitachi has already developed a system for transmitting digital data in quadrature with the main picture carrier. A three level dicode modulation scheme and comb filters are used to reduce audio and video interference. For a reception area where the video SNR is at least 30 dB, 1-2 Mbits/sec of data can be transmitted with a BER of 10\(^{-4}\). As with Matsushita, the data occupies ±1.25 MHz around the picture carrier.\(^7\) This system also uses the inverse Nyquist filter, and tries to reduce low frequency quadrature energy.

Matsushita's and Hitachi's systems are both compatible with the receivers used in Japan. The Nyquist filter in these receivers has a cutoff frequency -1.25 MHz below the picture carrier compared with -0.75 MHz for US receivers. Low frequency video interference is most noticeable, and this extra MHz of bandwidth allowed them to spread the data further away from the picture carrier. In Matsushita's extended definition TV system, the high frequency luminance information is correlated with the video and thus much less visible than random data. The Japanese audio baseband has a smaller bandwidth, and is less sensitive to high frequency interference. All of these

\(^6\)Yasumoto, New Extended Definition TV using Quadrature Modulation of Picture Carrier with Reverse Nyquist Filter, 1987 IEEE International Conference on Consumer Electronics Digest of Technical Papers, pp. 80.

\(^7\)Noda, Tsutomu, Multiplexed Digital Audio Signal Compatible with NTSC Television, Internal memo, Consumer Products Research Center, Hitachi, Ltd.
factors contributed to the feasibility of their system.

2.3 Conclusion

In order to remain compatible, the quadrature data signal can only extend ±750 KHz around the picture carrier. This is due to the NTSC spectrum and the design of television receiver Nyquist filters. Also, findings by DSRC and Hitachi show that audio reception is very sensitive to low frequency modulation of the quadrature carrier and that this should be avoided. Further details concerning audio compatibility will be presented in Section 3.3.
Chapter 3

Effects of Quadrature Modulation on Compatibility

3.1 Introduction

The added quadrature signal will affect both video and audio reception and the distortion created depends upon the type of demodulator used. The operation of synchronous, quasi-synchronous and envelope video detectors in the absence of a quadrature signal will analyzed. Next their operation with a quadrature signal will be studied to reveal the distortion created. A similar analysis for intercarrier demodulation will be done, showing specifically how audio reception is affected.

3.2 Video Compatibility

Television transmission in the region where the quadrature signal is located, is double side band amplitude modulated and therefore the vestigial side band filter will not affect the following derivations. It is only necessary that the added signal be in quadrature with the video, and this is guaranteed by the combination of Nyquist and inverse Nyquist filters.

Baseband video signals range from -40 IRE during horizontal sync to 100 IRE for peak white level. In the following discussion, this will be modelled as ranging from 0 to 0.875 volts. Television is transmitted using negative modulation; instead of AM modulating the baseband video signal \(v(t)\), it is
inverted and an offset added: $1 - v(t)$ is sent. By using negative modulation, impulses added during transmission appear as black instead of white dots, which are visually much less annoying. The transmitted RF video signal will therefore be modelled as

$$S_{RF}(t) = [1 - v(t)] \cos(\omega_c t)$$

$$= \cos(\omega_c t) - v(t) \cos(\omega_c t)$$

(3.1) (3.2)

where $\omega_c$ is the carrier frequency and the first term represents the video carrier. With the addition of the quadrature signal $q(t)$ the transmitted signal becomes

$$S_{RF}(t) = \cos(\omega_c t) - v(t) \cos(\omega_c t) + q(t) \sin(\omega_c t).$$

(3.3)

This model is also adequate for representing the RF signal after the Nyquist filter, because the data is completely in quadrature with the video.

The RF television signal attains its maximum amplitude during peak of sync, and this is a known, fixed quantity. The quadrature injection level $I_Q$, is set relative to this level, and is defined as the transmitted rms power due to the quadrature signal compared with the power at peak of sync:

$$I_Q = 10 \log \left[ \frac{\sigma^2_{RF \text{ data}}}{P_{\text{sync}}} \right].$$

(3.4)

This is in effect the increase in power needed to transmit the additional data.

Television signal strength is measured as a carrier to noise ratio,

$$CNR_{\text{video}} = 10 \log \left[ \frac{P_{\text{sync}}}{\text{noise in 6 MHz bandwidth}} \right].$$

(3.5)

If the channel has a flat noise spectrum, Equation 3.5 becomes

$$CNR_{\text{video}} = 10 \log \left[ \frac{P_{\text{sync}}}{\eta B_{\text{video}}} \right]$$

(3.6)

where $\eta$ is the noise power density in the channel and $B_{\text{video}}$ the RF bandwidth of the video signal (6 MHz). $CNR_{\text{video}}$ is a relatively good measure of received picture quality.
Video signal to noise ratios used in the evaluation of television images are not true signal to noise ratios. Rather, the rms noise in the video image, \( \sigma_{vn} \), is compared with a 100 IRE signal.

\[
SNR_{video} = 20 \log \left( \frac{100 \text{ IRE}}{\sigma_{vn}} \right) = 20 \log \text{[100IRE]} - 20 \log [\sigma_{vn}] \tag{3.7}
\]

For example, if the same noise is added to a 100 IRE white field, and to a 10 IRE black field, these images would be judged as having the same video SNR. Therefore it is only necessary to determine the added noise caused by the quadrature signal to evaluate video SNRs.

### 3.2.1 Envelope Detectors

Envelope detectors are nonlinear devices which demodulate AM signals by following their envelope. They function correctly only if the transmitted signal is always positive, and this is the case with video. The quadrature signal distorts the RF envelope, and an envelope detector would return

\[
1 - \tilde{v}(t) = \sqrt{[1 - v(t)]^2 + q^2(t)}. \tag{3.8}
\]

\( q(t) \) is always much smaller than \([1 - v(t)]\) and the preceding equation can be approximated and reduced to a more meaningful form through the following steps:

\[
1 - \tilde{v}(t) = [1 - v(t)] \sqrt{1 + \frac{q^2(t)}{[1 - v(t)]^2}} \tag{3.9}
\]

\[
= [1 - v(t)] \left[ 1 + \frac{q^2(t)}{2[1 - v(t)]^2} \right] \tag{3.10}
\]

\[
= \left[ \frac{1 - v(t)}{\text{signal}} + \frac{q^2(t)}{2[1 - v(t)]} \right]. \tag{3.11}
\]

This equation demonstrates that the quadrature signal will always cause some distortion in an envelope detector. However the introduced distortion is small due to the \( q^2(t) \) term in the numerator. Also notice that when the amplitude of the quadrature signal doubles, the interference increases 4 fold.
3.2.2 Synchronous Detectors

The next type of video detector that will be considered is the synchronous detector and is shown in Figure 3.1. This detector locally generates a copy of the video carrier using a phase locked loop. The received signal is demodulated by this copy and then low pass filtered. This process yields demodulated video.

The case when no quadrature signal is present will be considered first. Phase locked loops are not completely accurate and have some phase error $\phi$ associated with them. With this phase error, the recovered video becomes

$$1 - \hat{v}(t) = [1 - v(t)] \cos(\omega_c t) \times 2 \cos(\omega_c t + \phi)$$

(3.12)

$$= [1 - v(t)][\cos(2\omega_c t + \phi) + \cos \phi].$$

(3.13)

The first term in brackets is removed by the low pass filter leaving only

$$1 - \hat{v}(t) = [1 - v(t)] \cos \phi$$

(3.14)

at the output of the demodulator. This is the original video scaled by $\cos \phi$, no noise has been added. $\phi$ is usually only a few degrees and changes slowly, permitting automatic gain control circuits to correct for these variations in scale.

When the quadrature signal is added, the received signal before the demodulator becomes

$$[1 - v(t)] \cos(\omega_c t) + q(t) \sin(\omega_c t).$$

(3.15)

The output of the synchronous detector can be found by repeating the above steps with the new signal,

$$1 - \hat{v}(t) = [1 - v(t)] \cos \phi + q(t) \sin \phi.$$

(3.16)
Figure 3.2: Block diagram of quasi-synchronous demodulation.

The output is a fraction of the video plus a fraction of the quadrature signal. When the demodulator is working properly, no video interference will occur. Video interference in synchronous receivers depends on the accuracy of the demodulator, and this varies between receivers. Also video SNRs due to data interference do not depend upon the instantaneous video amplitude as in envelope detectors but only on the quadrature amplitude.

