Novel Hybrid CATV Network Architectures For Subcarrier Multiplexed Entertainment Video And Digital B-ISDN Services In The Local Loop

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ABSTRACT

Novel Hybrid CATV Network Architectures For Subcarrier Multiplexed Entertainment Video And Digital B-ISDN Services In The Local Loop

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This thesis addresses both the technical and economic performance issues of several novel local distribution fiber/coaxial CATV network architectures that are capable of supporting low-cost broad-band services including analog/digital entertainment video, and at the same time meet the demand of achieving a high optical power budget, required for deeper fiber penetration. Specifically, this work proposes and analyzes several novel local distribution CATV network architectures that utilize a 1.3 μm Semiconductor Optical Amplifier (SOA) based transmitter as well as a combination of fiber and coaxial cable as an upgrade to narrowband fiber-to-the-curb (FTTC) network

architectures. Fiber-to-the-home (FTTC) architectures will also be investigated. The main characteristic of the proposed network architectures is the innovation of using a 1.3 μm SOA as an external modulator at the transmitter end of the proposed network architectures. This represents a major milestone in achieving the above mentioned objectives. As will be shown, in addition to performing the modulation function, the required high optical power budget can now be obtained through 1.3 μm amplification and consequently there is no need to dismantle the already deployed 1.3 μm zero-dispersion single mode fibers.

The overall objective of this work is to investigate and analyze, through computer simulation and modeling, the performance of all the critical elements necessary for the implementation of high-capacity, high-performance, cost-effective local loop distribution CATV networks based on subcarrier-multiplexed fiber communications technology. We will implement a flexible, powerful computer modeling tool for evaluating the end to end performance of the proposed local distribution CATV network architecture.

Eventually, we want to be able to model the end-to-end performance of any signal that passes through the network. This will permit us to compare and devise candidate subcarrier-multiplexed local distribution CATV network architectures and determine the appropriate field of use for each. In addition, this will enable us to determine how to use the capabilities of each architecture to best advantage and provide insight into which Subcarrier-Multiplexed local distribution CATV architecture should be stressed.

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To my mother Carmen, wife Lindy-Ann and my beautiful daughters, Teraesa and Bria.

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CHAPTER 1

INTRODUCTION

1.1 Introduction

The concept of a broadband integrated services digital network (B-ISDN), as a means of providing integrated voice, data, and video services to residential and business customers has been established for a very long time [1], [2]. It is typically envisioned that single mode feeders will carry multigigabit/second signals to Remote Nodes (RNs) containing broadband electronic switching capability. From the remote node, each subscriber will be served by a dedicated fiber link delivering as much as 622 Mbps downstream to each home and carrying 155 Mbps upstream to the remote node.

Recently, however, there has been a growing recognition that the original proposal for B-ISDN may represent a technical solution which will not be cost-competitive, especially for the delivery of entertainment video to residential subscribers, and that there are other alternative approaches to provide broadband services that are likely to be easier to develop and less expensive to implement.

Subcarrier Multiplexing (SCM) of microwave or radio frequency (RF) carriers, applied in conjunction with passive optical splitting, is one of the alternatives which is now receiving a great deal of attention, as a means for building low-cost broadband Cable Television (CATV) networks. These subcarrier multiplexed systems [3 -5] have the ability to accomodate both analog and digital modulation, to handle voice, data, video, high definition video, and virtually any foreseeable combination of services. Their enormous flexibility makes them very attractive for broadband applications, especially if services are originating from a variety of different service providers, each requiring varying amounts of bandwidth.

Most of today's CATV industry use fiber for supertrunking : high quality point-to-point video interconnections between major system hubs, earth station / head -end connection, and links between head-ends. These fiber supertrunks have proven themselves to be highly reliable and cost effective, offering, in many situations, a viable alternative to microwave interconnection. A major objective of the telecommunication industry is to bring low-cost broad-band services to the subscriber using optical fiber. To justify the cost of an optical fiber subscriber network it is generally agreed that, in addition to voice and data, video services must be provided. First generation systems must be economically competitive with both the existing copper wire telephone network and CATV networks. They must also maintain a high degree of flexibility and upgradability so that they can evolve in step with an ever increasing consumer demand for broad-band services, and with future developments in video technology, such as digital or high definition television (HDTV).

Today's CATV networks can provide as many as 50 channels to a subscriber. It seems clear that a broad-band fiber video service must at a minimum be competitive with CATV, and provide either a comparable number of broadcast channels or on-demand video services, or both.

There is a growing consensus [6-7] that "POTS-only" fiber-based subscriber networks can be implemented in the near term with passive optical networks (PON) [8] and fiber-to-the-curb architectures. Such fiber architectures are expected to achieve cost parity with copper networks in the next few years by provisioning on the order of 64-128 living units with a PON having a 1 x 16 or 1 x 32 optical split. A number of previous studies have addressed the issues of how entertainment video services can be overlayed on such passive optical networks using AM / VSB [9], [10], broadcast FM [11] and switched FM [12], as well as high bit rate time-division multiplexed digital video [13], [14].

While each of these proposals has certain merits, each also has certain disadvantages which have prevented any one approach from achieving wide support. In brief, the AM-VSB approach has difficulty achieving the required power budget for a 1 x 16 PON, and is prone to many forms of transmission impairments. The FM approach requires high frequency electronic and optoelectronic components which limit the number of video channels, reduce the link budget, and increase the cost. The use of time-division multiplexed digital video at 70 Mb/s [13] or 45 Mb/s [14] requires digital electronics at gigbit rates which is costly and not really compatible with PON / FTTC architectures serving 64-128 subscribers per fiber loop.

The major underlying problem of all these approaches is that they have technical and economic difficulties achieving the high optical power budgets required for deeper fiber penetration in today's fiber / coax CATV networks. Two major factors contribute to this problem [15]. The first is that almost all optical fibers deployed are 1.3 μm zero-dispersion single-mode fibers (SMF's), and the second is that thousands of 1.3 μm distributed-feedback (DFB) laser transmitters and external modulation systems have been installed. All the existing 1.3 μm SMF's and 1.3 μm optical transmitters will continuously be used in their long lifetime. On the other hand, 1.55 μm transmitters have not been installed as widely as their 1.3 μm counterparts. Therefore, for deeper fiber penetrations, using 1.55 μm erbium-doped fiber amplifiers (EDFA's) may not be as economically viable as using 1.3 μm optical amplifiers. In addition, the high fiber dispersion (17-20 ps/km/nm) incurred by a 1.55 μm light beam in a 1.3 μm SMF can induce severe second-order nonlinear distortions (NLD's) [16] and consequently limit the transmission distance.

What is needed is a local distribution CATV network architecture, that is the focus of this thesis, capable of supporting low-cost broad-band services including entertainment video, and at the same time, meeting the demands of achieving high optical power budgets, by using 1.3 μm optical amplification, required for low-cost deeper fiber penetration.

