Performance of a 256-QAM Demodulator/Equalizer in a Cable Environment

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Abstract

The premise of this paper is to show that 256-QAM is a viable modulation type for the cable television environment. Performance results for various cable system impairments such as AM hum, FM phase noise, residual FM, SNR, and microreflections are presented for an actual prototype 256-QAM Equalizer/ Demodulator that verify this premise. The prototype demodulator has also undergone extensive testing at the Advanced Television Test Center (ATTC) in support of the cable portion of the HDTV testing (January 1994). All prototype testing results conclude that, with a well conceived architecture and design, 256-QAM is an eminently practical format for the transmission and reception of high-data-rate cable television signals. QAM and VSB modulations are compared. There are advantages in the use of 256-QAM over 16-VSB in areas involving carrier and phase tracking and blind equalization which are analyzed here and also supported by results obtained in carrier offset, residual FM, phase noise, and channel change acquisition time testing [8]. Cable microreflection environments are modeled and equalizer length trades are presented that indicate that 16-32 taps (possibly 64 taps in some cases) are generally required to overcome the effects of cable microreflections.

Introduction

The advanced methods of digital compression have brought new visions of interactive television which will soon be realized in practice. Although some cannot fathom the use of hundreds of new television channels, the broadcaster's vision is that as video-on-demand grows in popularity due to sports events, home shopping, etc., the search for more capacity will continue. This paper will present results of a QAM system, e.g., 256-QAM, that will increase the capacity by 33% over 64-QAM. Section 1.0 of this paper presents an overview of what is theoretically possible and what is or can be practical. Section 2.0 compares QAM with VSB which has recently been recommended for HDTV broadcast application. Section 3.0 presents results of extensive simulations and laboratory testing of a 256-QAM prototype modem for digital television over cable. Finally, results are summarized in Section 4.0.

1.0 SIGNAL DESIGN TRADES

1.1 Introduction

Over the years, many of the advances in modem design have been brought on by the desire to increase capacity over telephone lines. Many of the characteristics of the telephone line are similar to cable channels (strictly band-limited channels with amplitude and delay distortion) and thus, recent techniques that have increased the capacity of the phone lines from 9.6 to 14.4 kbps up to the mid twenties of kbps can be adopted. There are however, some significant practical differences as well. One of the obvious differences is that phone lines are point-to-point and therefore the channel between the transmitter and receiver can be sensed and adjusted prior to transmission. This is the basis of attaining channel capacity via tailoring the transmission spectrum to be highest where there is highest SNR, etc., and is made practical via Tomlinson and Harashima precoding techniques (see Reference [1] for further details) and multi-carrier techniques ([2]). Unfortunately, in the cable channel, the situation is point-to-multipoint and thus the channel can be significantly different from consumer to consumer depending on the neighborhood cable layout and the consumer's own premises cabling. Even in the situation of full video-ondemand, the subscriber would share a 6 MHz digital carrier with 5 to 10 other subscribers and thus the 6 MHz carrier could not be "tailored" to any particular household. Thus, as has been shown in a number of prior NCTA papers [3], [10], the requirement for adaptive equalization for each consumer is necessary.

The cable signal needs to achieve the most capacity for a given (reasonable) cost. In a sense, this can be considered a "cost-limited" channel situation. If operator revenue can be considered to increase linearly with number of channels transmitted (or data rate per 6 MHz carrier), it can be stated as a general rule that each increase of 2 bits/Hz requires 6 dB more SNR and thus one more bit of precision in the demodulator and equalizer. Thus, for example, a multiplier may have to be increased from 9 x 9 bits to 10 x 10 bits or a rough increase of 100/81 = 24% increase in die area/complexity. This is roughly linear with the 33% increase in capacity stated earlier for 256-QAM versus 64-QAM, and thus the semiconductor costs will be roughly proportional to the increased revenue from the operator.

Thus, what limits us now? Of course the complexity of the set-top box is not solely governed by the demodulator subsystem but also by the tuner, filters, quality of splitters, fiber, etc. (see Figure 1-1) which would tend to degrade the higher capacity signal and would also have to be included in the cost equation if they would have to be modified. As a result, it would be cost-effective if the signal design could work with minimal modifications to the entire cable infrastructure. It is for this reason that we have added a system cost-limited constraint on the signal design.

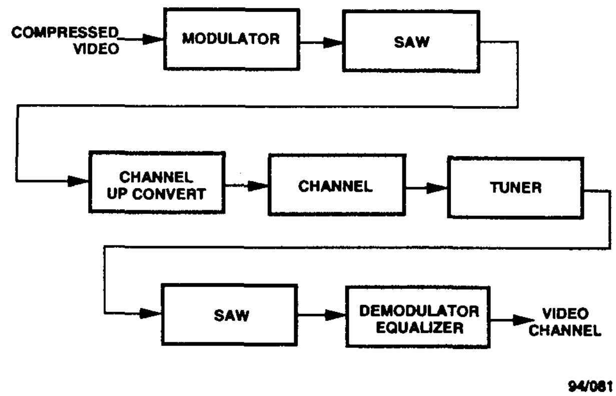

Figure 1-1. System Overview

1.2 BER versus SNR Bounds

Considering the theoretical bounds, it has been shown that for a strictly band-limited, high SNR, Gaussian noise/interference channel there is a 9 dB difference between the BER curves of an ideal Shannon system and uncoded MxM QAM [1, 4]. This 9 dB bound includes the knowledge of the subscriber channel and some shaping gain [4]. Practical coding schemes (e.g., Trellis, pre-coding, and shaping) have achieved as much as 7 dB gain [4]. Thus, for example, for 64-QAM, the required SNR (measured in the baud rate bandwidth) for 10^{-6} BER (using proper signal design techniques) can be as low as 19.6 dB and for 256-QAM, 25.6 dB, and 512-QAM, 28.6 dB. Thus, in an ideal band-limited cable channel with only amplitude/delay channel distortions, the above BER vs. SNR bounds hold.

The available SNR in reasonable worst-case cable environments have been presented in [5] as 30 dB SNR and in [6] as 32 dB SNR. Accounting for the fact of the inevitable losses in SNR due to consumer-caused cable wiring and losses due to other cable impairments (e.g., splits without pre-amplifications, ingress, etc.) a safe number might be 26 to 27 dB SNR with 3-6 dB of margin. Thus, in this environment the 256-QAM system appears to be a reasonable capacity-approaching compromise for the transmitted signal due to its requirement for only 25.6 dB SNR.

It should be noted that if constellation expanding trellis coding is applied to the signal, the actual signal may be a 512-QAM although the bps/Hz is still that of 256-QAM. As an example, a practical realization of a coded 256-QAM signal would be a 256-QAM signal with Reed-Solomon encoding which requires 27-28 dB SNR for 10^{-6} BER and results in an additional baud rate overhead of approximately 10%. (Note: Trellis coding would not result in any additional overhead.)

Finally, the above results depend on the demodulator circuits being properly designed to substantially remove (without additional significant SNR degradation) the effects of upconverter, tuner, etc., phase noise, AM hum, etc. This is very important since many cable environments are <u>not</u> SNR-limited but rather phase-noise limited.

1.3 Baud Rate Trades

It is interesting now to determine the maximum baud rate that can be obtained with the coded 256-QAM system with a given level of complexity of the demodulator. The 6 MHz channel is essentially band-limited by the combined effects of the transmit and receiver SAW filters. The maximum baud rate as a function of demodulator complexity will now be addressed with the constraint of the low-cost SAW filters.

Since the allocated channel bandwidth for each digital TV signal is 6 MHz, it is theoretically possible to operate a system at a symbol rate approaching this value. However, practical limitations arise due to:

1. the achievable frequency selectivity of input band-select filters,

2. the achievable SNR over the channel, and

3. the amount of digital processing and coding which can practically — from both a technical and economic standpoint — be applied to achieve a desirable level of performance (i.e., BER).

This technical and economic trade-off is illustrated in the family of curves shown in Figure 1-2. The details of these curves may change slightly depending upon the initial assumptions, but the character of the trade-off remains. Shown in this figure are curves of coded demodulation performance loss (at a BER of 10^{-6}) versus symbol rate for 256-QAM modulation. In these simulations, a typical SAW filter model has been used which exhibits 40 dB of rejection outside of the 6 MHz channel bandwidth. This channel-select filter model, derived from measurements on an actual device used in this application, is shown in Figure 1-3. Also assumed in these curves is the use of a particular Reed-Solomon code (t = 10) which can achieve roughly 4.5 dB of coding gain at a BER of 10^{-6}. (The vertical scale shows the loss relative to <u>uncoded</u> performance.) A fractional, T/2, equalizer has been assumed, and the curves show the performance loss for different length equalizer structures.

These curves illustrate that, as the symbol rate is forced higher for a given channel, in order to maintain a given level of performance (say 10^{-6} BER), a higher channel SNR must be achieved <u>and</u> a more powerful equalizer must be used. Both requirements lead to higher system costs.