3.2.3 Quasi-synchronous Detectors

Quasi-synchronous demodulation involves several steps and is more complicated to analyze than the previous two methods. The process is shown in Figure 3.2. Most commercial quasi-synchronous receivers use an integrated circuit manufactured by Toshiba and the exact design of this chip is not known. In theory, the visual carrier is extracted by a band pass filter, and then clipped, removing all amplitude modulation. This clipped signal resembles a square wave with frequency and phase equal to that of the carrier. This square wave is filtered, generating a reference cosine wave and this is used to perform the actual demodulation.

In practice, the carrier extractor filter has a bandwidth of roughly 1 MHz which is wide enough to pass most of the data. Quasi-synchronous receivers will function similar to an envelope detector for these low frequencies. Thus interference is expected in these receivers. After the carrier is clipped, the only information remaining is the location of the zero crossings, and these are affected by the quadrature signal. The effects on the final demodulated video are not obvious, but the zero crossings will be most perturbed when the ratio of video to data \( \frac{1 - v(t)}{q(t)} \) is smallest, and this occurs when white video is sent. Equation 3.11 shows that video interference for envelope detectors is also greatest when white video is sent.

It was originally believed that with the addition of the quadrature signal,
synchronous detectors would perform the best, quasi-synchronous next best, and envelope detectors worst. This presumption came from the known performance of these detectors in demodulating video. Hardware tests with the added quadrature signal however, showed that quasi-synchronous receivers were most affected and had the worst signal to noise performance.

3.3 Audio Compatibility

Television audio is recovered in all receivers by a process known as intercarrier sound demodulation shown in Figure 3.3. In order to understand the interference produced by the quadrature signal, standard FM modulation and demodulation must be understood. In FM modulation, the instantaneous frequency of the IF (intermediate frequency) signal is a linear function of the information signal \( f(t) \). Thus

\[
\omega(t) = \omega_0 + K_{FM} f(t)
\]  
(3.17)

where \( \omega_0 \) and \( K_{FM} \) are positive constants.

The carrier frequency \( \omega_0 \) has a value of 4.5 MHz in television transmission, and \( K_{FM} \) is the FM deviation constant. The instantaneous phase of this signal is the integral of the frequency,

\[
\phi(t) = \omega_0 t + K_{FM} \int f(t) \, dt
\]  
(3.18)

and the transmitted sinusoid is

\[
S_{FM}(t) = A \cos(\omega_0 t + K_{FM} \int f(t) \, dt)
\]  
(3.19)

where \( A \) is an arbitrary nonzero scaling constant.

At the receiver, the FM signal is demodulated in 2 stages. First the instantaneous phase is extracted

\[
\hat{\phi}(t) = \omega_0 t + K_{FM} \int f(t) \, dt
\]  
(3.20)

and then this is differentiated leaving

\[
\hat{\omega}(t) = \omega_0 + K_{FM} f(t).
\]  
(3.21)
Thus the original information signal can be recovered.

In television broadcast the IF sound originally at 4.5 MHz is modulated and added to the modulated video at the diplexor. The sound carrier is placed 4.5 MHz above the video carrier and they are transmitted together. At the receiver, a bandpass filter with a center frequency of \( \omega_c \) isolates the picture carrier, \( \cos(\omega_c t) \), while another at \( \omega_c + 4.5 \text{ MHz} \) isolates the RF sound. These two signals are mixed,

\[
= \cos(\omega_c t) \times \cos(\omega_c t) S_{FM}(t) \tag{3.22}
\]

\[
= \frac{1}{2} \left[ \cos(2\omega_c t) + S_{FM}(t) \right] \tag{3.23}
\]

and then low pass filtered leaving

\[
\frac{1}{2} S_{FM}(t). \tag{3.24}
\]

Thus the original FM signal can be recovered and demodulated by the steps described above.

The sound IF signal is derived by mixing to eliminate the effects of synchronous phase modulation and drift of picture and sound carriers caused by instability in various local oscillators in the transmission path, particularly
in cable TV set-top converters.\textsuperscript{1} Any extraneous modulation incurred during the transmission path will affect the picture and sound carriers equally. Intercarrier demodulation will remove this since the process is concerned with the difference in frequency between the two carriers, and not with their absolute frequencies.

When the quadrature signal is added, the output of the picture carrier extractor band pass filter is no longer a pure sinusoid, but contains phase modulation. The output of this filter becomes

\[
\cos[\omega_c t + P(t)]
\]

(3.25)

where \(P(t)\) is the instantaneous phase deviation of the picture carrier given by

\[
P(t) = \tan^{-1}\left(\frac{q(t)}{1 - v(t)}\right).
\]

(3.26)

This is used in the intercarrier demodulator, and after mixing and low pass filtering yields

\[
\hat{S}_{FM}(t) = \frac{1}{2} \cos[\omega_0 t + K_{FM} \int f(t) dt + P(t)]
\]

(3.27)

as input to the FM demodulator.

The FM demodulator first extracts the phase of the signal,

\[
\hat{\phi}(t) = \omega_0 t + K_{FM} \int f(t) dt + P(t)
\]

(3.28)

and then differentiates to recover the signal

\[
\ddot{\omega}(t) = \omega_0 + K_{FM} f(t) + \frac{d}{dt} P(t).
\]

(3.29)

Comparing this with Equation 3.21, you see that there is an extra term \(\frac{d}{dt} P(t)\) at the output of the demodulator. This represents distortion caused by the quadrature signal. This is additive noise, independent of the audio signal; it depends only on the current video and quadrature signals.

\textsuperscript{1}Gibson, \textit{Compatibility of Quadrature Modulation of the TV Picture with Regard to Audio Reception with Existing Mono and MTS Receivers}, DSRC Internal Memo, March 31, 1988, p. 7.
By expanding this term, the relationship between the quadrature and video signals and the induced noise can more clearly be seen. This will aid in the characterization of the induced noise and show the effect of changing the amplitude of the quadrature signal. Starting with Equation 3.29 and expanding,

\[
\text{noise} = \frac{d}{dt} P(t) \tag{3.30}
\]

\[
= \frac{d}{dt} \left[ \tan^{-1} \frac{q(t)}{1 - v(t)} \right] \tag{3.31}
\]

\[
= \frac{[1 - v(t)] \dot{q}(t) - q(t) \dot{v}(t)}{q(t)^2 + [1 - v(t)]^2}. \tag{3.32}
\]

If \( q(t) \ll v(t) \), this can be approximated as

\[
= \frac{[1 - v(t)] \dot{q}(t) - q(t) \dot{v}(t)}{[1 - v(t)]^2}. \tag{3.33}
\]

The above equation demonstrates that when the quadrature signal is much smaller than the video signal (this is always the case), the noise amplitude is a linear function of the amplitude of the quadrature component. When the amplitude of the quadrature component doubles, the noise amplitude also doubles. The noise is also inversely proportional to the video amplitude. Doubling the video signal, halves the amplitude of the noise. Therefore, the audio reception will be most affected when the amplitude of the RF video signal is smallest. Again, this worst case is for white fields.

In general, the video signal is mainly low frequency energy and from Equation 3.33 one sees that if \( v(t) \) is constant, there is a direct mapping from quadrature frequencies to audio frequencies. As seen in Figure 1.2, audio information is restricted to below 100 KHz in frequency. Therefore to reduce audio interference, the quadrature signal should have little or no energy below 100 KHz. This criterion will aid in selecting a suitable channel code for transmitting the data. Also, since the interference depends upon the derivative of the instantaneous phase, high frequency subchannels in the audio baseband will be affected more.
3.4 Conclusion

The above derivations characterize the quadrature interference and yield several useful design criteria for reducing it. Video interference cannot be eliminated and should be greatest in envelope detectors, rising as the square of the injection level. In all cases, audio and video reception is most impaired when the ratio $\frac{1-p(t)}{q(t)}$ is smallest and this occurs when white video is transmitted. Also, to minimize interference into audio, the quadrature signal should have little or no energy below 100 KHz.
Chapter 4

Digital Communications Systems

4.1 Introduction

A general digital data communications system consists of two parts, a transmitter and a receiver. The transmitter takes discrete symbols and converts them into a signal which can propagate easily through the channel. This mapping is called a channel code. The job of the receiver is to perform the inverse mapping: given a signal it must decide which symbol was sent. The transmitter and receiver are closely related, and their design depends upon the type of data sent and characteristics of the channel. The channel code is carefully chosen to compensate for defects in the transmission channel.