1.2 Thesis Statement

This thesis addresses both the technical and economic performance issues of several novel local distribution fiber/coaxial CATV network architectures that are capable of supporting low-cost broad-band services including analog/digital entertainment video, and at the same time meet the demand of achieving a high optical power budget, required for deeper fiber penetration. Specifically, this work proposes and analyzes several novel local distribution CATV network architectures that utilize a 1.3 μm Semiconductor Optical Amplifier (SOA) based transmitter as well as a combination of fiber and coaxial cable as an upgrade to narrowband fiber-to-the-curb (FTTC) network architectures. Fiber-to-the-home (FTTH) architectures will also be investigated. The main characteristic of the proposed network architectures is the innovation of using a 1.3 μm SOA as an external modulator at the transmitter end of the proposed network architectures. This represents a major milestone in achieving the above mentioned objectives. As will be shown, in addition to performing the modulation function, the required high optical power budget can now be obtained through 1.3 μm amplification and consequently there is no need to dismantle the already deployed 1.3 μm zero-dispersion single mode fibers. It is interesting to note that semiconductor optical amplifiers (SOA's) have long been considered not suitable for CATV systems which transmit multi-channel AM-VSB video signals. The main reason is due to the carrier-density-modulation induced second order nonlinear distortions (NLDs). However, as will be shown in this thesis, when SOAs are configured in the network as proposed here, as opposed to their conventional use, it will then be possible to meet both the NLD and power budget stringent requirements of CATV systems. The overall objective of this work is to investigate and analyze, through computer simulation and modeling, the performance of all the critical elements necessary for the implementation of high-capacity, high-performance, cost-effective local loop distribution CATV networks based on subcarrier-multiplexed fiber communications technology. We will implement a flexible. powerful computer modeling tool for evaluating the end to end performance of the proposed local distribution CATV network architectures. The tool will include wavelength driven and time driven capabilities. The wavelength driven capability is required to model the effects of cascading many optical amplifiers and optical filters on the output signal to noise ratio and crosstalk level of any signal on the network. The time driven capability is required to model the effects of fiber chromatic and polarization dispersion, crosstalk from adjacent channels, timing jitter due to noise and crosstalk, thermal and shot noise in the receiver, and non-perfect extinction ratio in external modulators, on the signal quality in the network.The model allows several candidate modulation formats : Amplitude Modulation (AM), Frequency Modulation (FM) and Quadrature Amplitude Modulation (QAM), to be compared, identifies the most important sources of performance degradation, and allows overall estimation of the proposed networks' capacities. These simulation models are interconnected into a system level simulation model of the proposed subcarrier-multiplexed architectures. We demonstrate some of the capabilities of the simulation model through a number of examples at both device and system level.We will investigate the proposed Subcarrier Multiplexed networking architectures and technology. The network technology element of this work aims to define the physical components and network elements that are necessary to assemble a reconfigurable, diverse Subcarrier-Multiplexed network. The network architecture element addresses several topologies such as star, mesh and bus. These networks provide subscribers with access to switched video channels as well as a menu of broadcast channels, voice, and data. These networks can provide these services in either all analog format, all digital format, or a combination of both analog/digital formats.

Eventually, we want to be able to model the end-to-end performance of any signal that passes through the network. This will permit us to compare and devise candidate subcarrier-multiplexed local distribution CATV network architectures and determine the appropriate field of use for each. In addition, this will enable us to determine how to use the capabilities of each architecture to best advantage and provide insight into which Subcarrier-Multiplexed local distribution CATV architecture should be stressed.

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Chapter 2

Principles of Subcarrier Multiplexed CATV Systems

2.1 Introduction

Significant progress has recently been made in research on high-capacity subcarrier-multiplexed lightwave systems, mainly as a result of increased interest in finding near-term, low-cost solutions for the distribution of multichannel video services on fiber [1-3]. This progress in subcarrier-multiplexing technology has been greatly facilitated by advances in novel optoelectronic components, particularly in the areas of high-speed semiconductor lasers, photodetectors, and optical amplifiers. The terminology "subcarrier multiplexing" should be understood to encompass the multiplexing of both multichannel analog and/or digital signals [4]. The basic subcarrier multiplexed system is shown in fig (2.1). A number of baseband analog or digital signals are first frequency-division multiplexed by using local oscillators (LO's) of different radio frequencies. The upconverted signals are then combined to drive a high-speed light source (typically a laser diode). The LO frequencies are the so-called subcarriers in contrast to the optical carrier frequencies. At the receiver site, a user can receive any one of the Frequency

Division Multiplexed (FDM) channels by tuning a local oscillator and downconvert the radio frequency (RF) or microwave signals to baseband or intermediate frequencies (IF), similar to the way we tune in radio or TV channels. The main difference is that Subcarrier Multiplexed (SCM) lightwave systems can carry many more video, data, or voice channels than radio systems. SCM systems have an advantage over time-division-multiplexed (TDM) baseband digital lightwave systems in that services carried by different subcarriers are independent of each other, and requires no synchronization. In addition, SCM systems are presently still more cost-effective than high-capacity TDM lightwave systems - which is an attractive feature for near-term deployment in broad-band subcarrier loop systems.

The capability of a subcarrier-multiplexed lightwave system can be enhanced by including one or more optical amplifiers which provide system gain, either for increasing the span length or for power splitting in a broadband distribution system. For simulation purposes, the microwave subcarriers are represented as unmodulated carriers, and the composite signal driving the optical modulator is given mathematically by [5] :

$$I(t) = I_0 + \sum_{j=1}^{N} I_{max} \sin(\omega_j t + \theta_j(t)) \qquad (2.1)$$

where :

I (t) - total drive current applied to the modulator.

I_{max} - amplitude of each carrier signal.

ω_j - angular frequency of the carrier signal.

θ_j- random phase of the carrier.

2.2 Limitations Of Current CATV Systems

In order to understand the ways in which fiber may be useful in CATV, it is important to focus on the limitations of the present system architecture [6]. Fig (2.2) illustrates the kind of tree-and-branch architecture used in current coaxial CATV systems. All of the signals which are to be delivered to subscribers are gathered at a central head end. Typical sources are satellite earth stations, off-air antennas, videotape playback facilities, and supertrunks providing the delivery of signals from remote locations. At the head end, the various video sources are amplitude-modulated with a vestigial sideband (AM-VSB) at various radio frequencies, combined into a single broad-band signal, and transmitted over a single coaxial cable. This coaxial cable undergoes repeated branching through power splitters until it passes down each street in the commmunity. Broad-band amplifiers are required every 1000 to 2000 ft in order to overcome cable and branching losses.

This architecture is exceedingly cost effective and is the historical basis of the CATV industry. Nevertheless, it has a number of inherent problems and limitations. Fundamental to many of these problems is the fact that a number of broad-band amplifiers are required to operate in series, or cascade, in order to transport signals to system extremities. Each of these amplifiers contains active components and must be provided with power, limiting the reliability which can be attained. In addition, each amplifier adds noise and intermodulation distortions to the signals passing through it. The addition of these phenomena over long cascades of amplifiers gives rise to systems which have real limitations in the achievable reliability and quality of the service delivered to subscribers.

Another effect of tree-and-branch systems with long amplifier cascades is on system operating tolerances. In order to realize design specifications, each amplifier in such a system must be adjusted to provide very flat gain over a wide range of frequencies and must provide rather precise signal output levels. Such close operating tolerances require frequent alignment by highly trained technicians.

Another obstacle arising from this system architecture is a practical limitation on channel capacity. The types of coaxial cables used in CATV systems have a relatively wide potential bandwidth, perhaps exceeding 1 GHz. Such cables have, in fact, been used for many years in small apartment houses and hotel master antenna TV (MATV) systems to carry UHF channels, often at frequencies above 700 MHz. Typical cable systems operate with a highest frequency of only 300-400 MHz, however, with a few recent systems operating to 550 MHz. The difficulty in realizing the potential bandwidth of coaxial cable arises from the limitations of the broad-band amplifiers themselves, particularly when these amplifiers are operated in cascade. It is expected that it will be difficult to push channel capacity dramatically further than today's levels as long as CATV systems employ long cascades of amplifiers.

Most of these performance limitations, including those of reliability, transmission quality, and useable bandwidth stem from the long amplifier cascades required by typical CATV tree-and-branch architectures when used in medium to large communities. This, in turn, is the product of the relatively high loss of coaxial cables (on the order of 1 dB per 1000 ft: a loss of half the signal voltage every 600 ft). This loss, and the large number of amplifiers needed in series to counteract it, requires that CATV system bandwidth be limited far below the potential of the coaxial transmission medium in order to achieve acceptable signal transmission performance.