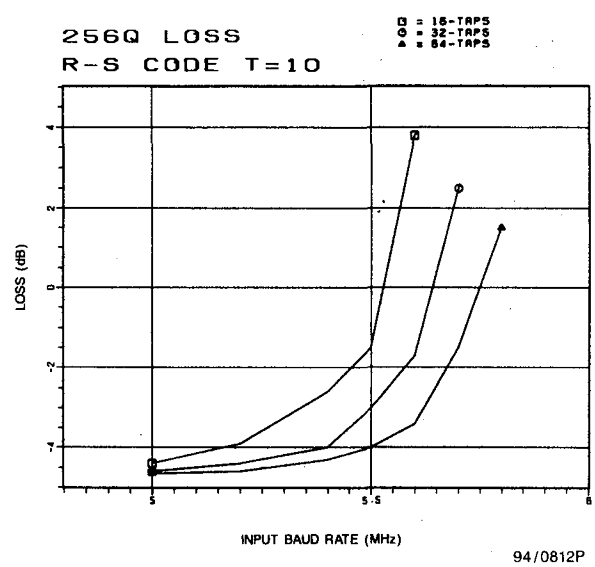

Figure 1-2. 256-QAM Loss versus Baud Rate for 6 MHz Channel and Reed-Solomon (t=10) Coding

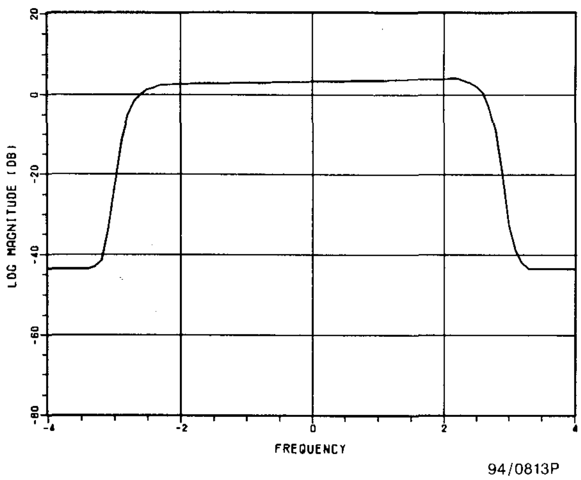

Figure 1-3. Model of Combined Transmitter/ Receiver SAW Channel-Select Filtering

2.0 QAM AND VSB TRANSMISSION TECHNIQUES

2.1 Introduction

Two transmission methods, QAM and VSB, are the principle candidates for use with digital television systems. Both make efficient use of the available transmission bandwidth and may be implemented using highly integrated digital architectures. Though the transmission and reception techniques for the two are quite different, the actual transmitted waveforms are similar. Both are generated using similar quadrature modulation techniques (VSB can also be generated by direct filtering at IF). Both methods require the same amount of transmission bandwidth for a given data rate and their spectra are very similar in appearance.

The transmitted spectrum of VSB is distinguished from that of QAM by the presence of a small pilot tone at the carrier frequency. VSB demodulation requires a coherent carrier reference in order to reconstitute the original double-sideband signal. QAM demodulation, on the other hand, is able to recover the coherent carrier from the quadrature waveforms and so needs no such reference. The waveform equalization of the VSB signal is less robust than that of QAM, due to the way in which the carrier pilot must be used in VSB reception. A VSB demodulator must acquire the carrier pilot, via a relatively narrow tracking loop, prior to any equalization. A QAM demodulator, on the other hand, can perform blind equalization in the presence of carrier offsets and final equalizer convergence takes place after the carrier is removed via a relatively wide bandwidth tracking loop. VSB transmissions also frequently include periodic "training sequences" (strings of known bit patterns) to assist in adapting the equalizer to the channel.

For the same transmission data rates, these two modulations exhibit the same theoretical performance in the presence of additive white Gaussian noise (AWGN). Hence, bit-error rates versus E_b/N_0 (bit-energy-to-noise-density ratio) are the same for 8-VSB and 64-QAM as they are also for 16-VSB and 256-QAM. Figure 2-1 shows these two performance characteristics. These characteristics do not account for the additional VSB power due to the pilot tone. The pilot will decrease the VSB SNR by a few tenths of a dB, as noted later. The equalization length (complexity) requirements for these two, in the presence of multi-path (echo) distortion, are virtually the same, as is noted later.

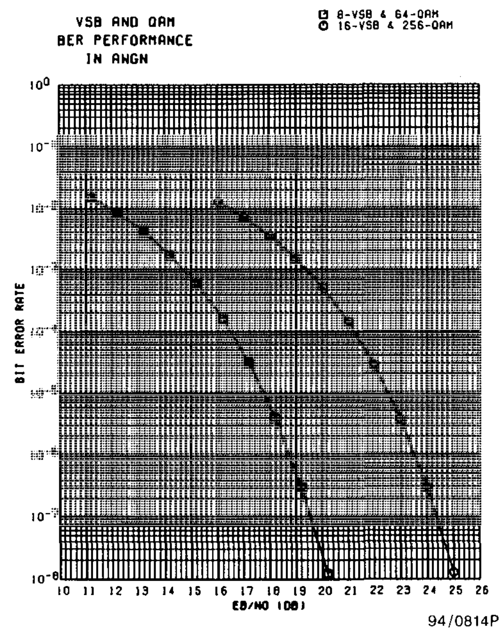

Figure 2-1. Ideal BER Performance for QAM and VSB

2.2 Generation of QAM and VSB Waveforms

In this section, the general methods for generating the two candidate waveforms are discussed. In order to make this discussion relevant to digital television transmission we will use in these discussions 64-QAM and 8-VSB signal types each operating at a gross bit rate of 32.4 Mbps — a nominal rate for digital TV transmission. With appropriate baseband filtering, either of these signals will fit within the 6 MHz channel allocation. (These same discussions can also apply to 256-QAM and 16-VSB — the number of amplitude levels and gross bit rates increase by a factor of 2.)

The block diagram of a typical 64-QAM generator is shown in Figure 2-2. The input bit stream (at 32.4 Mbps) is first multiplexed into two parallel streams of 16.2 Mbps. These two independent streams will be impressed upon the quadrature components of the QAM waveform. Each stream passes through an 8-level (3-bit) amplitude encoder. The rate of amplitude modulation at the encoder output is (16.2)/3 or 5.4 MHz. Since each stream carries 3 bits of information per level, the combined streams will convey 6 bits of data per transmitted symbol. The output data rate is then 32.4 Mbps, identical to the input rate. The amplitude encoded streams are passed through identical baseband filters in order to limit the transmitted spectrum to 6 MHz. The normal filtering used in this application has a square root, raised-cosine (SRRC) frequency response. A shaping factor of 10% will result in a transmitted bandwidth of 1.1x5.4 MHz or 5.94 MHz, just within the 6 MHz limit.

The filtered waveforms then are used to amplitude modulate two quadrature tones centered at the transmission center frequency. (The quadrature tones may actually be at an IF which is subsequently up-converted to the final transmission frequency.) The transmitted spectrum from this process is shown in Figure 2-3, with an SNR of 25 dB.

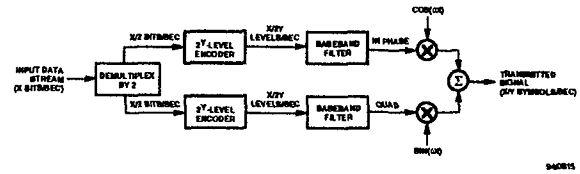

Figure 2-2. General QAM Generator and Modulator

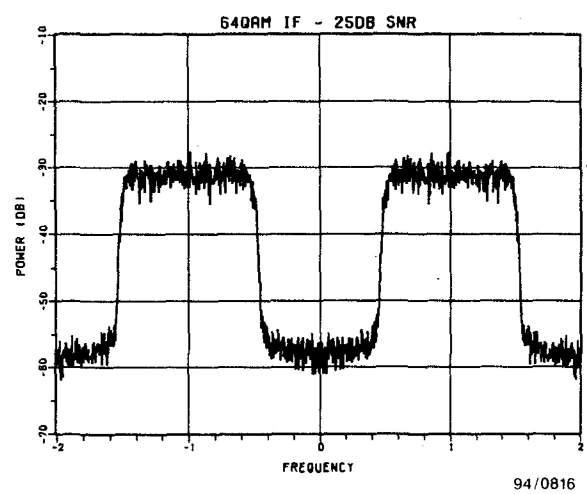

Figure 2-3. Spectrum of Transmitted 64-QAM Signal

The block diagram of the comparable 8-VSB generator is shown in Figure 2-4. In this method the input stream is first encoded into 8 levels (3 bits per level) resulting in an output symbol rate of (32.4)/3 or 10.8 MHz. This stream is next passed through a complex baseband filter producing two filtered output streams — a filtered version of the input stream and a phase-shifted (quadrature) version. The complex frequency response of this filter has a similar shape to that used in the QAM generator — in this example we will use the same 10%-shaped, SRRC response — with the filter center frequency offset from 0 Hz by one-half of the symbol rate. A small pilot tone at 0 Hz (carrier frequency) is also inserted at this point. The two streams are each carrying data at the full symbol rate of 10.8 MHz. The streams then amplitude modulate quadrature tones at the transmission frequency, which are summed together to produce the output waveform. Though the quadrature components carry full-rate data, the unique relationship between the two data streams results in a transmitted spectrum which is made up of one sideband of the quadrature components, a small vestige of the other sideband passed by the filter, and a small pilot tone at the carrier frequency. Figure 2-5 shows the transmitted spectrum of this signal. The pilot is approximately 11 dB below the signal power [6] and the SNR is 25 dB on the data portion of the signal. The pilot adds 0.3-0.4 dB to the total power. The required transmission bandwidth for 8-VSB is the same as that for 64-QAM.