The transmitter can usually be broken into two different functions. The first is a channel encoder which takes discrete time symbols (in this case binary digits) and converts them to a continuous time signal. The encoder will be modelled as consisting of a pulse generator and data shaping filter. The pulse generator handles the discrete time to continuous time conversion. It outputs regularly spaced impulses whose amplitudes are determined by the discrete time symbols. These impulses are fed to a data shaping filter which specifies the spectrum of the data. The waveform transmitted for each data symbol is determined by the impulse response $h(t)$ of this filter. This filter must be specifically tailored to the impulse rate of the pulse generator. The second component of the transmitter is the modulator. The output of the
data shaping filter is baseband and the modulator converts this to frequencies which can propagate through the channel. In this application, the frequency of this modulator equals the picture carrier frequency, and the phase differs by 90 degrees.

The receiver’s task is complicated by the fact that it has to isolate the signal from data being transmitted at other frequencies. Its first component is an RF data extraction filter. This serves to isolate the transmitted signal, and remove excess noise. The bandwidth of this filter must be wide enough to pass all of the data, and should be flat in the region of interest. The signal can then be demodulated by a sinusoid whose frequency and phase are the same as the sinusoid used in the transmitter. The output of this section is a baseband signal. The signal is then filtered by a matched filter. This filter is “matched” to the spectrum of the data shaping filter and optimally removes white noise incurred in the transmission channel. Lastly, the output of the matched filter is fed to a data slicer. This subsection compares the input signal at sampling times to a threshold and determines which symbol was transmitted. The output of the data slicer is a discrete data stream. The complete communication system is shown in Figure 4.1.

The overall system performance depends upon the transfer function from the output of the pulse generator to the input of the data slicer. This system transfer function can be thought of as the cascade of three filters: pulse shap-
Sampling Point

Figure 4.2: Eye pattern for an ideal Nyquist channel.

...ing, data extraction and matched. Ideally, in order to have no intersymbol interference, the system should have a Nyquist transfer function.

The requirements of a Nyquist filter are most easily seen in the time domain. The filter is fed with impulses equally spaced in time, and the output is a weighted sum of impulse responses, also equally spaced in time. For the symbols not to interfere with each other, the impulse response of these filters must be zero at all multiples of the symbol spacing except at the sampling instant. If the system has a Nyquist characteristic, and no noise is added during transmission, the data slicer should always be able to differentiate between symbols.

4.2 Measuring Performance

Real systems are usually not ideal, and data reception errors have four main sources: non Nyquist transfer functions (results in intersymbol interference), noise added during transmission, timing errors in the data slicer, and crosstalk between the quadrature and inphase signals. The effects of the first three of these imperfections can be quantified and analyzed using eye patterns.

An eye pattern is a diagram formed by repeatedly overlapping the received symbols at the input to the data slicer. Its two most important features are the separation between different symbols at the sampling instant (referred to as the eye height) and the rate at which the eye height decreases. An eye pattern for an ideal Nyquist channel is shown in Figure 4.2. Deviations away from an ideal channel cause the eye to close and become less defined. See Figure 4.3.
Along with an eye pattern are associated thresholds which divide the space of the received signal into different symbol regions. For an N level code, N-1 thresholds are needed. The function of the data slicer is to determine which region the received signal falls into by threshold comparisons, and then return the corresponding symbol.

A system's performance, measured by its bit error rate (BER), can be determined from its eye pattern. An error occurs when the channel noise is so great that the received signal is pushed over a threshold and interpreted incorrectly by the data slicer.

Calculating the BER is a two step process. First the signal amplitude probability density at the sampling instant $p_{S_i}(S)$ for each symbol must be known. This is equivalent to slicing the signal vertically through the sampling instant and recording the frequency of signal amplitudes. The eye pattern in Figure 4.3 has the density shown in Figure 4.4. The next step is to integrate this density multiplied by the conditional probability of error given symbol $S_i$ and amplitude $s$, and multiply by the probability that $S_i$ occurs.

$$\text{Prob}(\text{error}) = \sum_{i=1}^{N} \text{Prob}(\text{symbol} = S_i) \int_{\text{symbol} S_i} p_{S_i}(s) p_e(E \mid S_i, s) \, ds_i \quad (4.1)$$

$p_e(E \mid S_i, s)$ is the probability that the noise pushes symbol $i$ over a threshold given that the signal had amplitude $s$. For a two level code with amplitudes $a$ and $b$ corresponding to different symbols and a threshold $T$, this occurs when,

$$\text{noise} < T - a \quad \text{for symbol} \ S_A$$

and $$\text{noise} > T - b \quad \text{for symbol} \ S_B.$$
Figure 4.4: Signal distribution at sampling instant for non ideal Nyquist channel.

The probability of these events can be found by integrating the probability density for the noise, \( p_n(N) \),

\[
\text{Prob}(\text{noise} < T - a) = \int_{-\infty}^{T-a} p_n(s) \, ds \\
\text{and} \quad \text{Prob}(\text{noise} > T - b) = \int_{T-b}^{\infty} p_n(s) \, ds.
\]

The channel noise is modelled as additive Gaussian white noise, and the above two equations reduce to

\[
\text{Prob}(\text{noise} < T - a) = \Phi \left[ \frac{T-a}{\sigma_n} \right] \\
\text{and} \quad \text{Prob}(\text{noise} > T - b) = \Phi \left[ \frac{b-T}{\sigma_n} \right]
\]

and Equation 4.1 reduces to

\[
\text{BER} = \text{Prob}(\text{symbol}=S_A) \int_{S_A} p_{s_A}(s) \Phi \left[ \frac{T-s}{\sigma_n} \right] \, ds + \\
\text{Prob}(\text{symbol}=S_B) \int_{S_B} p_{s_B}(s) \Phi \left[ \frac{s-T}{\sigma_n} \right] \, ds \tag{4.2}
\]

for binary symbols. \( \Phi \) is the normalized Gaussian probability distribution function,

\[
\Phi(x) = \int_{-\infty}^{x} \frac{e^{-t^2}}{\sqrt{2\pi}} \, dt \tag{4.3}
\]
For codes with more than two symbols, the BER formula is quite similar, but errors occur if the noise pushes the signal over an upper or below a lower threshold. One must then be careful to integrate over the positive and negative tails of the noise PDF in determining the BER.

The normalized Gaussian distribution function is nonlinear and decays rapidly. The signal sample nearest to a threshold contributes most to the BER given in Equation 4.2 and it is reasonable and much simpler to model an eye pattern by its minimum eye height $\epsilon$ at the sampling instant. For equally likely symmetric symbols $S_A$ and $S_B$, Equation 4.2 reduces to

$$\text{BER} = \frac{1}{2} \Phi \left[ -\frac{\epsilon/2}{\sigma_n} \right] + \frac{1}{2} \Phi \left[ -\frac{-\epsilon/2}{\sigma_n} \right]$$

(4.4)

$$\text{BER} = 1 - \Phi \left[ \frac{\epsilon/2}{\sigma_n} \right].$$

(4.5)

This approximation is also valid for multilevel codes. The smallest noise disturbance needed to confuse the data slicer also equals $\frac{1}{2} \epsilon$.

Equation 4.2 can be evaluated if the variance of the demodulated noise, $\sigma_n^2$, is known. $\sigma_n^2$ depends on the SNR of the demodulated data and will be developed next.

As mentioned earlier, the quality of television reception is measured as a carrier to noise ratio, and this serves as a good starting point for obtaining $\text{SNR}_{data}$:

$$\text{SNR}_{data} = CNR_{video} + I_Q + 10 \log \left[ \frac{B_{video}}{B_{data}} \right] + 10 \log \left[ \frac{\sigma_{data}^2}{\sigma_{RF data}^2} \right].$$

(4.6)

By expanding terms and cancelling,

$$= 10 \log \left\{ \left[ \frac{P_{sync}}{\eta B_{video}} \right] \left[ \frac{\sigma_{RF data}^2}{P_{sync}} \right] \left[ \frac{B_{video}}{B_{data}} \right] \left[ \frac{\sigma_{data}^2}{\sigma_{RF data}^2} \right] \right\}$$

(4.7)

$$= 10 \log \left[ \frac{\sigma_{data}^2}{\eta B_{data}} \right]$$

(4.8)

$$\text{SNR}_{data} = 10 \log \left[ \frac{\sigma_{data}^2}{\sigma_n^2} \right]$$

(4.9)

one sees that this is the desired form. $B_{data}$ is the RF bandwidth of the data. The last term of Equation 4.6 is the ratio of demodulated data power to RF data power and will be referred to as $R_D$. The value of this parameter varies among channel codes and is usually a fraction of a dB.