2.3 The Fiber Backbone

In view of the shortcomings of today's CATV system architecture and the practical constraints on outright network replacement, the American Television and Communications Corporation (ATC) developed an evolutionary concept for the integration of fiber into our systems. We have termed this approach fiber backbone [7-12]. The approach is illustrated in Fig (2.3), and essentially consists of overlashing some percentage of the existing trunk system with optical fiber cables. Thus, a direct optical fiber path is established from the head end to conversion nodes, a number of feed points in the coaxial distribution system. From that point on, the existing coaxial plant is utilized, with some amplifiers being reversed in direction and some spans of trunk cable between node areas being abandoned. The effect of this is to break up the CATV system into a number of very short systems. The length of those systems can be described by the maximum trunk amplifier cascade which will be allowed. At one extreme, fiber could be taken to every existing bridger amplifier location and the resulting coaxial system would consist only of feeder cable and line extenders. At the other extreme, a maximum trunk amplifier cascade of eight or ten might be defined, breaking a typical cable system into a few node areas. To illustrate, if present power supply locations were used as fiber node feed points, the average trunk amplifier cascade would be two, with a maximum of three or four, and the average node would service several hundred subscribers.

2.3.1 Implementation

In examining implementation of a fiber backbone system, the smallest problem is provided by the installation of fiber cables.Single-mode fiber has become relatively inexpensive in recent years and is available in a variety of cabled packages containing from 1 to 144 fibers in a physically rugged cable of ½ in or less in diameter. Field-enterable cable packages have been developed which allow the extraction and splicing of one or a few fibers from a multifiber bundle within a cable without the need to splice the other fibers. This type of cable would be particularly helpful in routing a single fiber to each node location in the fiber backbone approach.

The more challenging part of fiber backbone system implementation lies with the electrooptical components. These consist first of a laser diode transmitter feeding each fiber (or split to feed several fibers) leaving the head end. At each node location, an environmentally rugged optical receiver must be installed, capable of converting optical signals back to a broad-band RF spectrum suitable for coaxial distribution. The optical link must be relatively transparent to the CATV signals if the advantages of the fiber backbone approach are to be realized. For the sake of investigation, we have postulated minimum performance specifications for such a link. These are illustrated in table 2.1 below.

Table 2.1

Channels	60 - 80
C/N	55 dB
CTB	65 dB
CSO	65 dB
CROSS MOD	65 dB
POWER BUDGET	10 dB

It is apparent that, in a hybrid/coaxial CATV system, it will be highly desirable to maintain AM-VSB modulation throughout. If this could be accomplished, the only signal conversion required outside of the head end would be that from optical to RF at the end of each optical trunk. This approach greatly simplifies the electronics needed at each conversion point since it should be possible to directly detect the intensity modulation of light on the fiber, with the resulting detected output being the broad-band RF spectrum, a complex waveform complete with all the original channel information, scrambling, data carriers, etc. Such a conversion point could be contained in a small weatherproof housing, directly powered from the coaxial portion of the CATV system. Because AM-VSB modulation is relativelly fragile, however, this approach is technically quite challenging.

Fig (2.4) shows a block diagram of an ideal approach. At its head end, the broad-band AM-VSB signal, containing all of the cable TV channels, is used to directly intensity modulate a laser by varying the drive current. The information is transmitted optically, through the fiber, to a conversion node deep in the cable system where it is converted to broad-band RF using a simple detector. While this is consequently straightforward, the modulating signal may contain as many as 80 discrete television channels and there maybe a requirement for even more in the future. The bandwidth of this frequency division multiplexed signal may exceed 500 MHz. Constructing a link capable of transmitting this amount of analog information with sufficient transparency is no small matter.

2.4 Link Performance Requirements

In order to effectively evaluate the performance requiremments of an AM fiber link or any other link in the transmission chain, it is necessary to examine the effects which all of the various portions of the transmission chain have on the overall performance of the system. Even the best fiber links cannot provide acceptable picture quality when combined with substandard distribution systems. It is the combined performance of all the processing equipment between the program origination point and the television set which determines whether the subscriber is satisfied with the quality of the signal.

A system designer needs to be concerned with the contributions of the head end, microwave links, supertrunks, optical links, trunks, bridgers. line extenders, and subscriber equipment such as converters and set-top terminals as illustrated in

Fig (2.5). An understanding of how these dissimilar devices interact with each other to arrive at a combined operating performance and how the minimum standards which we establish for RF transmission translate into visible performance at a subscriber's television set is critical.

2.4.1 Noise

The cable TV industry has come to accept a set of subjective standards for noise impairments in its system which was submitted to the Federal Communications Commission in 1959 by the Television Allocation Study Organisation (TASO). This report established subjective ratings for various measurable levels of noise in television pictures and defined an excellent picture with no perceptible noise as one having a signal-to-noise ratio (S/N) of 45 dB or greater. Many cable system designers have accepted 43 dB as the minimum acceptable carrier-to-noise ratio (C/N) at the subscriber's TV set. This RF measurement, which can be defined as the ratio of the power of the carrier to the power of the random noise in the bandwidth occupied by the carrier, translates to a signal-to-noise ratio of about 41 dB (TASO) or 42.8 dB (CCIR) which lies between just perceptible noise and and no noise in the TASO ratings [13]. Although it may be more accurate to measure picture quality in terms of signal-to-noise ratios, the relationship between S/N and C/N is well defined and C/N is much easier to measure with test equipment which is available to the average technician, and defines the performance of the RF portion of the signal delivery path to the subscriber.

Thus, typical cable system specifications require a minimum of 43 dB C/N at the television receiver or 46 dB C/N at the output of the directional tap before the set-top converter which, typically, also has a C/N contribution in the 46 dB range.

2.4.2 Intermodulation Products

Intermodulation distortion exists in all cable systems due to small nonlinearities in the operating characteristics of the components which make up the distribution network. Early cable systems, which operated with 12 or fewer channels, were quite tolerant of intermodulation distortion because second-order products fell outside of the frequency bands where the carriers resided and higher order products were few in number and were at a low level, making their contribution almost negligible. Enhanced capacity systems, however, which make use of the midband and superband for carriage of signals, must pay close attention to intermodulation products if they are to deliver quality signals to their subscribers.

There are several distortions in cable systems, which are collectively referred to as intermodulation because they have similar effects and follow the same rules of accumulation in cascades of devices. Those distortions with which we are most concerned are cross modulation (Xmod), composite second order (CSO), and composite triple beat (CTB).

Cross modulation is a form of distortion where modulation from one channel is imposed onto other carriers in the system. It is generally indicated by the presence of 15.75 kHz sidebands on an unmodulated carrier, which show that synchronizing pulse information from other channels is being impressed on the measured carrier. Cross modulation differs from second order and composite triple beat distortions in that the net effect of cross modulation does not introduce spurious carriers. Cross modulation is created because the amplifiers we use exhibit a cube law transfer characteristic. That is, any increases in input signal cause a reduction in gain of the amplifier. Therefore, as varying modulation on one channel causes the input level to an amplifier to vary, the gain of the amplifier changes in response to these variations. This variation in gain effectively modulates the other carriers being passed through the amplifier.

Composite second order (CSO) distortion is the sum of second order beat frequencies in a channel passband which are created at the sum or difference frequencies of two other carriers in a nonlinear system (A ± B). Second order performance of broad-band amplifiers is considerably enhanced by the use of push-pull circuitry, which causes cancellation of even order intermodulation products.

Composite triple beat (CTB) distortion refers to the sum of all third order beat frequencies which fall within the passband of a channel. These distortions are generated at the sum and/or difference frequencies of three other carriers in a nonlinear system (A ± B ± C).

Generally speaking, second and third order intermodulation products are of most concern to cable systems. Higher orders are of either insufficient amplitude or fall outside the system passband.