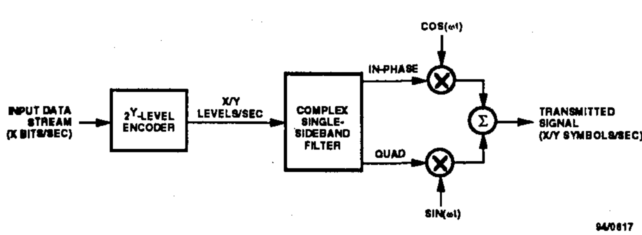

Figure 2-4. General VSB Generator and Modulator

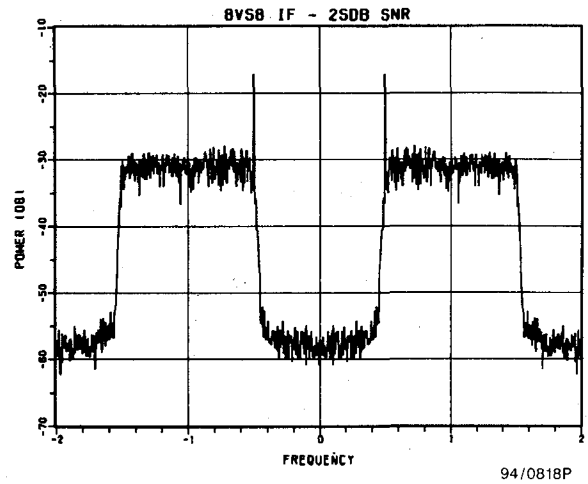

Figure 2-5. Spectrum of Transmitted 8-VSB Signal

2.3 Comparison of QAM and VSB Transmissions

The two techniques as discussed have very much in common. In Section 2.1 it was shown that for the same data rates and same AWGN channel the two have identical error rate performances ignoring the VSB power associated with the pilot. The baseband filtering used on each is very similar and can have identical frequency responses as shown in Section 2.2. Both transmitted waveforms consist of a carrier with quadrature amplitude modulation (for QAM the quadrature modulating data is independent, for VSB the data is identical but is a phase-shifted version of itself in quadrature). The required transmission bandwidths are the same for both modulations; any channel distortion will be observed to have a similar effect on the spectra of both signals.

The equalization complexity requirements are identical for the two modulations. QAM requires a complex structure (equivalent to 4 parallel real filters) while VSB requires only a single structure. However, for equivalent performance the equalizers must have the same duration impulse response. The VSB structure must run at twice the clock rate of the QAM, since VSB has twice the symbol rate. Therefore the VSB equalizer must have twice as many taps to achieve the same impulse response duration. For comparable equalization, the same number of multiply-accumulates per second are required of each structure.

Differences in the two techniques reside generally in the processing required for demodulation and in particular in the requirement for VSB to recover the carrier prior to equalization. A strong reliance is placed upon accurate carrier recovery in order to achieve high performance from VSB demodulation. Unlike QAM, the VSB equalizer cannot work in the presence of a non-zero carrier frequency, and small carrier phase errors or biases can cause performance degradations in the equalized signal. The use of training sequences in the data enable the demodulator to compensate for carrier phase errors and to assist the equalizer in initial adaptation. Figure 2-6 shows a complex baseband VSB constellation with distortion due to the transmitter baseband filtering. Since the received constellation possesses no angular symmetry it cannot be used for carrier acquisition; the carrier offset frequency is recovered from the received pilot. Reference [6] presents an architecture for performing some carrier phase correction following the equalizer; this, naturally, adds to the demodulator complexity.

QAM demodulation performance is also highly sensitive to received carrier phase. In fact, the performance loss resulting from carrier phase errors is virtually the same for both modulation types. Figure 2-7 shows the performance losses for QAM and VSB modulations resulting from carrier phase errors. However, the QAM demodulator is able to deal more effectively with system-induced carrier frequency and phase offsets because of the way the carrier is recovered in the demodulation process. The equalizer can begin the adaptation process in the presence of carrier frequency offsets. Figure 2-8 shows a noiseless 64-QAM constellation with distortion due to baseband filtering. The 4-quadrant symmetry of the 64-QAM constellation allows blind adaptation techniques to be applied during signal acquisition. Full carrier phase information is retained in the data at the equalizer output where noise and distortion has been largely removed from the signal. This allows more efficient carrier tracking to be used and thereby to very effectively track out carrier phase jitter. In addition, the complex equalizer is able to correct for any phase biases introduced in the tuner/demodulator hardware.

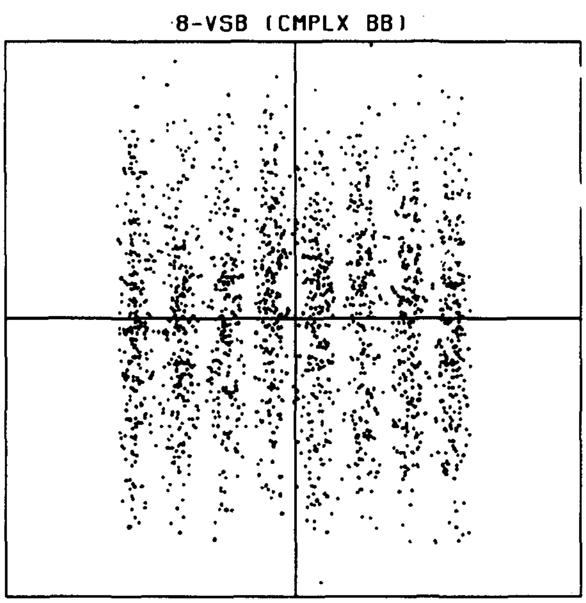

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Figure 2-6. 8-VSB Quadrature (Complex) Baseband Constellation

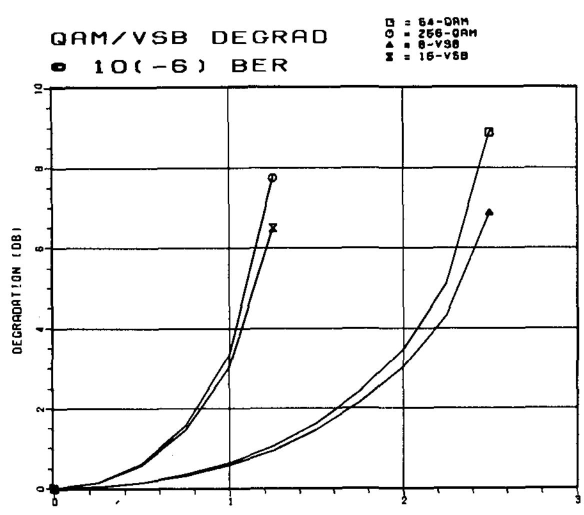

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Figure 2-7. QAM and VSB Performance Degradation from Carrier Phase Errors

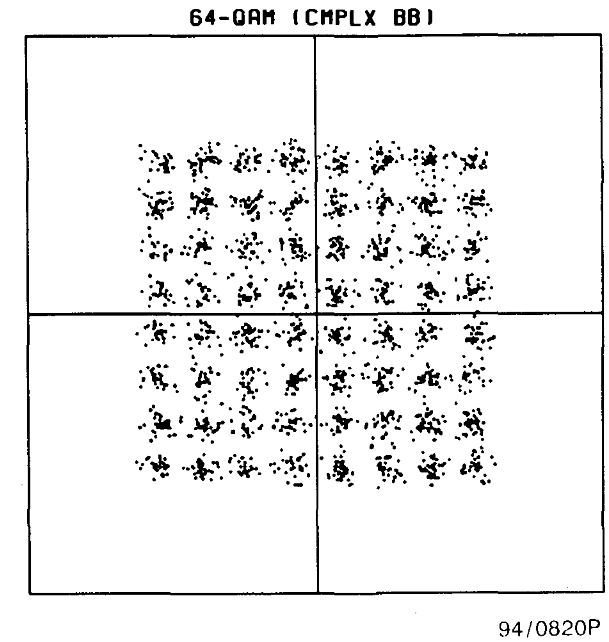

Figure 2-8. 64-QAM Quadrature (Complex) Baseband Constellation

3.0 CABLE IMPAIRMENTS

There are many impairments which need consideration when designing a QAM equalizer/demodulator for Cable TV. A brief description of some of these impairments is listed below.

3.1 Random Noise

Random or Gaussian noise is primarily added into the system via the trunk, bridge, and line extender amplifiers in a cable system. The figure of merit for these amplifiers is the noise figure, which is defined as the total measured output noise level of the amplifier minus the sum of the thermal noise and the amplifier gain. A cascade of identical trunk amplifiers with gain G dB is typically set up so that the cable between them attenuates the signal by –G dB. In this manner the noise level is increased solely by each amplifier's noise figure contribution. The noise figure of a cascade of identical amplifiers is then given as:

$$NF_n = NF_1 + 10 \log n$$

Thus, for example, every time the number of cascaded amplifiers is doubled, the noise level is increased by 3 dB. Increasing the random noise on a QAM signal has the effect of making each constellation point look more 'cloudy' or 'fuzzy.' Eventually, the amount of noise will begin to cause constellation points to cross the decision regions between points, causing errors or an increase in Bit Error Rate.

3.2 Composite Second Order (CSO) and Composite Triple Beat (CTB)

Of these two, CTB is typically the most problematic. Because of nonlinear characteristics in the trunk, bridger and line extender amplifier transfer characteristics, some mixing of the various video carriers occurs. Distortion products of the form $F1 \pm F2 \pm F3$ are called triple beats. F1, F2, and F3 are video modulated carriers. Thus, the more channels (or the more bandwidth) the system has, the more triple beats your system will incur. Triple Beats from different video carriers may fall at the same spectral location, adding with different phases and building upon one another to cause Composite Triple Beats. Second Order distortions are caused by the same mechanism, however push-pull amplifiers have overcome this problem, making it a non-issue. A very comprehensive study of the CTB/CSO problem was published in Reference [5]. The conclusion was that the CTB/CSO interference would cause little degradation (less than 1 dB) as long as the digital carriers were transmitted 8-10 dB below the analog carriers.