35
4.3 Channel Codes

NRZ and partial response channel codes were implemented and a description of each code and its defining characteristics are explained next. The combination of transmitter data filter and receiver matched filter should have a Nyquist response, and this was evenly distributed between these filters.

A second order RLC filter centered around the data and with a 1.5 MHz bandwidth was chosen as the RF data filter. This filter distorted the eye patterns slightly and more care could have gone into its design. However, from the ideal eye patterns, an upper limit on the performance with an optimal RF filter can be calculated.

4.3.1 NRZ

An NRZ channel code using a raised cosine transfer function was tested. Raised cosine filters have a Nyquist response defined by the following impulse response,

\[ h(t) = \frac{\sin(2\pi ft)}{2\pi ft} \cos[2\pi f (1 + \alpha)] \frac{1}{1 - [4f(1 + \alpha)t]^2}. \]  

(4.10)

\(\alpha\) is the roll off factor and \(f\) the cutoff frequency is twice the symbol rate. The filter tested had a roll off \(\alpha = 0.5\) and a symbol rate of 954,545 symbols/sec.\(^1\) The frequency and impulse responses for this filter are shown in Figure 4.5. Its eye pattern is shown in Figure 4.6. All eye patterns are scaled to have an average data power of 1.

4.3.2 Partial Response

NRZ is an example of pulse amplitude modulation (PAM) and signal waveforms are constrained in that they should not cause any intersymbol interference (ISI). In partial response signaling (PRS) systems, this constraint is relaxed, and a controlled amount of ISI is permitted. Since the ISI is known, its effect can be removed.\(^2\)

\(^1\)The hardware used to perform the tests samples at 4 times the color subcarrier frequency. Symbol rates are all integer fractions of this sampling frequency.

Figure 4.5: Frequency and impulse responses for NRZ channel code.
Figure 4.6: Eye pattern for NRZ channel code.
PRS has several advantages and disadvantages over traditional PAM. PRS waveforms have multilevel eye patterns making them harder to recover and more sensitive to random noise. Symbol precoding and decoding which is required to halt error propagation, adds to system complexity and cost.

The main merit of PRS is that by controlling ISI, the data spectrum can be shaped to compensate for a nonideal channel. Low frequency quadrature signals were known to cause interference and several PRS codes have a spectral null at DC making them suitable for this application.

At the transmitter, PRS waveforms are constructed from a weighted sum of $\sin{\frac{\pi}{T}}(t - nT)$ functions delayed at multiples of the symbol rate$^3$,

$$h(t) = \sum_{n=0}^{N-1} f_n \frac{\sin{\frac{\pi}{T}}(t - nT)}{\frac{\pi}{T}(t - nT)}.$$  \hspace{1cm} (4.11)

This is realized by cascading a tapped delay line with an ideal LPF. The tapped delay line is specified by its system polynomial

$$F(z) = \sum_{n=0}^{N-1} f_n z^{-n}$$  \hspace{1cm} (4.12)

and this determines the overall system response.

Two PRS waveforms with DC spectral nulls were tested. The first one which will be referred to as PRS1 had the following system polynomial,

$$F(z) = 1 - z^{-1}.$$  \hspace{1cm} (4.13)

Its frequency and impulse responses are shown in Figure 4.7 and it has the three level eye pattern seen in Figure 4.8. The second system tested, PRS2, was characterized by

$$F(z) = 1 - z^{-1} - z^{-2} + z^{-3}$$  \hspace{1cm} (4.14)

and its frequency and impulse responses and five level eye pattern are seen in Figures 4.9 and 4.10.

Two important characteristics for the channel codes are minimum eye height and $R_D$. A summary of these values is shown in Table 4.1. In addition, the maximum eye height for an ideal channel is also listed. The maximum eye height can be used to determine a lower limit on the BER.

$^3$Ibid., p. 923.
Figure 4.7: Frequency and impulse responses of PRS1 channel code.

<table>
<thead>
<tr>
<th>Channel Code</th>
<th>$\epsilon$ (volts)</th>
<th>$\epsilon_{opt}$ (volts)</th>
<th>$R_D$ (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>NRZ</td>
<td>1.8</td>
<td>2.1</td>
<td>-0.62</td>
</tr>
<tr>
<td>PRS1</td>
<td>1.1</td>
<td>1.4</td>
<td>-0.76</td>
</tr>
<tr>
<td>PRS2</td>
<td>0.8</td>
<td>1.0</td>
<td>-0.61</td>
</tr>
</tbody>
</table>

Table 4.1: Channel code characteristics.
Figure 4.8: Three level eye pattern of PRS1 channel code.
Figure 4.9: Frequency and impulse responses of PRS2 channel code.
Figure 4.10: Five level eye pattern of PRS2 channel code.
4.4 Video Crosstalk

Section 4.2 dealt with random errors caused by additive noise in the transmission channel. Errors also occur when the data is corrupted by video crosstalk. Video crosstalk occurs when the quadrature demodulator is out of phase and a portion of the video signal is demodulated along with the data. The effects of this crosstalk are best modelled as a decrease in eye height.

In the analysis which follows, the demodulated data is scaled to have an average power of 1. From the definition of $SNR_{data}$,

$$\sigma_n = 10^{-\frac{SNR_{data}}{20}}$$

and the approximate BER may be found from Equation 4.5.

$$BER = 1 - \Phi \left[ \frac{\epsilon}{2} 10^{\frac{SNR_{data}}{20}} \right].$$

This is plotted in Figure 4.11 for varying eye heights and data signal to noise ratios. Lines on the graph represent points of constant $SNR_{data}$.

When a phase error of $\phi$ degrees occurs, the demodulated signal is,

$$\tilde{d}(t) = d(t) \cos \phi + [1 - v(t)] \sin \phi.$$
Figure 4.12: Maximum crosstalk for varying demodulator phase errors.

Scale this so that is data component has an average power of 1,

\[
\tilde{\sigma}(t) = \frac{d(t)}{\sigma_{data}} + \frac{[1 - v(t)]}{\sigma_{data}} \tan \phi. \quad (4.18)
\]

Assuming data is transmitted only during active video, maximum crosstalk occurs when black video is transmitted. This assumption will be qualified in Section 6.3, and \([1 - v(t)]\) attains a value of \(\frac{3}{4} A_{sync}\), where \(A_{sync}\) is the peak video carrier amplitude.

The maximum amplitude of the video crosstalk will then be

\[
\frac{3A_{sync}}{4\sigma_{data}} \tan \phi. \quad (4.19)
\]

\(\sigma_{RF data}\) is approximately equal to \(\sigma_{data}\) and this further reduces to,

\[
\text{maximum crosstalk} = \frac{3}{4} 10^{-\frac{12}{30}} \tan \phi. \quad (4.20)
\]

This is plotted in Figure 4.12.

The crosstalk is always positive because \([1 - v(t)]\) is always positive. The lower portion of the eye would move closer to the threshold and the net effect
would be to decrease the eye height by twice the crosstalk amplitude. Figures 4.11 and 4.12 can be used together to show the effect of video crosstalk on the BER. For example if PRS1 is used the eye opening size is 1.1 volts. Under the following conditions:

\[
\begin{align*}
CNR_{video} & = 35 \text{dB} \quad (4.21) \\
I_Q &= -25 \text{dB} \quad (4.22) \\
B_{video} &= 6 \text{MHz} \quad (4.23) \\
B_{data} &= 1.5 \text{MHz} \quad (4.24) \\
R_D &= -0.7 \text{dB} \quad (4.25) \\
\phi &= 0 \quad (4.26)
\end{align*}
\]

\[SNR_{data} = 15.3 \text{dB}\] and from Figure 4.11 the BER would be \(6 \times 10^{-4}\). If \(\phi = \frac{1}{2}\) degree then from Figure 4.12 the crosstalk would have a peak amplitude of 0.21 volts and the eye opening would reduce to 0.68 volts. The BER would increase to \(2.5 \times 10^{-2}\).

From the analysis, it is clear that the data is quite vulnerable to video crosstalk. The low injection level makes it extremely difficult to recover the data properly. Any error phase error in the data demodulator would overwhelm the data with video. A possible solution to this problem would be to estimate the phase error, and then use this to subtract out a portion of the video from the data.
Chapter 5

Software Simulations

5.1 Introduction

Software simulations were performed using the Digital Video Facility (DVF) of the David Sarnoff Research Center. This section describes the DVF and software simulations in detail followed by results obtained from the simulation and how they compare with the theoretical framework presented in Chapter 3.