Recall that nonlinear distortions are of two types : static and dynamic. Dynamic nonlinearities are detemined from numerical simulations of the rate equations of the travelling wave semiconductor optical amplifier or laser. Static distortions are calculated from the light versus current characteristic curve of the device. The static CSO, CTB, and Second Harmonic Distortion (2HD) in the ith channel can be obtained from the following equations :

$$CSO_i = \frac{\left(2\dfrac{L^{*2}}{L^{**}L_o m}\right)^2}{C_{i2}} \qquad (2.2)$$

$$CTB_i = \frac{\left(4\dfrac{L^{*3}}{L^{***}L_o^2 m^2}\right)^2}{C_{i3}} \qquad (2.3)$$

$$\frac{2HD}{C} = 20\log\left(\frac{mL^{**}L_o}{4L^{*2}}\right) \ dBc \qquad (2.4)$$

where $C_{i2,3}$ is the number of second- and third-order products in the ith channel respectively, m is the modulation index per channel, L is the output light power, L_0 is the output power at the bias point, L^* is the first derivative of the light - current curve at the bias point, L^{**} and L^{***} are the second and third derivatives at the bias point respectively.

Although there has not yet been a subjective study by an independent organization of the impact of intermodulation products on television picture quality, as in the case of the TASO noise impairment studies, most cable television companies have adopted minimum performance criteria for these distortions which are equal to, or very close to, the ratios listed below.

Minimum Intermodulation Performmance Criteria (**Table 2.2**)

Distortion	Minimum Performance w/o Set-Top Converter	Minimum Performance w/Set-Top Converter
Composite second order (CSO)	60 dB	55 dB
Composite triple beat (CTB)	53 dB	51 dB
Cross modulation (XMOD)	53 dB	51 dB

Note : The minimum performance criteria listed above are stated as a ratio in decibels of the carrier level of the channel being measured to the level of the intermodulation.

As cable systems added more and more carriers to increase the product offering available to their subscribers, intermodulation products displaced noise accummulation as the limiting factor in cascadeability of amplifiers. Manufacturers have developed new circuits which make use of feed forward and parallel hybrid technologies to limit the distortion contributions of amplifier stations but, in the case of systems approaching 80-channel loading, it is difficult to design equipment which allows cascades which could be achieved with 30 or fewer channels. This reduced cascadeability is one of the primary driving factors which led to the development of fiber backbone architecture. as well as the earlier supertrunking architectures which are in use today.

2.5 Optical Noise Performance

Now that we have looked at the effects of noise and distortions on the various combinations of CATV networks, we need to look at the noise and distortion parameters of the optical link and determine where improvements can be made.

The noise in a CATV optical link has several sources. We will calculate the noise contribution of the quantum behavior of light, thermal noise, and laser noise. We will assume that the fiber does not contribute to the noise. This assumption may be wrong if poor connectors or splices are being used [14], [15].

The calculations of an optical link carrier-to-noise ratio will be done with reference to currents at the detector in order to make it as simple as possible. The average photocurrent at the detector depends upon the laser power output, losses at the fiber, and the detector response.

Signal current :

$$<I_{ph}> = R * P * 10^{-\alpha * L/10} \qquad (2.5)$$

where

I_{ph} = photocurrent (A),

R = detector response (A/W),

P = laser output power (W) (average),

α = fiber loss (dB/km),

L = fiber length (km).

The signal current at each video channel is dependent upon the modulation index m, which is a measure of the ratio of the peak modulation of the individual channel to the average optical power. In common AM systems, the modulation index is twice or more of the reciprocal of the channel number, i.e., summation of all the modulation indices will give 200% modulation. System manufacturers rely upon the fact that the individual channels are not modulated coherently and the signals add up statistically, seldom reaching 100%, so the laser is operating in it's linear region [16].

Channel current :

$$\langle I_{ch} \rangle = \frac{1}{\sqrt{2}} * m * I_{ph} \qquad (2.6)$$

where

I_{ch} = rms channel current,

m = modulation depth (peak) per channel.

The quantum behavior of light and photocurrent gives rise to a statistical noise known as shot noise. The shot noise is a variation of the average current in the detector. We are neglecting the dark current contribution to the shot noise, since the detector dark current is usually much smaller than the photocurrent in AM systems.

Shot noise :

$$<I_{sh}^2> = 2 * q * I_{ph} * B \qquad (2.7)$$

where

I_{sh} = shot noise current (A),

q = charge of the electron ($1.6 * 10^{-19}$ C),

B = bandwidth (Hz).

The photodiode, being a current source, has its current converted to voltage over a load resistor. A thermally created current at this resistor is named thermal noise. The preamplifier noise is referred to as an increase of the thermal noise by a noise figure.

Preamplifier thermal noise :

$$<I_{th}^2> = 4 * K * T * B * F/R_l \qquad (2.8)$$

where

I_{th} = thermal noise current (A),

K = Boltzmann's constant (1.38 * 10^{-23} J/K),

T = temperature (K),

F = amplifier noise figure,

R_l = load resistance (Ω).

Laser noise is an inherent property of the lasers. It depends on the ratio between the drive current and the threshold current, and the higher the ratio, the better the performance [17]. Laser noise performance deteriorates if light is allowed to be reflected back into the laser from the transmission path [17], [18]. Distributed feedback lasers have less noise than Fabry-Perot lasers and the noise may be minimized by proper coating of the laser facets [19].

We will use the definition

$$RIN = 10\log_{10}\frac{d}{dF}\frac{<P_{\ln}^2>}{P^2} \qquad (2.9)$$

where

RIN = relative intensity noise of the laser (dB/Hz),

$P_{\ln}$ = laser power noise (W),

P = laser power (W).

Using [11] and assuming white RIN noise, we can obtain

$$<I_{\ln}^2> = I_{ph}^2 * 10^{RIN/10} * B \qquad (2.10)$$

where

$I_{\ln}$ = laser RIN current (A),

Carrier-to-noise ratio

$$CNR = \frac{I_{ch}^2}{<I_{sh}^2> + <I_{th}^2> + <I_{\ln}^2>} \qquad (2.11)$$

where

CNR = carrier to noise ratio (dB)

or

$$CNR = \frac{0.5 m^2 R^2 10^{-2\alpha L/10}}{2qRP10^{-\alpha L/10}B + 4KTBF/R_l + R^2 P^2 10^{-2\alpha L/10} 10^{RIN/10} B} \qquad (2.12)$$

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Chapter 3

Modeling and Performance Analysis Of Travelling Wave Semiconductor Optical Amplifiers For Use As External Modulators At The Transmitter End Of The Proposed CATV Networks

3.1 Background

Semiconductor laser optical amplifiers will play an important role in future optical systems, both for optical fiber transmission and for optical data processing [1]. These amplifiers can be used in both linear and non-linear modes of operation. Direct amplification of optical signals using semiconductor optical amplifiers is expected to increase the versatility of optical fiber transmission systems. It has been confirmed that a semiconductor optical amplifier, when used as a pre-amplifier in front of the photodetector, can improve the minimum detectable power compared to that in the direct detection scheme. Semiconductor optical amplifiers can also be used as optical linear repeaters to greatly expand the digital regenerative repeater spacing as well as reduce repeater size. Furthermore, they operate as simple, flexible repeaters, facilitating expansion of optical frequency multiplexed channels and changes of data rate and data formats. They can additionally be used as optical booster amplifiers to compensate for insertion loss and/or branching loss in optical circuits, optical switching systems, and local area networks.

Other areas of potential application include optical routing switches, pulse shaping and bistable elements, and external modulation.

This chapter focuses on using TWSOAs, for the first time to our knowledge, as external modulators at the transmitter end of the proposed subcarrier multiplexed CATV networks.