3.3 Microreflections or Multipath

Microreflections are caused by impedance mismatches in cable systems. If the terminating impedance is different than the characteristic impedance of the line, a portion of the incident wave is reflected back towards the source. The portion of the incident wave reflected is dependant upon the difference between the terminating impedance and the characteristic impedance, measured in return loss. If the incident signal ray arrives at the receive end accompanied by a delayed, attenuated version of itself the receiver must be robust enough to eliminate this distortion. Reflections can cause scalloping in the signal spectrum which the adaptive equalizer must equalize out by building a filter whose frequency response enhances the notches in the input spectrum, thus creating the inverse of the channel frequency response. Reflections between subscribers on separate taps are typically not a problem since the tap to tap isolation is usually around 30 dB. An example of a scenario where microreflections start to become a problem might be when the signal on one tap is connected to a splitter with low port to port isolation and sent to a separate TV with a relatively short run of cable. In this case, subscriber 1 may be watching a particular channel in one location while a viewer 2 may change to this same channel. In this case, the equalizer at subscriber 1 must respond to the change in channel conditions (i.e., level of reflection from the viewer 2 TV input port) that occurred during the step in return loss that occurred when the channel was changed. In Sections 3.8 through 3.10, results of actual and simulated equalizers on realistic microreflection environments are presented.

3.4 AM Hum

The trunk, bridger and line extender amplifiers are usually powered by a 60V, 60 Hz quasi-rectangular AC supply. This quasi-rectangular waveform can amplitude modulate the video signal within the amplifier. Although a single amplifier does not contribute significantly to hum, a long cascade of amplifiers can cause a buildup of hum to the point where it can appreciably distort the video signal. Other sources of hum can come from a wide variety of sources at the subscriber site. For example, if the cable carrying the video is in close proximity to fluorescent lights, a significant amount of AM hum can be imposed on the signal.

AM hum affects a QAM signal by elongating the constellation points in the radial direction. The elongation is proportional to the amount of AM hum present. As the elongation extends closer to the constellation point decision regions, an increase in the BER occurs.

3.5 Phase Noise

Phase noise can be thought of as random phase fluctuations imposed on the video carrier. In Cable TV systems, phase noise is imposed on the carrier by the LOs in upconverters, AML links, head-end modulators, and set top converter tuners. Phase noise forces the constellation points of a QAM signal to traverse a circular arc whose length is proportional to the amount of phase noise present. As the arc length extends closer to the constellation point decision regions, an increase in the BER occurs.

3.6 Residual FM

Low cost power supplies in set top converters which supply VCOs are the prime contributors to residual 120 Hz FM. Other residual FM sources include upconverters and modulators, however these are usually less significant contributors. The effect of the residual FM is to cause the signal frequency sweep between ±Fmax (the maximum frequency deviation) at 120 Hz and harmonics rates. The amount of deviation, Fmax, varies with tuners and can be from several kHz up to 50-60 kHz or more for the most inexpensive tuners.

3.7 Lab Tests

A series of in-house tests were performed on the Applied Signal Technology 256-QAM demodulator prototype to determine the viability of using this design as the basis for a VLSI implementation for the Cable TV industry. The tests were selected from a CableLabs report [7]. Only a subset of the tests were performed, which characterized the demodulator in the presence of:

1. Random Noise Interference
2. 120 Hz AM Hum
3. Residual FM
4. Residual FM and Noise
5. Channel Change
6. CW Interference

The results were preliminary in the sense that Applied Signal Technology intended to duplicate these tests at CableLabs in the near future. The following tests were not performed.

1. Composite Second Order (CW) Interference
2. Composite Third Order (CW) Interference
3. Composite Third Order (ATV)
4. CTB (CW) and Noise

These tests would be characterized later at the Advanced Television Test Center [8]. The QAM demodulator employed a 64-tap equalizer throughout the tests.

3.7.1 Random Noise Interference

3.7.1.1 Summary of Test Method

An HP3708 Noise and Interference Test Set was used to add Gaussian noise to the signal. The noise level was varied until an uncoded BER of approximately 10^{-4} was obtained (accounting for the effects of a simple error coding scheme). The noise power in a 5 MHz bandwidth was recorded.

3.7.1.2 Test Results

(Note that no coding gain has been applied. These are raw BER measurements.)

256-QAM: SNR = 31.4 dB
(~1.4 dB loss from theory)

3.7.2 AM Hum

3.7.2.1 Summary of Test Method

The HP 8782A Vector Signal Generator was used to produce the 256-QAM signal. The generator AM port was driven by a signal generator. The QAM signal was amplitude modulated using both a triangular and squarewave 120 Hz signal. The level of the modulation was varied to obtain a BER of approximately 10^{-4}.

3.7.2.2 Test Results

Hum Modulation:
256-QAM: 5.9%
(triangular modulation source)
4.2%
(squarewave modulation source)

3.7.3 Residual FM

3.7.3.1 Summary of Test Method

The HP 8780A Vector Signal Generator was used to produce the 256-QAM signal. This generator has an FM port which was driven by a signal generator. The QAM signal was then frequency modulated using a 120 Hz sinewave. The modulation was increased to obtain a BER of approximately 10^{-4}.

3.7.3.2 Test Results

Residual FM 256-QAM: 300 kHz

This result implies that low cost tuners can be used with 256-QAM.

3.7.4 Channel Change

3.7.4.1 Summary of Test Method

This test was performed using two different methods. The first method involved changing the channel from channel 13 to channel 12 on the set top converter and filming the constellation display using a camcorder. By counting frames and knowing the frame time, one could determine how long it took for the equalizer to acquire the new channel. The second method involved using a digital storage scope and triggering with a TTL signal from the set top box which indicated a channel change. The 'Error Output' on the back of the Anritsu BER Receiver was also monitored on the scope. The time difference between start of trigger and the Error Output signal reaching 0 errors was taken to be the channel change time.

3.7.4.2 Test Results

~ 0.5 seconds (AGC, carrier, and equalizer lock)

3.7.5 FM Interference and Noise

3.7.5.1 Summary of Test Method

The residual FM deviation is varied over the course of the test. At each value of FM deviation random noise in a 5MHz bandwidth is added until a BER of 10^{-4} is achieved. The Carrier to Noise ratio (C/N) is then recorded for each level of FM impairment.

3.7.5.2 Test Results

256-QAM	
Res. FM (kHz)	C/N (dB)
10.0	31.4
20.0	31.6
37.0	31.7
100.0	32.3

64-QAM	
Res. FM (kHz)	C/N (dB)
500	28
600	30 BER=10^{-5}
700	32 BER=10^{-5}

3.7.6 CW Carrier Interference

3.7.6.1 Summary of Test Method

A CW interference carrier was inserted at 0.5 MHz offset from the center of the 6 MHz band. The level of the interferer was raised until a BER of 10^{-4} was achieved.

3.7.6.2 Test Results

256QAM: CW C/I =26.9 dB (0.5 MHz from center of band)

3.7.7 Performance With a Low-Cost Tuner

In addition to these tests, the BER performance with standard off-the-shelf consumer grade tuners was evaluated. It could be hypothesized that the integrity of the IF filtering and the phase noise of future LOs in QAM tuners would have to be improved due to the complexity of the digital signals. This translates into a higher cost for the tuner. It would be advantageous from a cost standpoint if the demodulator were robust enough to work with low cost tuners.

Applied Signal Technology selected several consumer-grade tuners to determine the relative degradations imposed on a QAM signal. The SER performance (the BER curve can be derived by dividing the SER value by eight) for one of these tuners is plotted with white gaussian noise added to the signal. The results are shown in Figure 3-1. The 256-QAM test curve represents the results of sending a 256-QAM signal through the set-top box tuner and a 256-QAM prototype demodulator. The uncoded BER results are encouraging considering the coding gains that can be realized using Forward Error Correction.

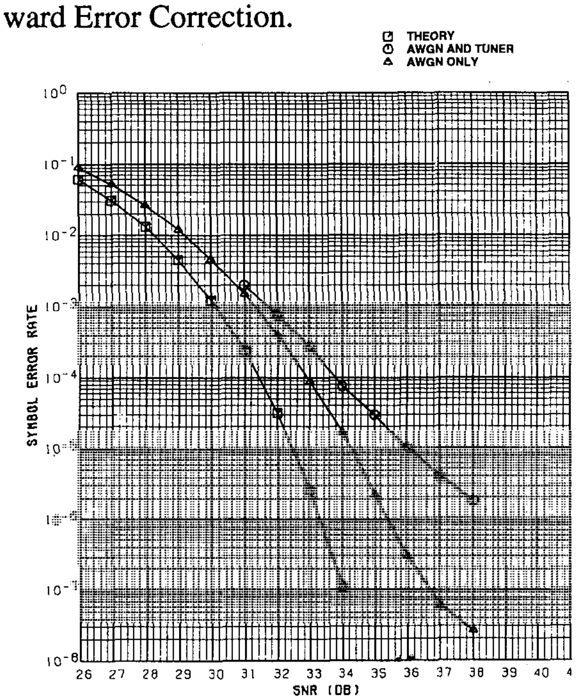

Figure 3-1. Uncoded 256-QAM Prototype Performance With and Without Low-Cost Tuner Distortion

3.8 Microreflection Degradations and Equalizer Performance

Computer simulations were performed to estimate the performance of a linear adaptive equalizer for 64- and 256-QAM for a variety of cable channels. Based on cable models of [9], microreflection models were derived for the trunk, distribution, and subscriber cable channels. In addition, models of some of the cable channels which were used in the recent CableLabs tests [8] were also evaluated.