The entire system was simulated in software including video and data modulation and demodulation. Output in the form of video images was studied subjectively and with a Rhode-Schwartz video SNR meter. Audio baseband noise power densities and SNRs in individual audio channels were used to evaluate audio compatibility. Eye patterns of the demodulated data and BER vs. $CNR_{video}$ curves showed data ruggedness.

The DVF provides real time sampling and display of video images in addition to off line processing. Images are sampled at 4 times the color subcarrier frequency (14,318,180 samples/sec) by an 8 bit A/D converter. 100 IRE corresponds to 255 DVF units and -40 IRE (bottom of sync tip) to 0 DVF units. Samples were converted to floating point to eliminate roundoff and overflow errors during computation. Their floating point representations ranged from 0 to 1. Simulations were performed in digitized sync mode. In this mode the entire video signal (active video, horizontal and vertical blanking intervals) is sampled. Four field video sequences containing 954,545 samples were used.
The simulation is a combination of the block diagrams in Figures 2.5 and 4.1 and is shown in Figure 5.1. The implementation details of each subsection will be described next.

Baseband signals are modelled by real sequences and RF signals by complex sequences. RF signals are represented as,

\[ s(t) = i(t) \cos \omega t - q(t) \sin \omega t \]
\[ = \Re[z(t)e^{j\omega t}] \]  \hspace{1cm} (5.1) (5.2)

allowing the inphase and quadrature components to be manipulated independently. The modulator converts a baseband signal \( x(t) \) to RF through the following transformation,

\[ z(t) = x(t)e^{j\phi} \] \hspace{1cm} (5.3)
\[ z(t) = x(t)\cos \phi + jx(t)\sin \phi \] \hspace{1cm} (5.4)

where \( \phi \) is the phase of a cosine modulator.
Two demodulators were implemented, envelope and synchronous. Envelope detectors compute the magnitude of the RF signal,

$$x(t) = \sqrt{i^2(t) + q^2(t)}.$$  \hspace{1cm} (5.5)

Synchronous demodulators are specified by a phase angle and the baseband signal they return is

$$x(t) = i(t) \cos \phi + q(t) \sin \phi.$$ \hspace{1cm} (5.6)

Both baseband and RF filters operated on the same principle. Time sequences were converted to frequency samples by a 1024 point FFT and these were then multiplied with the filter's frequency response and then reconverted to time samples by an inverse FFT. Much care was taken to eliminate circular aliasing. When an M length sequence is convolved with one of length N, the result has length M+N-1. Filter impulse responses were kept below 512 samples and padded with zeros to length 1024. 513 time samples were padded to length 1024 and these filtered by the method previously described. The output has length 512 + 513 − 1 = 1024 so no circular aliasing occurred and was reconstructed using an overlap add algorithm.

5.2 Data Demodulator

The data demodulator was one of the most complicated pieces of the software simulation. It had to phase lock to the data by determining the optimum sampling time and then calculate bit error rates. It operated correctly for a variety of multi-level channel codes with different sampling thresholds. The data demodulator also plotted an eye pattern of the data to check for idealness and sensitivity to sampler timing errors.

The filters used were causal, and thus added phase delay to the signal as it propagated through the system. The data demodulator was provided with the data symbol rate and had to determine the phase of the data. For each sample in the symbol interval, a signal distribution histogram, such as seen in Figure 4.4 was produced. The sampling instant was chosen to be the time with largest average eye width. Although this is not the optimum test, it was sufficient to correctly synchronize the data.

The data BER was determined by the steps described in Section 4.2. The data signal was first scaled to have an average power of 1 and then the noise
variance was found using Equation 4.6. Then the BER was evaluated directly using Equation 4.2 and the signal distribution histogram found above. The BER was calculated for varying levels of channel noise and this result plotted.

5.3 Audio Demodulator

As shown in Section 3.3, audio interference is an additive noise caused by the phase of the picture carrier. Audio interference was measured as a SNR in each of the audio subchannels and several steps are needed in this calculation. First the noise power density in the audio baseband will be determined followed by SNR calculations for each subchannel.

Intercarrier audio demodulation was simulated. The picture carrier was first extracted by a second order filter with bandwidth 1 MHz centered around the picture carrier. The phase of this signal was extracted,

\[ P(t) = \tan^{-1} \left[ \frac{q(t)}{i(t)} \right] \]  

(5.7)

differentiated and scaled by \( \frac{1}{2\pi} \). The result is an implementation of Equation 3.31 with units Hz.

\[ n(t) = \frac{1}{2\pi} \frac{d}{dt} \tan^{-1} \left[ \frac{q(t)}{i(t)} \right] \]  

(5.8)

and represents the total noise. To find the SNR in each subchannel, the noise in each subchannel was integrated in the frequency domain.

The bandwidths of the audio channels are on the order of KHz and the sampling rate on the order of MHz. To get a reasonable amount of detail in the frequency domain a very large FFT is necessary. \( 2^{18} \) samples of \( n(t) \) were transformed and each frequency sample represented the noise in a \( B_o = 54.6 \) Hz bandwidth. To obtain the noise power density, take the magnitude squared of the frequency samples (the power) and divide by \( B_o \). The quantity that results has units \( \frac{Hz^2}{Hz} \) and will be referred to as \( N(f) \).

5.3.1 L+R Subchannel

The L+R subchannel extends from 0 to 15 KHz and in order to reduce noise, the signal is pre-emphasized at the transmitter and de-emphasized at the
receiver. The de-emphasis function is a simple RC circuit with \( \tau = 75 \mu \text{sec} \) and its power transfer function is

\[
d(f) = \left| \frac{1}{1 + 2\pi f \tau} \right|^2.
\]  
(5.9)

Then the total noise in the L+R subchannel is

\[
N_{L+R} = \int_{0}^{15\text{KHz}} N(f)d(f)df
\]  
(5.10)

and has units Hz\(^2\). This noise is compared with the signal power in the L+R subchannel

\[
S_{L+R} = \frac{(25 \text{ KHz})^2}{2}
\]  
(5.11)

to obtain

\[
SNR_{L+R} = 10 \log \left[ \frac{S_{L+R}}{N_{L+R}} \right].
\]  
(5.12)

25 KHz represents the peak deviation of the FM carrier due to the L+R subchannel.

### 5.3.2 L-R Subchannel

The L-R subchannel is similar to the L+R and the SNR calculation is basically the same. The L-R subchannel extends \( \pm 15 \text{ KHz} \) around 31.5 KHz and uses the same de-emphasis function centered around 31.5 KHz.

\[
N_{L-R} = \int_{-15\text{KHz}}^{15\text{KHz}} N(f + 31.5 \text{ KHz})d(f)df.
\]  
(5.13)

The peak deviation is 50 KHz instead of 25 KHz and

\[
S_{L-R} = \frac{(50 \text{ KHz})^2}{2}
\]  
(5.14)

and

\[
SNR_{L-R} = 10 \log \left[ \frac{S_{L-R}}{N_{L-R}} \right].
\]  
(5.15)

The effects of companding in the L-R and SAP channels were ignored because they were very difficult to simulate due to the long time constants involved in their operation.
5.3.3 SAP Subchannel

The second audio program (SAP) subchannel differs from the other two in that it is FM modulated, and calculating its SNR requires considerably more work. First the relationship between \( SNR_{SAP} \) and \( CNR_{SAP} \) will be established. When flat white noise is FM demodulated, the noise it creates has a quadratic spectrum.

\[
n(f) = \frac{r f^2}{C} \tag{5.16}
\]

where \( r \) equals the noise power density at the input in \( \frac{Hz^2}{Hz} \) and \( C \) equals the carrier power in \( Hz^2 \).