3.2 Recombination Mechanisms in Travelling Wave Semiconductor Optical Amplifiers

Recombination mechanisms are classified as either radiative or nonradiative in nature. The processes associated with the radiative recombination of electron-hole pairs in semiconductors are spontaneous emission, optical absorption or gain, and stimulated emission. Electrons and holes can also combine nonradiatively. Nonradiative mechanisms include recombination at defects, surface recombination, and Auger recombination, among others.The Auger recombination process involves four particle states (three electrons and one hole, two electrons and two holes, and so forth). In this process, the energy released during the electron-hole recombination is transferred to another electron (hole), which gets excited to a high energy state in the band. This electron or hole then relaxes back to achieve thermal equilibrium by losing its energy to lattice vibrations or phonons. The Auger recombination R_a may be approximately written as

$$R_a = CN^3 \quad (3.1)$$

where C is the Auger coefficient and N is the carrier density. The carrier lifetime τ_A for the Auger process is

$$\tau_A = \left(CN^2\right)^{-1} \quad (3.2)$$

Total recombination rate can be written as $R(N) = AN + BN^2 + CN^3$ **(3.3).** A is the inverse of the nonradiative lifetime in the presence of a trap, used to account for nonradiative recombinations, and B is the radiative recombination coefficient

3.3 Amplfier Model

In this section, the equations used to characterize and simulate the operation of the travelling wave semiconductor optical amplifier (TWSOA), are presented.

Specifically, the nonlinear model developed is used to examine the characteristics of the device when used for external modulation at the transmitter end of a subcarrier multiplexed CATV system. In particular, we will investigate the performance of the amplifier, when used as an optical modulator, as a function of the frequency and amplitude of the RF driving signal. Our formulation, which is a pair of coupled partial differential equations, is basically along the line of the analysis reported in [2-4]. The analysis with space- and time- dependence is necessary because the gain saturation behavior is, in essence, spatially nonuniform and temporally dynamic. The characteristics of the present treatment lies in the fact that, in contrast to the frequently used linear recombination model [2], [3], [5], [6], [7] which assume that the gain and carrier recombination vary linearly with carrier density, we have also taken into account the detailed nonlinear carrier recombination rate, R(N). The model used here makes several reasonable simplifying assumptions, which makes it uncomplicated and economical on computational power.

In the analysis, the signal light is assumed to be single mode. The carrier density N(z,t) is obtained by solving the z-dependent rate equation [4] :

$$\frac{dN(z,t)}{dt} = \frac{I(t)}{qV} - R(N) - \frac{g(N)}{\Gamma a E_{sat}} |E(z,t)|^2 \qquad (3.4)$$

where the diffusion term is ignored by assuming that the transverse waveguide dimensions are smaller than the diffusion length. E(z,t) is the total optical field in the waveguide, normalized such that $|E(z,t)|^2$ represents the optical power in watts, $E_{sat} = (h\nu Ar/\Gamma a)$ is the saturation energy of the amplifier, Ar is the cross-sectional area of the active region, and $h\nu$ is the photon energy. The gain coefficient g (N) is defined by :

$$g(N) = \Gamma a [N(z,t) - N_o]. \qquad \textbf{(3.5)}$$

Other parameters and their numerical values used in the simulation are taken from references [4], [12]. The propagation of the electromagnetic field inside the amplifier is governed by the wave equation [2], [3], [5]. Observing that the optical period is much shorter than the photon transit time through the amplifier, which, in turn, is much smaller than the timescale of envelope and gain variations, [8-9], one can integrate the wave equation to get [2-3] :

$$E(z,t) = E(0,t)\, e^{(1+j\alpha)\frac{1}{2}\int_0^z g(z,t)dz} \qquad (3.6)$$

or equivalently :

$$|E(z,t)|^2 = p_{in}(t)\, e^{\int_0^z g(z,t)dz} \qquad (3.7)$$

At z = L, equation (3.6) can be written as

$$E_{out}(t) = E_{in}(t)\, e^{G(t)\frac{(1+j\alpha)}{2}} \qquad (3.8)$$

where $E_{in}(t) = E(0, t)$ is the input field, $E_{out}(t) = E(L,t)$ is the output field, α is the linewidth enhancement factor, $P_{in}(t)$ is the input power to the amplifier, $G(t) = \int_0^L g(z,t)\, dz$, and $e^{G(t)}$ is the total large-signal optical power gain. To take amplified spontaneous emission (ASE) into account in our nonlinear model, an equivalent average spontaneous emission input power term [10-11] is added to the original input power $p_{in}(t)$ in equation (3.4).

The rate equation (3.4) and the wave equation (3.7) were numerically solved using a fourth-order Runge-Kutta routine. This gives the values of N (t) at any point z along the amplifier, which can be used with equation (3.7) to calculate the output power of the amplifier.

Unlike the approximate models previously proposed [2], [3], [5], to solve equation (3.4) the simulation technique used here provides an exact solution. In the following, the solution claimed as "exact" is intended to mean the exact solution of (3.4). The technique used here does not restrict in any way the class of input signals, and can handle very large gain saturation and gain fluctuations. The simulation block diagram of a Travelling Wave Semiconductor Optical Amplifier when used as an external modulator in the proposed Subcarrier Multiplexed CATV system transmitter configuration is shown in figure. (3.1). The video channel generator (VCG) generates a number (equal to the desired number of channels) of randomly phased tones at the desired video carrier frequencies. The output power of the CW laser is coupled into the amplifier modulator which is driven by the sum of a constant bias term (amplifier bias current) plus the output composite signal of the VCG, which, in turn, intensity-modulate the CW optical carrier via the amplifier modulator. The detailed amplifier model and the parameters used here are taken from [4], [12].

3.4 Amplifier Noise

The dominant noise generated by the amplifier is the amplified spontaneous emission (ASE). This noise results from the spontaneous recombination of electrons and holes in the amplifier medium. The noise can be modeled as a random process in which the events are infinitely short pulses distributed all along the active medium. Characterizing this random process by a flat noise - power spectrum, the ASE power at the output of the amplifier is given by [13 - 15] :

$$P_{ASE} = 2\chi N_{sp} h\nu (G-1) B_o \qquad (3.9)$$

where

B_o : is the optical filter bandwidth.

2 : two polarization.

h : is Planck's constant.

ν : is the optical frequency.

G : is the amplifier's optical power gain.

N_{sp} : is the spontaneous emission factor, given by :

$$N_{sp} = \frac{N}{N - N_o} \frac{\Gamma g}{\Gamma g - \alpha} \qquad (3.10)$$

where

Γ : is the optical gain confinement factor of the active layer.

α : is the effective loss coefficient per unit length which includes scattering and absorption losses both inside and outside the active region.

g : is the material gain coefficient.

N : is the carrier density.

N_o : is the carrier density at transparency.

In general, N_{sp} ranges from 1.4 to more than 4 depending both on the pumping rate and the operating wavelength [14-16].

χ : is the excess noise factor [13-14] given by :

$$\chi = \frac{(1 + R_1 G)(G - 1)}{(1 - R_1)G} \qquad (3.11)$$

where

R_1 is the input facet reflectivity of the amplifier.

From (3.10) and (3.11) we see that the noise power decreases when the input facet reflectivity R_1 decreases. The reason is that the part of the noise which has the greatest impact on the amplifier performance degradation is that which occurs where the signal is the weakest, that is, at the input to the amplifier. With R_1 small, the spontaneous emission directed toward the input facet can escape from the cavity and no longer contributes to the ASE. Since for a travelling wave amplifier, $R_1 \approx 0$ and $G \gg 1$, $\chi \approx 1$.

When the total output from the amplifier (amplified signal plus ASE) is detected by a photodiode, two distinct types of noise are created in addition to electronic (thermal) noise and shot noise. Since the power spectrum of the electrical current out of the photodiode is the result of a squaring operation upon the input optical power, the power spectrum of the photodiode output will contain not only a spike at DC, the signal-signal cross-product, which is the squared optical signal, but will also contain two noise cross-product continuum spectra, signal-spontaneous, and spontaneou-spontaneous. The spontaneous - spontaneous term has a spectral density which is much smaller than that of the signal-spontaneous component.

The photocurrent generated by the ASE is :

$$I_{ASE} = q\eta \frac{[P_{ASE}]}{h\nu_s} = q\eta N_{sp}(G-1)B_o \qquad (3.12)$$

where

q : is the electronic charge.

η : is the quantum efficiency of the photodiode.

P_{ASE} : is the average ASE input optical power.

ν_s : is the optical frequency of the input signal.