The block diagram of the system which was simulated is shown in Figure 3-2. The baud rate was set to 5.287 Mbaud. The transmit filter which was used in the simulations was based on an actual transmit SAW filter.

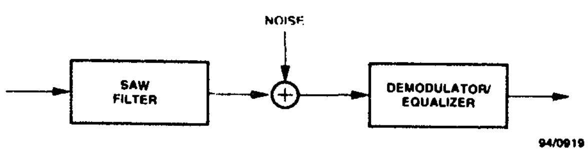

Figure 3-2. Simulated System

For each of the test channels, the optimal minimum-mean-square error equalizer coefficients were computed and the resulting error variance was determined as a function of input SNR and equalizer length. The error variance values were used to compute the symbol error rates for both 64- and 256-QAM modulations. The performance of T/2 spaced equalizers with 8, 16, 32 and 64 taps were compared to determine the minimum equalizer length required to produce performance close to ideal.

The reflections that occur within cable systems will now be evaluated. These reflections result in Inter-Symbol Interference (ISI) on the signal and can result in bit errors at the demodulator. Reflections will be evaluated at three different locations within a cable system: the Trunk environment, the Distribution environment, and the Subscriber environment. Each environment has been studied and simulated to determine the amount of equalization needed for the different cases.

3.8.1 Trunk Environment

The Trunk environment is a series of amplifiers which transmit the cable signal from a head-end or fiber node to a particular neighborhood. A typical trunk system has approximately 20 amplifiers that are spaced by 22 dB (2472 feet at 300 MHz) [10]. Figure 3-3 shows a typical microreflection scenario between amplifiers #2 and #3.

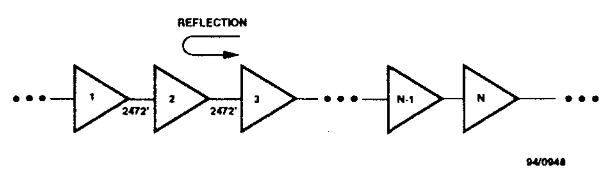

Figure 3-3. Trunk System

The power of the reflection (in dB) at the Nth stage is $10 \log (N * 10^{R_a/10}) + L_a$, where N is the number of amplifiers, R_a is the return loss of the amplifiers in dB, and L_a is the loss due to the attenuation of the line. Reflections in the trunks are characterized by relatively low amplitude reflected signals that are delayed in time on the order of microseconds.

Typical reflected powers with associated delays are:

Table 3-1. Reflected Power and Delay versus Frequency to the Trunk Channel

f (MHz)	Power of Reflection (dB)	Delay (µs)
55	–36.8	5.69
300	–63.0	5.69
450	–74.4	5.69
550	–82.8	5.69

From these results, it is clear that for the higher cable frequencies where digital signals will operate, the trunk environment results in essentially no degradation due to microreflections.

3.8.2 Distribution System

The distribution system is characterized by N equally spaced taps between line extenders, with each tap system assumed to have 5 unterminated taps. Reflections result from single reflections from nearby taps, double reflections from adjacent taps, and coupling between adjacent drops. See Figure 3-4. For analysis purposes, a software algorithm was used to generate the channel characteristics modeled in Reference [10].

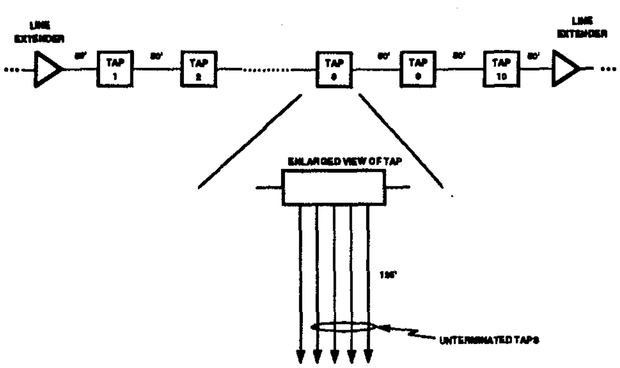

Figure 3-4. Distribution System

The assumptions placed on the channel [10] are shown in Table 3-2.

Table 3-2. Distribution Channel Parameters [10]

Number of Taps	10
Number of Drops per Tap	5
Line Loss (Taps & Drops)	0.46 dB/100ft and 1.55 dB/100ft
Delay	1.15 ns/ft
Tap Return Loss (Both sides)	20 dB
Line Extender Return Loss	16 dB
Distance between Taps	50 ft
Insertion Loss of Taps	~.8 dB
Tap Value	~21 dB
Tap to Output Isolation	25 dB
Tap to Tap Isolation	25 dB
Drop Length	125 ft
Return Loss of Drops	0 dB
Reflection of Drop	20 dB

The corresponding reflections at 55 MHz are illustrated in Figure 3-5a) and b).

The channel reflections are higher in the trunk environment than in the distribution system, but they are delayed by less time.

3.8.3 Simulation Results

Figure 3-6 illustrates the equalizer performance for the distribution system only for 64-QAM and 256-QAM, respectively. For 64-QAM, the 16-tap equalizer produces about 1 dB of degradation from ideal performance. For 256-QAM, the 16-tap equalizer suffers several dB of degradation at an uncoded symbol error rate (SER) = 1 x 10^{-6} and the 32-tap equalizer is about 1 dB from ideal. The 64-tap equalizer tracks the ideal curve with about 0.5 dB of loss.

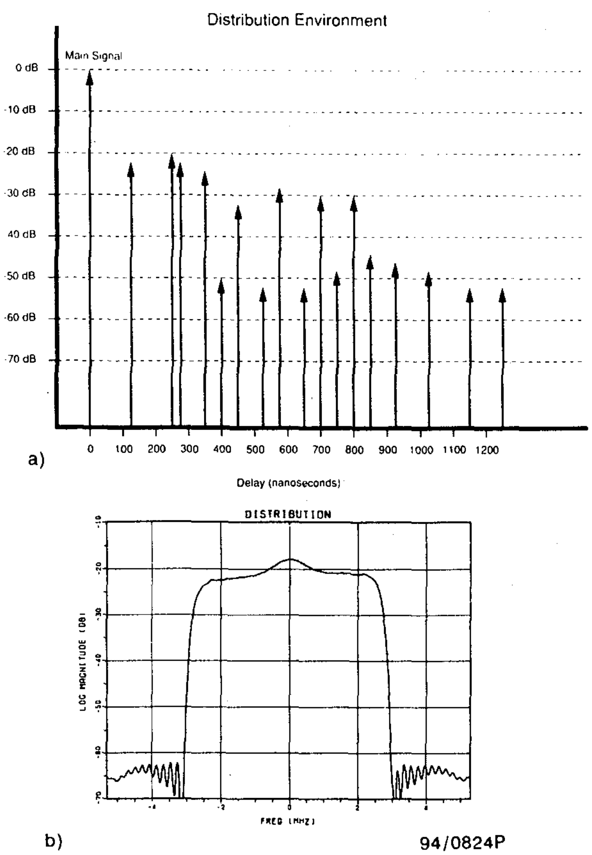

Figure 3-5. Distribution Environment Impulse and Frequency Response

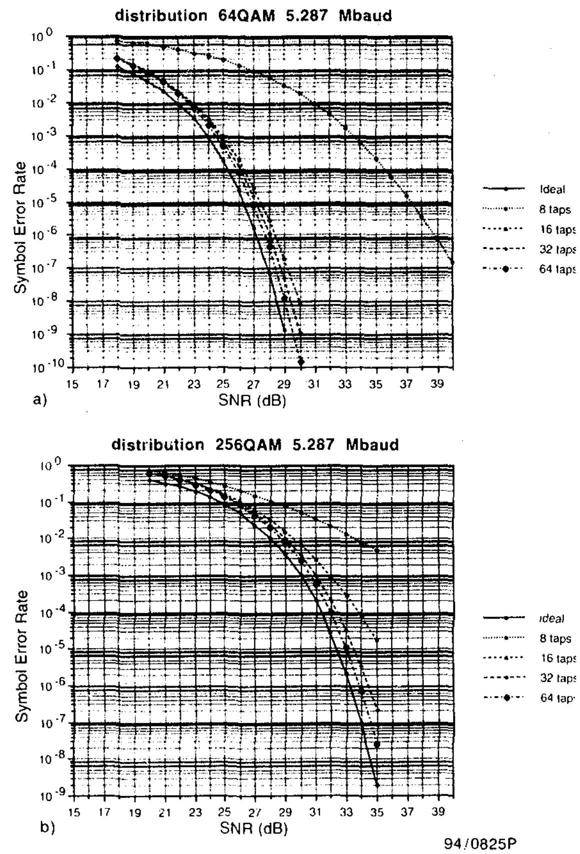

Figure 3-6. Equalizer Performance for Distribution Environments 64-QAM and 256-QAM

3.8.4 Subscriber

The subscriber environment is the reflections that take place once the cable signal enters an end-user's home. The different configurations are numerous, and a few reasonable scenarios were examined.

Two different channels were simulated. The first had splitter cable lengths of 5 feet, the length of typical cable in the living room and is illustrated in Figure 3-7. The second configuration used cable lengths of 50 feet, which is typical when TVs and VCRs, etc., are located in different rooms. The assumptions are shown in Table 3-3.