This equation applies to the composite audio baseband; noise increases as \( f^2 \) and this is why pre/de-emphasis is used. Noise in the composite audio baseband in the location of the SAP subchannel will be modelled as begin flat and Equation 5.16 applied to this situation. If the average noise power density is \( \bar{r} \), then the noise in the demodulated SAP subchannel is

\[
n(f) = \frac{\bar{r} f^2}{C} \tag{5.17}
\]

as before. Multiplying top and bottom by the IF bandwidth of the SAP subchannel and rearranging terms leaves

\[
n(f) = \frac{f^2}{B} \cdot \frac{C}{B\bar{r}} = \frac{f^2}{B} / CNR_{SAP}. \tag{5.18}
\]

Integrating this over the SAP audio output bandwidth \( W \) gives the total noise in the SAP,

\[
N_{SAP} = \frac{1}{3} \frac{W^3}{B} / CNR_{SAP} \tag{5.19}
\]

with units \( \frac{Hz^2}{Hz} \). The SAP signal power is \( \frac{D^2}{2} \) where \( D \) is the peak deviation. The final SNR in dB is

\[
SNR_{SAP} = 10 \log \left[ \frac{3 D^2 B}{2 W^3 CNR_{SAP}} \right] \tag{5.20}
\]

and using the values \( D = 10 \text{ KHz}, B = 30 \text{ KHz} \) and \( W = 15 \text{ KHz} \) one obtains the relation

\[
SNR_{SAP} = CNR_{SAP} + 1.2 \text{ dB}. \tag{5.21}
\]

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<table>
<thead>
<tr>
<th>Subchannel</th>
<th>Simulation SNR (dB)</th>
<th>Expected SNR (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>L+R</td>
<td>64.9</td>
<td>65.4</td>
</tr>
<tr>
<td>L-R</td>
<td>50.9</td>
<td>50.6</td>
</tr>
<tr>
<td>SAP</td>
<td>25.5</td>
<td>26.8</td>
</tr>
</tbody>
</table>

Table 5.1: Audio SNRs for 0 IRE black field.

$CNR_{SAP}$ is found by integrating the noise in the audio baseband over the SAP subchannel

$$N_{SAP\,\text{carrier}} = \int_{63.75\,KHz}^{93.75\,KHz} N(f)\,df$$  \hspace{1cm} (5.22)

and then comparing this with the SAP carrier power $\frac{(15\,KHz)^2}{2}$.

5.3.4 Audio SNR Accuracy Check

The above audio SNR calculations were checked with the known performance of the BTSC audio system when white noise is added in the transmission channel. An RF video signal was created and 30 dB $CNR_{video}$ white noise added. This was demodulated by the simulation and audio SNRs calculated. Table 5.1 lists the audio SNRs determined by the simulation along with their actual values.\(^1\) The values shown are for a 0 IRE black field and a -10 dB visual to aural carrier peak power ratio. As can be seen, the simulations accurately predict SNRs.

In addition to the audio SNRs, $n(f)$ vs. $f$ was plotted in the audio baseband interval. Overlaid upon this was the theoretical noise power density caused by 30 dB $CNR_{video}$ noise. This curve is Equation 5.16 evaluated when the audio carrier to noise ratio is

$$\frac{C}{r} = CNR_{audio}$$  \hspace{1cm} (5.23)

$$= CNR_{video} + 10 \log \left[ \frac{\text{audio carrier power}}{\text{video carrier power}} \right]$$  \hspace{1cm} (5.24)

10 \log \left[ \frac{B_{\text{video}}}{B_{\text{audio}}} \right]. \tag{5.25}

When \( CNR_{\text{video}} = 30 \text{ dB} \), \( B_{\text{video}} = 6 \text{ MHz} \), \( B_{\text{audio}} = 400 \text{ KHz} \) and the audio carrier power is 1/10 that of the video, \( CNR_{\text{audio}} = 31.8 \text{ dB} \).

## 5.4 Simulation Results

The complete simulation was performed using the 3 channel codes described in Section 4.3 for a variety of injection levels and using black and white fields. The data rate for all channel codes is 954,545 bits/sec. Ideal Nyquist, inverse Nyquist and VSB filters were used. The simulations examined the performance of envelope detectors and not synchronous detectors because interference in the latter case depends on the demodulation phase error which varies among receivers.

Flat white (100 IRE), grey (50 IRE) and black (10 IRE) fields were digitized and served as source video sequences. Flat fields were chosen because the Rhode Schwartz video SNR meter required them. Video SNR figures are for a 4.2 MHz unweighted bandwidth. The simulation results are summarized in Table 5.2.

The noise floor of the simulation is the optimal performance in the absence of any quadrature modulation. This best achievable video SNR was 59.6 dB for the white field and 75.4 dB for the black field. The main source of residual noise was determined to be from the video test pattern generator and not from the simulation.

The quadrature modulation appeared as low frequency white noise in the video. When the injection level was -20 dB, the noise was easily visible in both black and white fields. At -25 dB it could still be seen in the white field but was almost invisible in the black field. At -30 dB it was impossible to see in the black field and could only be seen in the white field by adjusting the display contrast. This level of interference was deemed acceptable.

From a SNR point of view, the white field is most sensitive to quadrature modulation. Subjectively this is not true because nonlinearities in the picture tube and in the human eye may at high amplitudes masking the noise. A video SNR of 42 dB is usually considered excellent when the noise is flat in a 4.2 MHz bandwidth. The noise in this case was bandlimited to 750 KHz and the high video SNRs may be misleading.

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### 100 IRE White Field

<table>
<thead>
<tr>
<th>Channel Code</th>
<th>Injection Level (dB)</th>
<th>BER $w/CNR_{\text{video}}$</th>
<th>Video SNR (dB)</th>
<th>Audio SNR (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>30 dB</td>
<td>35 dB</td>
<td>L+R</td>
<td>L-R</td>
</tr>
<tr>
<td>NRZ</td>
<td>-25</td>
<td>$3.07 \times 10^{-4}$</td>
<td>43.9</td>
<td>45.2</td>
</tr>
<tr>
<td></td>
<td>-30</td>
<td>$2.54 \times 10^{-2}$</td>
<td>52.2</td>
<td>50.0</td>
</tr>
<tr>
<td>PRS1</td>
<td>-20</td>
<td>$1.37 \times 10^{-4}$</td>
<td>31.8</td>
<td>53.1</td>
</tr>
<tr>
<td></td>
<td>-25</td>
<td>$1.78 \times 10^{-2}$</td>
<td>41.0</td>
<td>57.8</td>
</tr>
<tr>
<td></td>
<td>-30</td>
<td>$1.44 \times 10^{-1}$</td>
<td>50.8</td>
<td>62.7</td>
</tr>
<tr>
<td>PRS2</td>
<td>-20</td>
<td>$3.52 \times 10^{-3}$</td>
<td>30.4</td>
<td>58.8</td>
</tr>
<tr>
<td></td>
<td>-25</td>
<td>$9.16 \times 10^{-2}$</td>
<td>39.7</td>
<td>62.8</td>
</tr>
<tr>
<td></td>
<td>-30</td>
<td>$3.27 \times 10^{-1}$</td>
<td>49.5</td>
<td>67.5</td>
</tr>
</tbody>
</table>

### 0 IRE Black Field

<table>
<thead>
<tr>
<th>Channel Code</th>
<th>Injection Level (dB)</th>
<th>BER $w/CNR_{\text{video}}$</th>
<th>Video SNR (dB)</th>
<th>Audio SNR (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>30 dB</td>
<td>35 dB</td>
<td>L+R</td>
<td>L-R</td>
</tr>
<tr>
<td>NRZ</td>
<td>-25</td>
<td>$3.07 \times 10^{-8}$</td>
<td>63.0</td>
<td>56.3</td>
</tr>
<tr>
<td></td>
<td>-30</td>
<td>$2.54 \times 10^{-4}$</td>
<td>63.0</td>
<td>63.3</td>
</tr>
<tr>
<td>PRS1</td>
<td>-20</td>
<td>$1.37 \times 10^{-8}$</td>
<td>44.2</td>
<td>63.5</td>
</tr>
<tr>
<td></td>
<td>-25</td>
<td>$1.78 \times 10^{-4}$</td>
<td>55.0</td>
<td>68.5</td>
</tr>
<tr>
<td></td>
<td>-30</td>
<td>$1.44 \times 10^{-1}$</td>
<td>61.7</td>
<td>73.5</td>
</tr>
<tr>
<td>PRS2</td>
<td>-20</td>
<td>$3.52 \times 10^{-7}$</td>
<td>42.8</td>
<td>73.6</td>
</tr>
<tr>
<td></td>
<td>-25</td>
<td>$9.16 \times 10^{-3}$</td>
<td>52.9</td>
<td>78.6</td>
</tr>
<tr>
<td></td>
<td>-30</td>
<td>$3.27 \times 10^{-2}$</td>
<td>59.0</td>
<td>83.6</td>
</tr>
</tbody>
</table>

Table 5.2: Results of software simulations.
Figure 5.2: Audio baseband distortion for NRZ channel code. 100 IRE white field, -30 dB injection level.