The noise terms due to the signal-spontaneous and the spontaneous - spontaneous beat terms are, respectively :

$$i_{sig-sp}^2 = \frac{2q^2F(G-1)}{h\upsilon} \qquad (3.13)$$

$$i_{sp-sp}^2 = 2q^2m_t n_{sp}^2(G-1)^2\Delta f \qquad (3.14)$$

where

F : is the excess noise figure.

n_{sp} : is the population inversion factor.

m_t : is the number of transverse modes.

Δf : is the optical bandwidth of the amplifier.

3.5 Small Signal Frequency Response Of The Travelling Wave Semiconductor Optical Amplifier.

The simulation block diagram in fig. (3.2) was used to investigate the small signal modulation response curves for the Travelling Wave Semiconductor Optical Amplifier. The curves were plotted for different combinations of CW input optical power and DC bias current. For each combination, small signal sinusoidal frequencies were used to modulate the DC bias. By performing an FFT on the output from the amplifier, the magnitude of the output at the applied frequency was determined.Graphs of relative magnitude (relative to the zero frequency component) versus modulating frequency were then sketched. These curves, shown in fig. (3.3) and fig. (3.4), show that for a given CW input power, the 3 dB bandwidth increases as the bias current increases. Also, there is very little dependence of the bandwidth on the magnitude of the input CW input power.

3.6 Characterizing The Amplifier Linearity

To assess the performance of the amplifier when used as an external analog modulator, both nonlinear distortion (NLD) mechanisms, i.e., the static distortion, which is caused by the nonlinearity of the amplifier output power -

current (P-I) curve under the CW condition, and the dynamic distortion, which is caused by signal-induced carrier density modulation, need to be carefully examined. The linearity of the amplifier is first characterized by using a single tone with a modulation depth m = 4 %. Here, the optical depth (OMD) per channel, m, is defined as :

$m = I / (I_b - I_o)$, where I is half of the peak-to-peak amplitude of the modulating signal, I_b is the amplifier bias current, and I_o is the current required to achieve transparency.

The static NLD is calculated directly from the P-I curve [12]. The main objective is to select the most linear region of the P-I curve over a wide bias current range, and then place the dc operating point roughly at the middle of this region. This is achieved by first calculating a set of P-I curves, each with a different value of the amplifier coupled input optical power P_{in}, over a wide range of bias current (0-300 mA) in 0.1-mA steps. For each value of P_{in} and bias current I_b, the output power, P, is obtained from a steady-state solution of the rate equation and the wave equation (equ (3.4) and equ (3.7)). Representative examples of these curves are shown in Fig (3.5) to Fig. (3.9). Then, second and third order harmonic distortions for constant OMD are calculated [12], for each curve, at various bias currents ranging from 40 to 150 mA. Calculations are performed by first getting a polynomial approximation for each P-I curve. From the formula obtained, dP / dI, d^2P / dI^2, d^3P / dI^3, etc. can be determined for any value of bias current. Once these derivatives are found, static distortion components can be calculated.

The simulation results indicate that the linearity of the P-I curve is strongly dependent on P_{in} and that the higher the value of P_{in}, the stronger the linearity (amplifier output power increases almost linearly with bias current when it operates well into saturation). At P_{in} = 4 mW, dP / dI is almost flat above I_o (about 10 mA), but starts to decrease above 200 mA. This indicates that the 10 - 190 mA region is the most linear region of the P-I curve for the typical amplifier parameters used here. With these optimal operating conditions (for best P-I curve linearity), the calculated values of the second (-91 dBc) and third (-118 dBc) order static harmonic distortions should meet the typical device specifications for a CATV trunk system [17].

The dynamic nonlinearity is calculated by analyzing the Fourier transform of the time-domain output optical power, P_{out}, from the amplifier which is calculated based on the detailed amplifier model reported in [18], [19]. For the dynamic measurements, single frequency tones at 4 % modulation index were applied to the dc bias of the amplifier. A dc bias of 100 mA was used, approximately at the middle of the 10-190 mA operating range of the static P-I curve obtained for P_{in} = 4 mW. Fig. (3.10) shows the second and third order harmonic distortions as a function of the modulating frequency for various values of P_{in}. As can be seen from the Fig, the dynamic distortion is also strongly dependent on P_{in} and that the higher the value of P_{in}, the lower the distortion (there is about 10 dB improvement when P_{in} increases from 0.1 mW to 4 mW). It can be seen that the second-order harmonic distortion does not meet the typical specifications for an AM -VSB CATV trunk system irrespective of the subcarrier frequency. Note, however, that at high input signal level (P_{in} > 1 mW), there is a window of low third order distortion (-85 to -100 dBc over the standard U.S. NTSC frequency plan) wide enough for CATV requirements. Accordingly, for AM-VSB modulation, system calculations are carried out by keeping the AM-VSB channels within an octave of frequency range and by operating the amplifier well into saturation.

Note, however, despite the fact that these second order nonlinear distortions cannot be tolerated by AM-VSB signals, they may be acceptable to subcarrier multiplexed digital video channels using M-QAM format and analog FM format.

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CHAPTER 4

AM-VSB CATV Distribution Networks Using Travelling Wave Semiconductor Optical Amplifier As An External Modulator

4.1 Introduction

In the last few years, several thousand fiber-optic AM-VSB CATV links have been deployed using distributed feedback laser (DFB) transmitters [1]. DFB lasers currently offer a working solution to AM fiber transmission, though not free of problems [2], [3]. Another alternative that has been used for AM video transmission is the use of $LiNbO_3$ external modulators and high power solid - state lasers [3].

In this chapter we propose a novel AM-VSB CATV link which utilizes a travelling wave semiconductor optical amplifier (TWSOA) as an external modulator and a high-power low RIN DFB laser as the CW source for the external modulator as another alternative to directly modulated DFB laser based transmitters [4]. We use computer simulation techniques to investigate the potential and limitations of TWSOAs when used as external analog modulators at the transmitter end of multi-channel AM-VSB CATV lightwave systems. The simulation block diagram of the proposed link is shown in Fig

(4.1). The video channel generator (VCG) generates a number (equal to the desired number of channels) of randomly phased tones at the desired video carrier frequencies. The output power from the CW laser is coupled into the amplifier modulator which is driven by the sum of a constant bias term (amplifier bias current) plus the output composite signal of the VCG, which, in turn, intensity - modulate the CW optical carrier via the amplifier modulator. The detailed amplifier model and its parameters used here are taken from [5], [6].