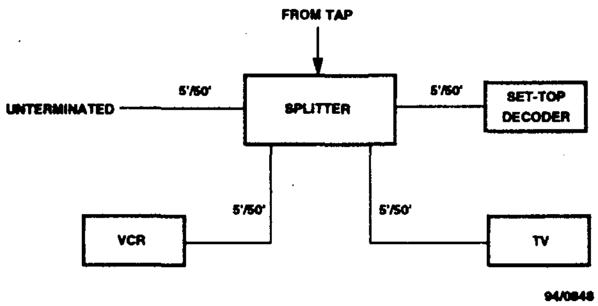

Figure 3-7. Subscriber System

Table 3-3. Subscriber Channel Parameters

Frequency	55 MHz
Number of Ports	4
Length of Cable on Ports	5 ft (case 1) 50 ft (case 2)
Termination Reflection Coefficient	0 dB
Port Reflection Coefficient	8 dB
Line Loss	1.9 dB/100ft
Delay	1.15 nS/ft
Isolation between Ports	15 dB

The corresponding 5-foot cable (case 1) reflections are shown in Figure 3-8a) and b), impulse response model and power spectra, respectively, and the 50-foot cable (case 2) impulse response model in Figure 3-8c).

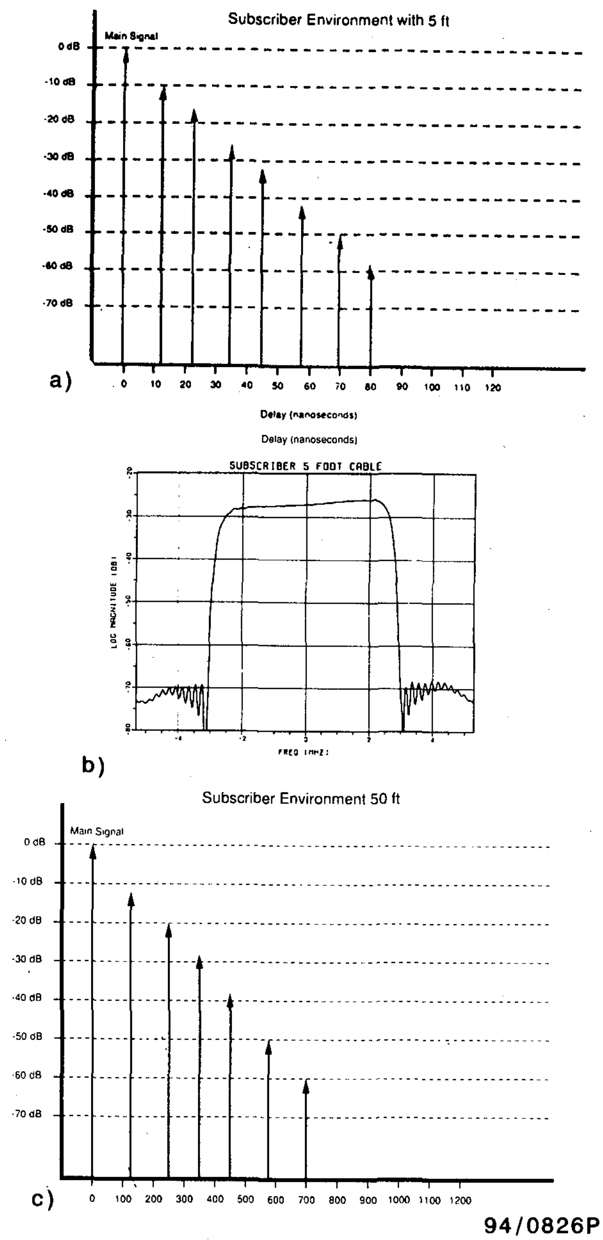

Figure 3-8. Subscriber Environment Models

Figure 3-9a) and b) illustrate the performance for the subscriber channel with 50-foot splitter cables for 64- and 256-QAM, respectively. For 64-QAM, the equalizer performance is within two dB of ideal for the 16-tap equalizer and within one dB of ideal for 32- and 64-tap equalizers. For 256-QAM, it is apparent that the 16-tap equalizer diverges from the ideal curve by a considerable amount; the 32-tap equalizer is within 1.5 dB of ideal while the 64-tap equalizer is within one dB of ideal.

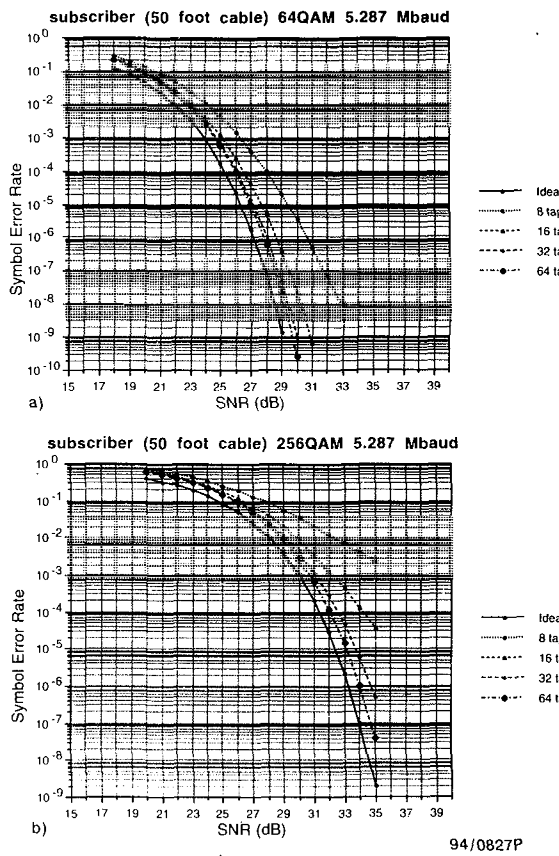

Figure 3-9. Subscriber Performance (50-Foot Cable)

Figure 3-10a) and b) illustrate the performance for the subscriber channel with 5-foot splitter cables for 64- and 256-QAM, respectively. For 64-QAM, the equalizer performance is within one dB of ideal for 16-, 32- and 64-tap equalizers. In particular, the 32- and 64-tap equalizers track the ideal curve very closely. For 256-QAM, it is apparent that the 16-tap equalizer diverges from the ideal curve by 2 dB at a SER of 10^{-6} while the 32-tap equalizer is within one dB of ideal. The 64-tap equalizer tracks the ideal curve very closely.

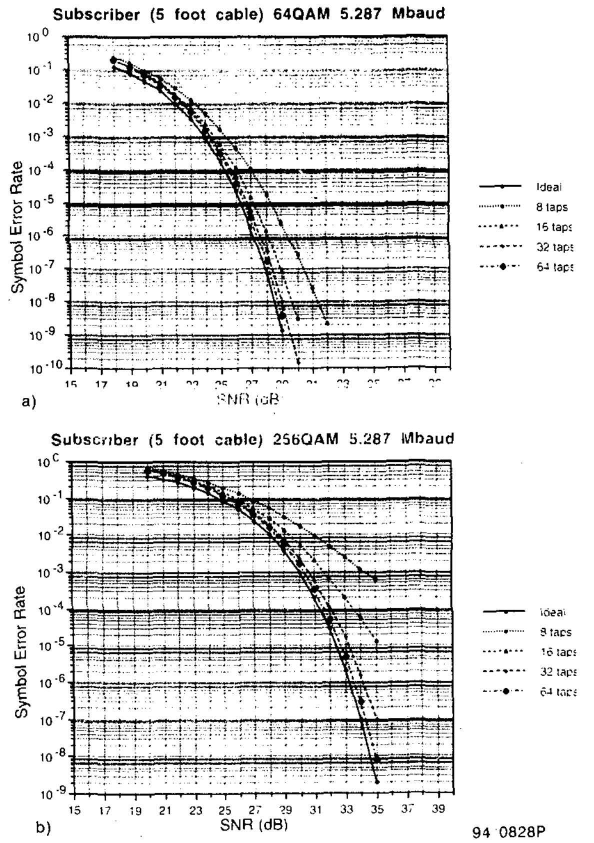

Figure 3-10. Subscriber Performance (5-Foot Cable)

3.8.5 Test Channels

Five of the channels which were used in the recent CableLabs tests [8] were also analyzed. Three channels were derived from Section 3.16 of the CableLabs tests which were used to measure the TV channel change acquisition time for the equalizer-demodulator. The parameters for CableLabs channels are shown in Table 3-4. The frequency response of these channels are illustrated in Figures 3-11 through 3-13.