The results of the simulations agree with the theoretical framework developed in Chapter 3. As predicted by Equation 3.11 the video interference increases approximately 10 dB for every 5 dB increase in injection level. This relation holds for both black and which fields. Also, video interference is worst for white fields with video SNRs being 10-12 dB lower than for black fields.

Audio results also agree with the theoretical results of Section 3.3. As predicted, low frequency quadrature modulation is transferred to the audio baseband and channel codes with less low frequency energy produce less audio interference. The audio baseband distortion graphs for the 3 channel codes are shown in Figures 5.2 - 5.4. As expected from their data spectrums, NRZ causes the greatest audio degradation followed by PRS1 and then PRS2.

Audio noise is also a linear function of the quadrature amplitude as pre-
Figure 5.3: Audio baseband distortion for PRS1 channel code. 100 IRE white field, -30 dB injection level.
Figure 5.4: Audio baseband distortion for PRS2 channel code. 100 IRE white field, -30 dB injection level.
<table>
<thead>
<tr>
<th>Subchannel</th>
<th>Actual SNR (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>L+R</td>
<td>56.7</td>
</tr>
<tr>
<td>L-R</td>
<td>41.9</td>
</tr>
<tr>
<td>SAP</td>
<td>18.1</td>
</tr>
</tbody>
</table>

Table 5.3: Audio SNRs for 100 IRE white field.

dicted by Equation 3.33. A 5 dB increase in the injection level causes audio SNRs to drop by 5 dB. Audio is most impaired during white fields. As in the case of video interference, audio interference is 10-12 dB greater for white over black fields.

Audio SNRs for black fields should be compared with the right hand column of Table 5.1 and for white fields with Table 5.3. These list SNRs when the interference is caused by 30 dB $CN_{R_{video}}$ channel noise. From an audio compatibility point of view, NRZ at all injection levels tested as well as PRS1 at -20 dB are not suitable. PRS2 has acceptable levels of interference at all injection levels tested.

The BERs presented in Table 5.2 are for random errors caused by white channel noise. Video crosstalk and sampling errors would further degrade these values. The multilevel codes are most difficult to recover due to their small eye heights. When the data is invisible (-30 dB injection), none of the channel codes provide suitable (BER < $10^{-3}$) performance at 30 dB $CN_{R_{video}}$ except NRZ but it fails audio compatibility. Operation of this system would then have to be limited to areas where $CN_{R_{video}} \geq 35$ dB.

5.5 Conclusion

Software simulations supported the theoretical results of Chapter 3 and provided insight into the compatibility of quadrature modulation with NTSC. They reduce the number of candidate modulation schemes and provide bounds on the injection levels. The NRZ channel code should be excluded on grounds of audio incompatibility. The area where data can successfully be recovered is limited to having $CN_{R_{video}} \geq 35$ dB and some tradeoffs could be made between injection levels and service area range. Also PRS2 is ex-
tremely difficult to recover reliably and it may have to be removed from the list of possible channel codes. The only viable channel code left is PRS1 and this will be studied further in the following chapter on hardware tests.
Chapter 6

Hardware Compatibility Tests

The software simulations described in the last chapter used ideal models of television receivers and therefore had to agree with the theoretical framework. The tests described next, repeated the software experiments but with real receivers and are a true test of compatibility. The experiments were not limited to envelope detectors, but were carried out using all three types of video demodulators. Also, the data could be added to off air signals for subjective testing during motion sequences.

6.1 Hardware Setup

The hardware test bed provides an environment for evaluating modifications to NTSC and is well suited for this application. It contains a complete channel 3 transmitter operating at 61.25 MHz, a channel simulator and several receivers. The main transmitter was augmented with a quadrature modulator, NQM-1, and Norm Hurst used this in the sine wave tests mentioned in Section 2.2. His setup was used with two minor modifications: the input to the NQM-1 was data generated by the DVF, and an inverse Nyquist filter was added. The hardware setup used for these experiments can be seen in Figure 6.1.

When the quadrature signal is bandlimited to ±750 KHz, an inverse Nyquist filter can be constructed by adding a fraction of the derivative of the signal to the inphase component. The data and its derivative were generated by the DVF and sent down to the test bed. Two attenuators adjusted
Figure 6.1: Hardware test bed setup.
the inverse Nyquist slope and set the injection level.

The quadrature data component fed the NQM-1 and the inphase component was added to the baseband video and this composite signal modulated the main carrier. The main modulator had a 6.3 $\mu$s delay relative to the NQM-1 and this was compensated for on the DVF by delaying the quadrature data component 90 samples.

Four television receivers were evaluated: *synchronous* – Tektronixs 1450, Sony 27TX20; *quasi-synchronous* – RCA ColorTrak 2000; *envelope* Tektronixs 1450 in envelope detection mode. Video interference was analyzed subjectively and with the Rhode-Schwartz video SNR meter.

### 6.2 System Calibration

Calibrating the test bed was a tedious process due to the large number of adjustments it had. Three important quantities had to be carefully set for accurate results:

- *quadrature phase* – phase of the NQM-1 relative to the main modulator.
- *quadrature injection level* – as before, the ratio of RF data power to the video carrier power at peak of sync.
- *derivative scaling constant* – the difference in amplitude between the inphase and quadrature data components. Used in constructing the inverse Nyquist slope.

The quadrature phase was set by adjusting trombones connected to the quadrature modulator. A meter on the NQM-1 displayed quadrature phase and the trombones were adjusted until this read zero.

Next the derivative scaling constant was set using a data sequence consisting of 750 KHz sync pulses generated on the DVF. This repetitive sequence and its derivative were delayed as the data would be, and fed the modulators. With video modulation removed, the RF output of the transmitter was viewed on a spectrum analyzer. The attenuators were adjusted until the correct inverse Nyquist slope appeared. It was found that the quadrature data component had to be attenuated 8.9 dB relative to the inphase component to get a “clean” inverse Nyquist slope. This slope can be seen in Figure 6.2 and the scaling constant was used for all tests. The three channel codes
after passing through the inverse Nyquist filter are shown in Figures 6.3 - 6.5. These spectral plots certify that the inverse Nyquist slope was correctly adjusted during the tests. Any errors in phase or relative amplitudes would manifest themselves here.

The design of the NQM-1 is such that 1 volt peak-to-peak input signal modulates the quadrature carrier with amplitude equal to the video carrier at peak of sync. By measuring the rms NQM-1 input voltage, the RF data power could be determined and from this, the quadrature injection level. Remembering that the RF data power contains both quadrature and inphase components,

\[ I_Q = 20 \log \left[ \frac{V_{rms}}{0.5} \right] - R_D \]  \hspace{1cm} (6.1)
Figure 6.3: NRZ channel code after inverse Nyquist slope.
Figure 6.4: Three level partial response channel code after inverse Nyquist slope.
Figure 6.5: Five level partial response channel code after inverse Nyquist slope.
<table>
<thead>
<tr>
<th>Video (flat field)</th>
<th>Receiver</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Tek. Sync.</td>
</tr>
<tr>
<td>100 IRE</td>
<td>58.0</td>
</tr>
<tr>
<td>50 IRE</td>
<td>57.6</td>
</tr>
<tr>
<td>10 IRE</td>
<td>57.5</td>
</tr>
</tbody>
</table>

Table 6.1: Noise floor of hardware system in dB.

or alternatively

\[ V_{rms} = 2 \left[ 10^{\frac{I_q + n_d}{20}} \right]. \] (6.2)

The overall noise floor of the system was measured for clean flat fields. The results for different fields and video detectors is shown in Table 6.1. These values will later be used in the evaluation of the results.

6.3 Hardware Results

The tests were similar to the ones performed by software simulation. Four field sequences of data were used and video SNRs measured for the three channel codes, varying injection levels, and multiple video sources. In addition, off air signals were subjectively tested.

The video SNRs were significantly lower (5-7 dB) than those predicted by the simulation and the interference had a different character. The noise still appeared random, but was correlated along individual scan lines; often an entire line was displayed either too dark or too bright. It was determined that the data located in the horizontal blanking interval would hinder the measurement of clamping levels (front porch, color burst, etc.) by the receivers and subsequently complete lines were incorrectly displayed.