4.2 Proposed AM - VSB Link Performance

As an illustrative example, consider a 42-channel AM-VSB CATV system at 6-MHz intervals within an octave of frequency range from 300 to 546 MHz. An octave of frequency is used to avoid the high second order distortion of the device. Taking into account the fact that the NLD's are discrete intermodulation products as opposed to continuous noise sources, it is appropriate to treat the carrier-to-NLD (C / NLD) ratio and CNR (ratio of the carrier to the total noise power in 4 MHz bandwidth) separately. Our approach is to first determine the maximum value of m which satisfy the C / NLD requirements for a CATV trunk system. Second, by substituting the calculated value of m in (4.2) below, the maximum allowable system power budget is determined for a given CNR. Based on the power budgets, different system designs could be determined. The numerical simulation of the dynamic NLD is very simple conceptually. However, the execution of this simulation is quite computation-intensive [7]. We have chosen to tackle the problem numerically in a way similar to that of Phillips [7]. First, the composite time-domain modulation signal is generated as the sum of a constant bias term (amplifier bias current I_b) plus 42 randomly phased tones. Second, assuming 4 mW of CW input optical power coupled into the amplifier and an electrical drive current equal to that of the modulation signal plus a bias current of 100 mA, the time-domain output optical power of the amplifier, P_{out}, is calculated based on the detailed amplifier model reported in [5], [6]. Then, a Fourier transform is taken of P_{out} and the spectrum analyzed for distortion levels. The simulation distortion at a particular channel is determined by turning off the tone at that channel and summing the power that falls within its 6-MHz bandwidth. Because the result is strongly dependent on the phase of the carriers, the average is taken of a large number of runs, each with a different set of random phases. The number of runs was high enough to achieve a standard deviation of less than ± 1 dB, with a maximum of 200 runs. Fig.(4.2) shows the 42 AM-VSB channels from 300-546 MHz, and Fig. (4.3) shows the distortion in channel 41 when the carrier is suppressed. Fig. (4.4) show the numerically simulated carrier-to-composite triple-beat ratio (C / CTB) for channels 6, 21, and 41 (worst CTB distortion), as a function of the total rms modulation index $\mu = m \surd N /2$, where N is the number of channels. Also included in the figures is the static CTB distortion for the three channels.. As can be seen from the figure, the CTB distortion are < -60 dBc through the entire system as long as $m \leq 3\%$. Accordingly, when considering the link CNR at the receiver, a value of 3% will be taken as an upper limit for m. It should be pointed out that no clipping-induced distortion was observed in our simulation. For the link shown in Fig. (4.1), the CNR at the receiver is calculated based on the analysis presented in [2] where the extra noise generated by the TWSOA is taken into account. When the amplifier is used with coupled input power, P_{in}, and modulation depth per channel, m, the electronic CNR for any channel is [2] [see 4.1 shown below] where B_e is the electronic bandwidth (4 MHz), and the noise terms are, respectively, receiver thermal noise (7 pA / √ Hz), signal-spontaneous beat noise, relative intensity noise (RIN = -160 dB / Hz), and shot noise. Pin is the amplifier coupled input power (4 mW), G is the amplifier gain (10 dB), L is the total optical loss between the amplifier and the receiver (including the amplifier output coupling loss), F_0 is the amplifier's noise figure

(5.2 dB), η is the detector quantum efficiency (0.81), $\Re$ is the detector responsivity (0.85), and q is the electronic charge.

$$CNR = \frac{i_{sig}^{\;2}}{B_e[\,i_{th}^2 + i_{sig-sp}^2 + i_{RIN}^2 + i_{shot}^2]} \qquad (4.1)$$

$$= \frac{\dfrac{(m\Re P_{in}GL)^2}{2}}{B_e[\,i_{th}^2 + 2q\eta\Re P_{in}GL^2F_o(G-1) + (\Re P_{in}GL)^2 10^{\frac{RIN}{10}} + 2q\Re P_{in}GL]}$$

(4.2)

Fig (4.5) shows the CNR per channel for the 42 - channel system as a function of the total power budget, for various values of m (m = 2%, 2.8%, and 3%). As can be seen from the figure, at m = 2%, 2.8%, and 3%, a system budget of at least 12, 16.5, and 17 dB, respectively, is attained, while achieving a 50 dB CNR. The corresponding values of the worst CTB distortion are, -67, -62, and -60.2 dBc, respectively, indicating that the overall link performance can nearly always satisfy both the strict CNR and distortion requirements for a CATV trunk system. Even if the CATV trunk specifications were more stringent than those given above, our propose scheme can still meet these stringent specifications at the expense of a reduced power budget. For instance, a system power budget of about 5-6 dB can be attained while delivering > 52 dB CNR and < -67 dBc CTB distortion. Note that the signal - spontaneous noise alone imposes a limit on the CNR of $(m^2 \Re P_{in} / 4q\eta F_o B_e)$. Thus, to attain the high CNR values required for AM-VSB reception, it is necessary to operate the amplifier with a high coupled input power causing the amplifier to operate well into saturation, that is a power amplifier rather than an in-line or preamplifier. Consequently, operating the amplifier well into saturation is necessary to satisfy both the strict CNR and distortion requirements for a CATV trunk system. This indicate that the proposed scheme can still meet more stringent CATV trunk specifications and achieve higher system power budget than those given above by increasing the amplifier coupled input power ($P_{in} > 4$ mW).

It is interesting to compare the overall system performance of our proposed scheme with that of the more conventional approach reported in [8] that uses a directly modulated DFB laser followed by a TWSOA. In this conventional scheme, it was possible to transport 12 - channel AM - VSB television signals, provided the signal frequencies can be kept within one octave to avoid high second - order nonlinear distortion, and provided the amplifier can be operated in the linear gain region ($P_{in} < -15$ dBm), that is as an in-line or preamplifier, but not as a power amplifier, to avoid third-order nonlinear distortion [8]. In fact, when the amplifier output power exceeded + 2 dBm, the measured CTB distortion was greater than -40 dBc [8]. We have also carried out extensive simulations for the conventional scheme and the results are in qualitative agreement with these experimental results. In this case, however, in contrast to our proposed scheme, the simulation results show that adequate CTB distortion can only be obtained provided the amplifier is operated with $P_{out} < P_{sat} / 10$ (P_{sat} is the amplifier's saturation power) [9]. However, with such small output power, the amplifier's usefulness becomes marginal.On the other hand, if the OMI / channel is reduced significantly, the CTB distortion induced by the amplifier may become acceptably small, however, it would not be possible to obtain the high CNR values required for AM-VSB reception.

4.3 Network Architectures

4.3.1 General

Fiber to the curb (FTTC) systems have been proposed as a near-term, cost effective means for providing video services over fiber [10]. In an FTTC system , shown in Fig (4.6) , fiber is bought only as far as the Service Access Point

(SAP) where a shared Optical Network Unit (ONU) is installed. The ONU contains an optical transmitter and receiver, as well as multi/demultiplexing electronics and other hardware needed to process the video signals. Coaxial drops carry video signals from the SAP to several residences. Several SAPs may communicate with the central office (CO) over the same fiber through Subcarrier Multiplexing (SCM), Time Division Multiplexing (TDM), or some other technique. Due to the extensive sharing of the ONU hardware, the transmitter and receiver at the central office, and the fibers running to the service access point, the per subscriber cost of an FTTC system is reduced relative to the cost of a Fiber- to- the- home (FTTH) system that extends fiber all the way to individual customer premises Fig (4.7).

For FTTC systems, a broadcast delivery method (point-to-multipoint transmission of the full video menu) is the simplest and most straightforward way to provide video services in the near term. Either amplitude modulation (AM), frequency modulation (FM), digital modulation techniques, or a combination of analog and digital modulation formats can be used. These baseband signals can be combined, as shown in this thesis, by using subcarrier multiplexing.

While Fiber to the curb architectures may soon provide a cost effective means for delivering near term services over fiber, a variety of factors should eventually cause the fiber to be extended all the way to the subscriber premises. The availability of mass-produced, low cost, fiber optic transmitters and receivers should reduce or eliminate the need to share these components among subscribers, thus allowing for the installation of ONUs at individual customer premises rather than at shared service access point locations. Elimination of coaxial cable and twisted pair from the distribution and drop cables should greatly reduce or completely suppress the problems of corrosion and lightning damage that often plague metallic media.

4.3.1.1 Fiber To The Curb Systems

A possible Fiber To The Curb architecture is shown in Fig (4.8) The Remote Node (RN) houses the optical power splitter [11]. Each Video Optical Network Unit (V-ONU) does optical to electrical conversion and contains at least one electrical post amplifier. Four coaxial cables emanate from the V-ONU, with each bus having four " 15 dB " taps, and each tap servicing four subscribers. Sixty four subscribers are therefore served by each V-ONU. The network uses no coaxial amplifiers between the video ONU and the subscriber.