Table 3-4. CableLabs 1 Channel Parameters

	Delay (ns)	Power (dB)
CableLabs 1	300	–18
CableLabs 2	2500	–20
CableLabs 3	600	–20

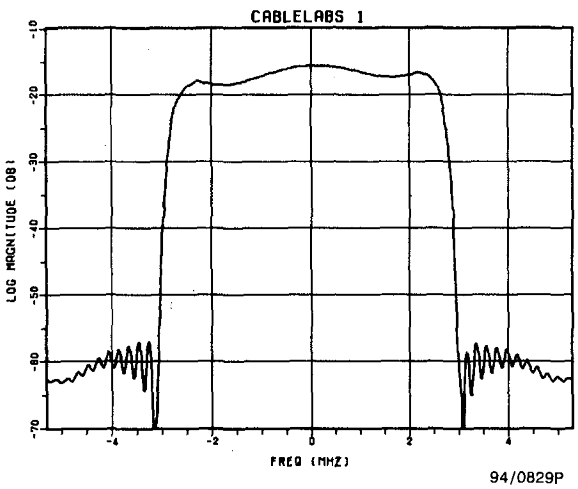

Figure 3-11. CableLabs 1 Channel Frequency Response

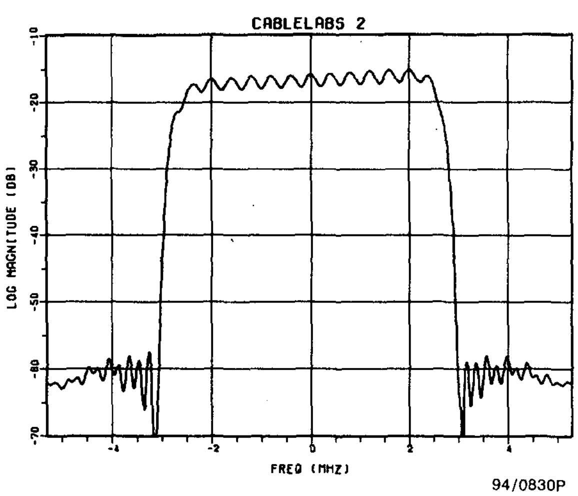

Figure 3-12. CableLabs 2 Channel Frequency Response

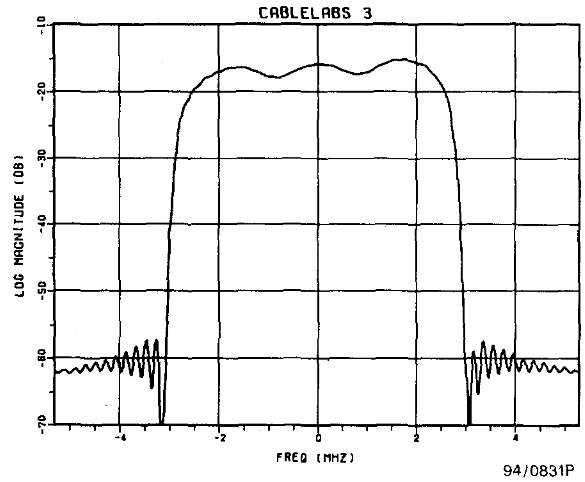

Figure 3-13. CableLabs 3 Channel Frequency Response

A strong reflection channel with reflection only –13 dBc was modeled in Figure 3-14 and a CableLabs recommended microreflection ensemble for general cable environments is shown in Figure 3-15. Parameters for the strong reflection model are given in Table 3-5 and for the microreflection ensemble in Table 3-6.

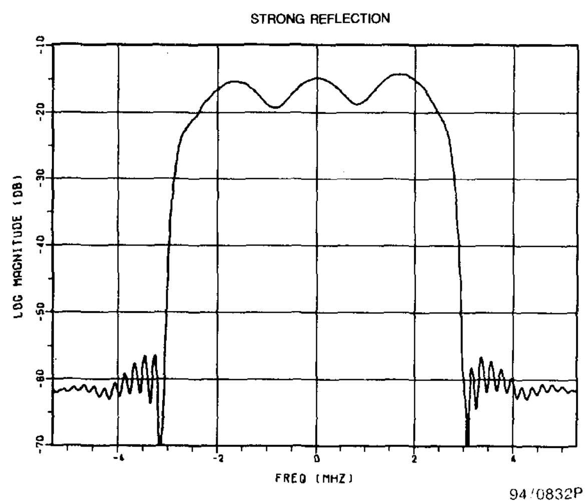

Figure 3-14. Strong Reflection Channel Frequency Response

Table 3-5. Strong Reflection Channel Parameters

Delay (ns)	Power (dB)
600	–13

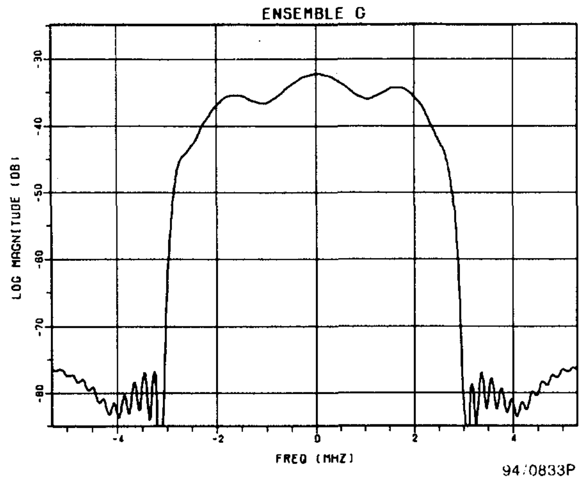

Figure 3-15. Microreflection Ensemble Channel Frequency Response

Table 3-6. Microreflection Ensemble Channel Parameters

Delay (ns)	Power (dB)
0	0
–200	–19
80	–22
150	–17
300	–22
600	–19

The SER performance vs. SNR curves for the CableLabs 1 channel for 64- and 256-QAM are shown in Figure 3-16 a) and b). It is apparent that for 64-QAM, the 16-, 32- and 64-tap equalizers produce less than 1 dB of degradation with respect to the ideal symbol error rate performance. For 256-QAM, the 16-tap equalizer produces about 2.5 dB of loss with respect to ideal and the 32-tap equalizer produces about one dB of loss. The 64-tap equalizer tracks the ideal curve fairly closely.

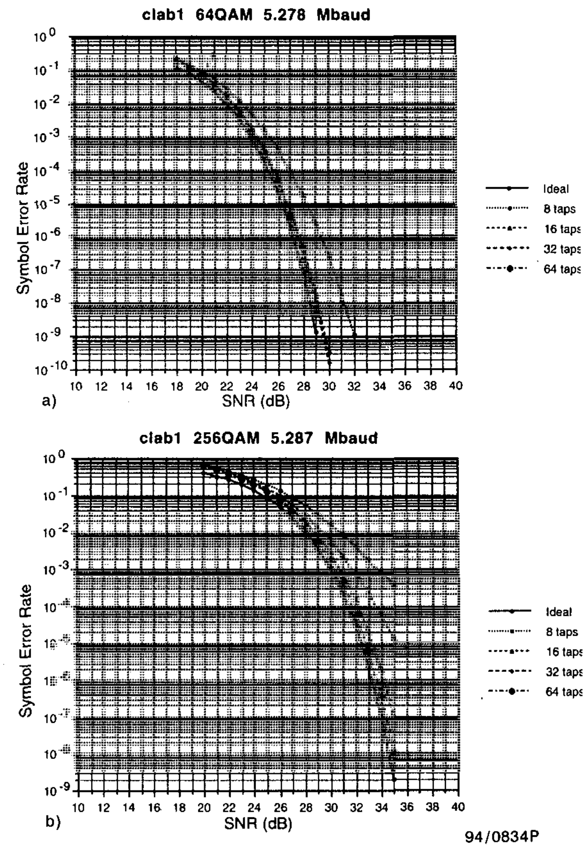

Figure 3-16. CableLabs Channel 1 Equalizer Performance a) 64-QAM b) 256-QAM

The SER performance vs. SNR curves for the CableLabs 2 channel are shown in Figure 3-17a) and b). Because of the long 2.5 microsecond delay of this channel, it is apparent that a 64-tap equalizer is required to produce performance close to ideal for 64-QAM. However, for 256-QAM the 64-tap equalizer produces about 2 dB of loss with respect to ideal; a longer equalizer or more coding gain may be required.