To alleviate this problem, data was removed during the horizontal blanking interval and by this the data rate was reduced by 16%. This gating is similar to that used by Matsushita in their quadrature transmission system. The tests were rerun with this gated data and the results now agreed better with the simulations. All channel codes tested gave similar video SNRs and only differed by approximately 1-1.5 dB. The results are shown in Table 6.2

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<table>
<thead>
<tr>
<th>Video Source</th>
<th>Injection Level (dB)</th>
<th>Video SNR in dB for receiver Tek. Sync.</th>
<th>Tek. Env.</th>
<th>RCA</th>
<th>Sony</th>
<th>Simulation Video SNR (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>100 IRE</td>
<td>-20</td>
<td>39.5</td>
<td>30.8</td>
<td>27.4</td>
<td>35.3</td>
<td>31.8</td>
</tr>
<tr>
<td></td>
<td>-25</td>
<td>44.2</td>
<td>39.3</td>
<td>33.1</td>
<td>38.0</td>
<td>41.0</td>
</tr>
<tr>
<td></td>
<td>-30</td>
<td>48.7</td>
<td>46.2</td>
<td>38.0</td>
<td>41.0</td>
<td>50.8</td>
</tr>
<tr>
<td></td>
<td>-35</td>
<td>49.0</td>
<td>48.2</td>
<td>42.0</td>
<td>46.0</td>
<td></td>
</tr>
<tr>
<td>50 IRE</td>
<td>-20</td>
<td>47.0</td>
<td>40.9</td>
<td>32.3</td>
<td>40.0</td>
<td></td>
</tr>
<tr>
<td></td>
<td>-25</td>
<td>50.1</td>
<td>47.2</td>
<td>37.8</td>
<td>43.3</td>
<td></td>
</tr>
<tr>
<td></td>
<td>-30</td>
<td>53.5</td>
<td>52.8</td>
<td>41.9</td>
<td>45.6</td>
<td></td>
</tr>
<tr>
<td></td>
<td>-35</td>
<td>51.0</td>
<td>51.2</td>
<td>44.4</td>
<td>47.2</td>
<td></td>
</tr>
<tr>
<td>10 IRE</td>
<td>-20</td>
<td>50.0</td>
<td>45.2</td>
<td>33.6</td>
<td>42.1</td>
<td>44.2</td>
</tr>
<tr>
<td></td>
<td>-25</td>
<td>52.7</td>
<td>50.3</td>
<td>39.4</td>
<td>44.8</td>
<td>55.0</td>
</tr>
<tr>
<td></td>
<td>-30</td>
<td>54.5</td>
<td>54.2</td>
<td>43.1</td>
<td>46.0</td>
<td>61.7</td>
</tr>
<tr>
<td></td>
<td>-35</td>
<td>51.7</td>
<td>52.2</td>
<td>45.5</td>
<td>48.3</td>
<td></td>
</tr>
</tbody>
</table>

Table 6.2: Video SNRs with gated data. Results for PRS1 code.

along with values from the simulations. Overall, gating the data improved the video SNR by approximately 2 dB and decreased the visibility of the data by removing line correlation.

Although the video SNRs were closer, subjectively the interference was more visible in the hardware tests than in the simulations. The quadrature modulation was visible on all receivers even when the injection level was -35 dB. Visibility of the quadrature signal could also have been exaggerated due to the short data sequences used. The 4 fields of data caused stationary interference, which is more noticeable than random noise.

The noise floor of the hardware system was also much higher than in the simulations. Many of the values obtained by the simulations could not be expected in the hardware system. However, picture quality without data was excellent and so the lower noise floor does not account for the discrepancies in video SNRs. The system was recalibrated several times and the experiments repeated, but the results obtained were unchanged.

Examination of the data shows that as expected, the Tektronix synchronous receiver was least affected by quadrature modulation. This is a
professional quality receiver and so it is not surprising that it even outperformed the Sony synchronous receiver while in envelope detection mode. The envelope detector performed reasonably close to the simulation, as measured by its video SNR, but the data was visible even for low injection levels. The quasi-synchronous receiver consistently performed worst, and this is disturbing since the majority of American receivers use this type of detector. This result by itself is enough to rule out compatible operation of this data channel. The interference into the quasi-synchronous receiver can most likely be traced to phase modulation of the picture carrier. The demodulator had a phase error proportional to the picture carrier phase, and severe cross talk resulted.

The discrepancies between hardware experiments and simulations could be caused by several factors. First, correct compatible operation depends on the exact shape of the inverse Nyquist filter. Any deviations and data will no longer be completely in quadrature. Normally, the exact cutoff frequency of the Nyquist slope is not critical; as long as it has a symmetric shape it functions properly. Some of the receivers tested might not have standard Nyquist filters. It would be worthwhile to test several receiver IF sections and determine their exact response. From this data an inverse filter could be determined which minimizes overall interference.

Errors could also have been caused by nonlinear phase distortions in any of the RF receiver filters. Nyquist slope alignment was checked after the transmitter, so any phase irregularities must have been introduced in the receivers. This nonlinear could be compensated for at the transmitter if all receivers were known to have similar phase characteristics. Another test was performed to pinpoint the source of this error.

6.3.1 Sine Wave Injection Test

This test was specifically aimed at the envelope detector. The data was replaced by a 150 KHz sine wave generated on the DVF. It was filtered by the inverse Nyquist filter and gated during the horizontal and vertical blanking intervals and appeared only during active video. If the Nyquist receiver filter was functioning properly, then the signal would be completely in quadrature and appear as a 300 KHz interfering sine wave. This however was not the case, the video contained both 150 KHz and 300 KHz components. Therefore the receiver Nyquist filter must have a phase or magnitude error or both.
<table>
<thead>
<tr>
<th>Injection Level (dB)</th>
<th>Stereo Separation (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>-∞</td>
<td>43</td>
</tr>
<tr>
<td>-35</td>
<td>25</td>
</tr>
<tr>
<td>-30</td>
<td>12</td>
</tr>
</tbody>
</table>

Table 6.3: Effects of quadrature modulation on stereo separation.

### 6.3.2 Stereo Separation Test

In addition to the video tests performed, audio was analyzed subjectively. Interference in the L and R subchannels was demodulated and listened to. The distortion was objectionable for the NRZ code at all injection levels tested. The partial response codes were acceptable if the injective level was below -25 dB. The degradation of stereo separation was also measured.

Stereo separation is defined as the leakage between stereo channels and is one measure of reception quality. The measurement was performed by placing a 1 KHz test tone in the left channel at the transmitter. The rms voltages in the received L and R channels were measured.

\[
\text{Stereo Separation} = 20 \log \left[ \frac{L_{\text{rms}}}{R_{\text{rms}}} \right].
\] (6.3)

The test was performed with no quadrature modulation and with PRS1 modulating the quadrature carrier. Results for varying injection levels are shown in Table 6.3.

As is shown, even a low injection level (-35 dB) decreases stereo reception by 18 dB. Loss of separation is not as bothersome as additive noise, but some consumers would probably be annoyed by this. Stereo reception is a feature on new receivers, something people have had to pay extra for, and quadrature modulation would make their receivers sound monaural.
Chapter 7

Conclusion and Recommendations

The results demonstrate that the three channel codes tested are not suitable for NTSC compatible transmission. Several conflicting design issues could not be balanced in order to produce a working system. The main trade off dealt with the injection level of the data. On the one hand, a high injection level is needed for robust data transmission in a broad service area. On the other hand, with a high injection level the data is visible in the received television picture and incompatible. An injection level could not be chosen to meet both of these design requirements.

The compatible operation of this system depends upon matching the receiver Nyquist filter with an inverse filter. Theoretically the system should function with minimal interference. However hardware tests demonstrate that variation exists in the RF section of commercial television receivers, and that an inverse Nyquist filter compatible with all receivers does not exist.

Hardware tests show that of all receivers, quasi-synchronous are most affected by quadrature modulation. This is disturbing since the bulk of commercial receivers use this type of demodulator. Quasi-synchronous receivers should be studied more closely, and the precise source of interference identified. Using this information, a channel code with lower video interference might be designed.

In addition to video interference, the data is fragile due to its low injection level. Channel noise alone creates BERs on the order of $10^{-2}$ and any imperfections in the data demodulator would increase this figure. The
data is very sensitive to phase errors; an error of 1 degree is usually sufficient to make recovery impossible. In the BER calculations timing jitter in the data demodulator and multipath were ignored, and these would further degrade data performance. Due to the low injection levels, multilevel codes are impossible to recover reliably. One possibility would be to use a two level code with most of its energy concentrated at high frequencies. By using an extremely low injection level (< -35 dB), it might be compatible and still recoverable.

In conclusion, three channel codes were examined, and none were able to meet all design specifications. Although these results do not rule out a compatible system, they do demonstrate the inherent difficulties associated with applying quadrature modulation to compatible television transmission.
Bibliography