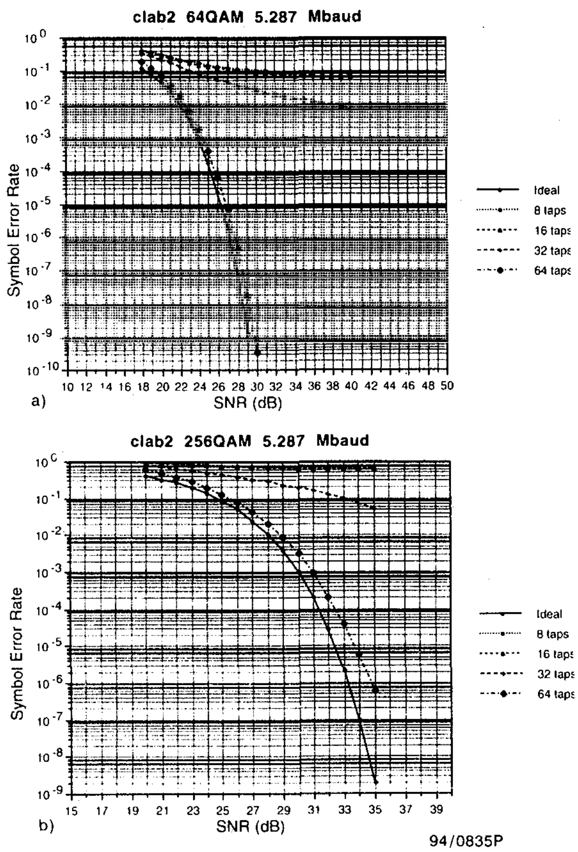

Figure 3-17. CableLabs Channel 2 Equalizer Performance a) 64-QAM b) 256-QAM

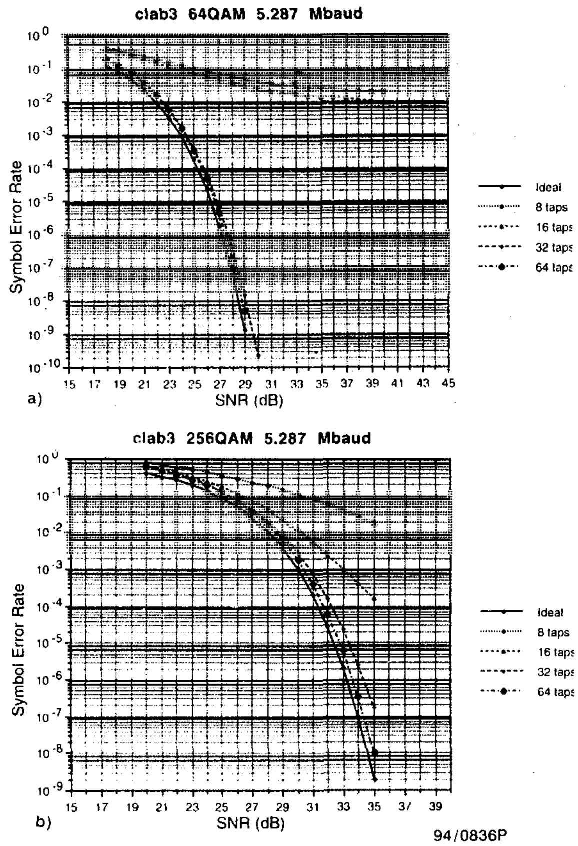

Figure 3-18. CableLabs Channel 3 Equalizer Performance a) 64-QAM b) 256-QAM

The SER performance vs. SNR curves for the CableLabs 3 channel are shown in Figure 3-18a) and b). For 64-QAM, the 16-tap equalizer diverges from the ideal curve while the 32-tap and 64-tap equalizers produce performance within one dB of ideal. For 256-QAM the 32-tap equalizer produces about one dB of loss at a SER of 10^{-6} while the 64-tap equalizer is within about 0.5 dB from ideal.

The SER performance vs. SNR curves for the strong reflection channel are shown in Figure 3-19a) and b). For 64-QAM, the 16-tap equalizer diverges from the ideal curve while the 64- and 32-tap equalizers produce performance within one dB of ideal. However, for 256-QAM the 32-tap equalizer is about 2 dB worse than ideal at a SER of 10^{-6} while the 64-tap equalizer is about one dB worse than ideal. Note that even though the echo delay for this channel is equal to that of the CableLab 3 channel, the equalizer performance with this channel is worse because the reflection is 7 dB larger.

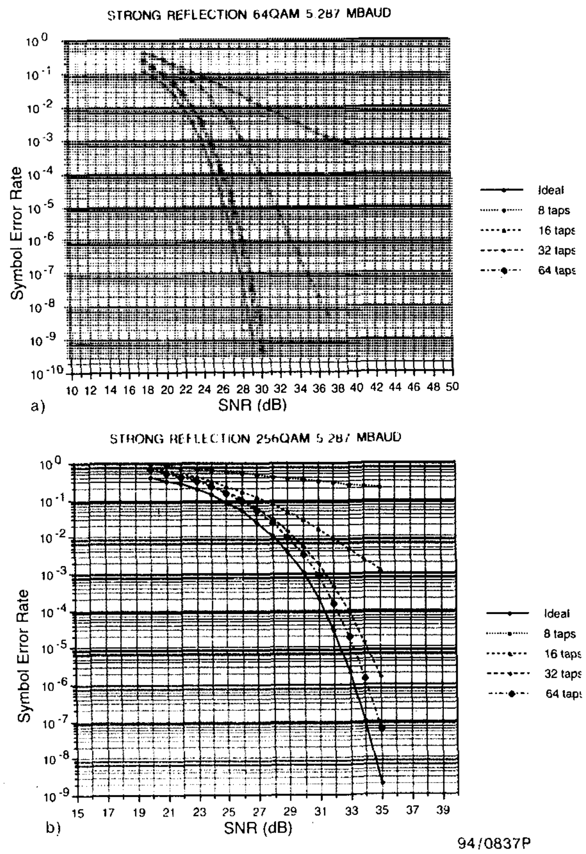

Figure 3-19. Strong Reflection Equalizer Performance a) 64-QAM b) 256-QAM

The SER performance vs. SNR curves for the Microreflection Ensemble channel are shown in Figure 3-20a) and b). For 64-QAM, both the 64 and 32-tap equalizers produce about 2 dB of loss with respect to ideal. For 256-QAM the 32-tap equalizer diverges from ideal and the 64-tap equalizer is about two dB worse than ideal.

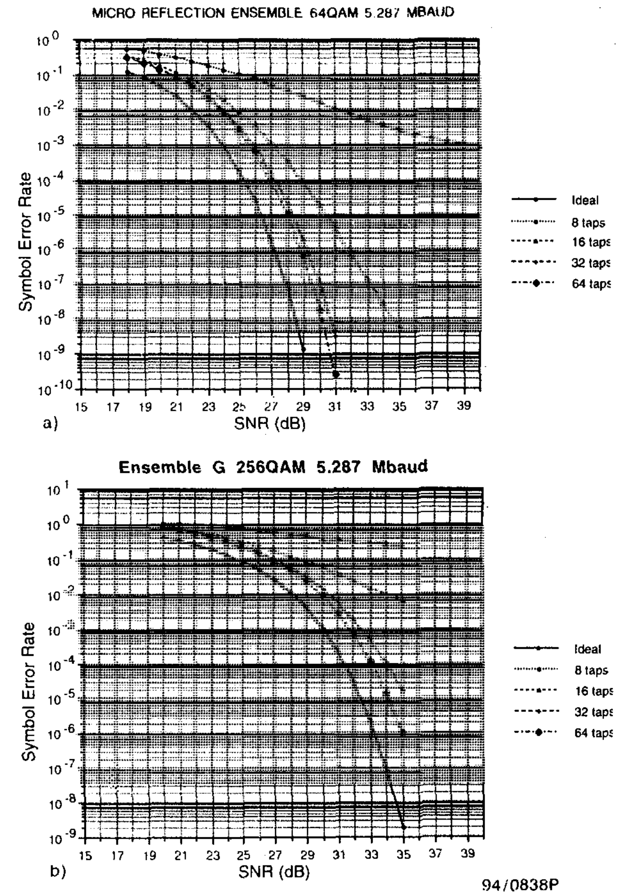

Figure 3-20. Microreflection Ensemble Equalizer Performance a) 64-QAM b) 256-QAM

3.8.6 Summary

The performance of a linear T/2 spaced equalizer was evaluated for a variety of cable channel models for 64- and 256-QAM modulations. The simulations indicate that for a 10^{-6} uncoded BER, 64-tap equalizer is sufficient to produce performance within one dB of ideal for both modulation types for the majority of channel models. There were a couple of cases with long delays such as the CableLabs 2 and Microreflection Ensemble channels where a longer equalizer is desirable to improve performance. However, for reasonably coded systems, the equalizer lengths of 16-32 taps appear to be sufficient.

3.9 ATTC Testing

In January 1994, a series of tests were conducted by CableLabs, Inc. at the Advanced Television Test Center as part of the evaluation of HDTV transmission techniques. The testing was extensive and comprehensive and included simulations of a number of realistic cable impairments. The results of these tests serve to support the conclusions in Section 2.0 regarding the comparative performance of QAM and VSB techniques. Performance tests of 256-QAM and 16-VSB demodulators were performed in the presence of phase noise, residual FM, carrier frequency offsets, and channel switching. The QAM performance was found to be comparable to or better than the VSB. These results could be attributed to features in the QAM demodulator architecture (as mentioned in Section 2.0) such as blind equalizer acquisition and robust carrier tracking on data only.

4.0 SUMMARY AND CONCLUSIONS

The 256-QAM modulation signal has been suggested as a reasonable capacity maximizing approach when cost, SNR, and various cable environment impairments are considered. Blind QAM equalization and carrier recovery used in the prototype 256-QAM demodulator was shown conceptually and in lab tests [8] (also compared to results of [7]) to offer equal or better performance than VSB systems without the need for an equalizer training sequence or pilot tone.

A BER of 10^{-4} was achieved in the presence of 120 Hz residual FM distortion consisting of a 100 kHz peak to peak deviation and a C/N of 32 dB. For 256-QAM, the prototype has been shown to track out the phase noise and residual FM inherent in an off-the-shelf analog tuner. The 256-QAM demodulator can perform channel change acquisition in 0.5 seconds or less by virtue of acquiring blindly without the need of a training sequence. Carrier offsets on the order of hundreds of kHz have been shown to be within the pull-in range of the demodulator on a consistent basis.

It has also been illustrated that the minimum required equalizer length appears to be 16-32 taps for most multipath scenarios encountered. Two different subscriber scenarios were modeled for simulation and certain worst-case assumptions were made concerning such parameters as isolation, terminations and return loss of cable elements. The resultant channel was simulated with several different equalizer lengths. It was found that in both cases 32 taps were sufficient. A 32-tap equalizer also proved adequate in handling several microreflection models recommended by CableLabs including a microreflection ensemble which consists of five rays between 80 and 600nsec with relative levels ranging from –17 to –22 dB from the main signal. However, for 256-QAM it was shown that a longer equalizer with 64 taps performs closer to ideal (within 2 dB at 10^{-6}) for microreflections which are –20 dB and delayed by 2.5msec, while the 32 tap equalizer BER curve diverged considerably for the same case. There are many microreflection scenarios that must be considered when designing a QAM equalizer for the cable environment. Several typical scenarios have been presented with results which are achievable with a low cost demodulator implementation.

This paper has described the virtues of 256-QAM for transmission of high capacity digital television signals, and contrasted some of these virtues with VSB modulation. Clearly, if careful architecture and design techniques are employed, the simulated and laboratory data presented in this paper confirm that 256-QAM modulation is eminently usable in the cable television environment.

ACKNOWLEDGEMENTS

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