The video-distribution system thus requires active components only at the Central Office (CO), the V-ONUs, and subscriber set-top boxes. The set-top box at the customer premises is capable of tuning across the entire range of system carrier frequencies, demodulating individual carriers. We assume that each set-top box provides access to two channels, and that subscribers desiring four channels will have two boxes and a coaxial two-way splitter. The CNR (CNDR) at the set-top box can be written as the following ratio of mean-squared voltages :

$$CNR = \frac{<V_{st}^2>}{<V_{RIN}^2>_{st} + <V_{sh}^2>_{st} + <V_{revr}^2>_{st} + <V_{coax}^2>_{st} + <V_{NLD}^2>_{st} + <V_{sig-sp}^2>_{st}} \qquad (4.3)$$

where V_{st} is the signal voltage associated with a single carrier entering the set - top , $<V_{RIN}^2>_{st}$ is due to laser RIN, $<V_{sh}^2>_{st}$ is because of shot noise at the optical receiver, $<V_{revr}^2>_{st}$ due to receiver thermal noise, $<V_{coax}^2>_{st}$ from the coaxial system, $<V_{NLD}^2>_{st}$ and $<V_{sig-sp}^2>_{st}$ are TWSOA nonlinearity and signal - spontaneous noise terms respectively. $<V_{coax}^2>_{st}$ is different for each type of modulation format. Assuming negligible noise coming from the electrical post amplifier in the V-ONU, a 9 dB noise figure for the amplifier at the set-top box, and thermal noise power of -59.2 dBmV at the set - top, the following values are obtained for each format :

$$\text{AM-VSB} \quad - \quad 9.6499 \times 10^{-12}\ V^2$$

$$\text{M-QAM} \quad - \quad 1.3 \times 10^{-12}\ V^2$$

It can be shown that a combination of 1 / 2 inch coaxial cable, length 40 m, between the " 15 dB " multitaps and RG 59 cable in the drop, is sufficient to provide a worst-case signal power of -6 dBmV at the set - top box, for both modulation schemes. In addition to a 40 m drop length, the power budget also allows for 40 m of RG 59 inside wiring as well as a two-way splitter for those subscribers utilizing two set-top boxes. A typical design curve for these systems is shown in Fig. (4.9).

4.3.1.2 Fiber To The Home Systems

One of the simplest means of providing fiber sharing in an Fiber to the home architecture is shown in Fig (4.10). A passive 1:N optical power splitter located at the remote node (RN) divides downstream optical signals roughly equally amongst the N subscribers, and also combines light transmitted upstream by the subscribers. The passive nature of the link between the central office and customer premises has given rise to the generic term Passive Optical Network

(PON) to describe these splitter-based Fiber to the Home designs. The splitting of optical power among several subscribers in a passive optical network raises important power budget issues. With N = 16, for example, the power splitting results in a received optical power at each ONU approximately 14 dB lower (12 dB inherent in the 1: 16 splitting, plus 2 dB due to excess loss and splitting nonuniformities) than in the case of the fiber to the curb architecture; hence, any video distribution techniques employed in a passive optical network must be able to maintain adequate signal to noise ratio (SNR) performance at these lower levels of received power.Other possible FTTH system configurations are shown in Fig. (4.11) (bus) and Fig.(4.12) (ring topology).

In both fiber-to-the-curb and fiber-to-the-home architectures, Erbium Doped Fiber Amplifiers (EDFAs) can be deployed, so as to improve the system power budget [12] and therefore allow longer transmission distances or additional power splitting. EDFAs have several advantages over conventional optical repeaters in which the input optical signal is first optically-to-electrically converted, then processed in the electrical domain and reconverted to an optical signal. The main advantages of EDFAs include signal amplification regardless of its format or bit rate [13-14], simple configuration, and low cost [15].

EDFAs, however, will introduce optical noise and distortions. The latter is caused by the amplifier's gain tilt. The optical noise is due to three sources : Spontaneous shot noise, Spontaneous-Spontaneous beat noise, and Signal-Spontaneous beat noise. At high pump levels, only signal-spontaneous beat noise is significant, and is given by the following equation :

$$N_{sig-spon} = (4 I_d^2 \frac{n_{sp}}{C_1} h \frac{\nu}{P_{input}}) B \qquad (4.4)$$

where :

$N_{sig-spon}$ is the signal-spontaneous beat noise.

I_d is the average detected photocurrent

n_{sp} is the EDFA inversion factor.

C_l is the EDFA input coupling losses.

h is Planck's constant.

ν is the optical frequency.

B is the signal bandwidth per channel

P_{input} is the input optical power to the amplifier.

Also, if the chirp from the modulator is negligible, then the distortion resulting from gain tilt is very small.

At present, optical amplification is only practical in the 1550 nm wavelength region. Current EDFAs only operate efficiently in the 1550 nm wavelength regime and can boost signal power up to 20 dBm.

4.3.2 Proposed AM-VSB CATV Network Architectures

For the AM-VSB system, we will design both a Fiber-to-the-curb (FTTC) and a Fiber-to-the-home (FTTH) network architecture.In these and all subsequent designs in the ensuing chapters, the following design parameters will be used :

- Optical Wavelength - 1.33 μm
- Fiber attenuaion - 0.4 dB/km
- Length of fiber cable to distribution neighbourhood - 10 km
- Loss per two-way split in power splitter - 3.5 dB
- Amplifier output coupling losses - 4.0 dB

Also, all FTTC systems will use the design procedure and parameters found in 4.3.1.1

For the AM-VSB link, an optical power budget of 17 dB was obtained for the worst channel (# 41). Using a Passive Optical Network, only four subscribers can be served, Fig (4.13). Using the FTTC approach, 128 subscribers can be serviced, Fig (4.14).

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Chapter 5

M-QAM Compressed Digital Video Signals Transmission for CATV Distribution Networks Using Semiconductor Optical Amplifier As An External Modulator

5.1 Introduction

New digital compression standards, being developed by the Motion Picture Experts Group are expected to convey full-motion video signals at 4-5 Mb/s at a quality high enough for subscriber distribution [1]. Up to 4 / 6 such signals can be placed on one 16 / 64- level quadrature amplitude-modulated (16 / 64-QAM) carrier operating with a total carrier-to-noise-and distortion ratio (CNDR) of 20 / 32 dB with forward error correcting codes [1-2], and transmitted on a 6 MHz bandwidth.

This chapter proposes an optical CATV digital link that transmit hundreds of M-QAM compressed digital video signals by using a TWSOA as an external modulator and a high-power DFB laser as the CW source for the external modulator. Semiconductor optical amplifiers have long been considered unsuitable for CATV systems which transmit multi-channel AM-VSB video signals, due to the carrier-density-modulation induced second order nonlinear distortions (NLDs) [4]. However, despite the fact that these second-order NLD levels cannot be tolerated by AM-VSB signals, they may be acceptable to subcarrier-multiplexed digital video channels using M-QAM format [3]. We use computer simulation techniques to investigate the potential and limitations of TWSOAs when used as external modulators at the transmitter end of multi - channel M-QAM CATV digital lightwave systems. The simulation block diagram of the proposed optical link is shown in Fig (5.1). A " MATRIX Generator " (MG) generates a number of randomly phased unmodulated carriers at 6 - MHz intervals, at the desired subcarrier frequencies. The output power of the CW laser is coupled into the amplifier modulator which is driven by the sum of a constant bias term (amplifier bias current) plus the output composite signal of the MG, which, in turn, intensity modulate the CW optical carrier via the amplifier modulator. The detailed amplifier model and its parameters used here are taken from [5]. As for the AM-VSB system in chapter 4, the amplifier bias current is 100 mA and the CW input optical power is 4 mW.

5.2 Proposed M-QAM Link Performance

As a proposed system, we consider M-QAM digital video signals transmitted in the proposed CATV link shown in Fig. (5.1). We assume that there are N channels in this system, and that the M-QAM digital video signal of each channel is compressed to fit into 6 MHz bandwidth of the standard NTSC frequency allocation. Specifically, we consider a 100- channel 16 / 64 QAM CATV system at 6 MHz intervals from 139.25 MHz - 733.25 MHz. Since this system operates over a bandwidth exceeding one octave, second order intermodulation products formed by the sum and difference frequencies of subcarriers now lie within the transmission band and should be taken into account. For digitally modulated subcarriers, the intermodulation products can be treated as white noise [1-2]. Therefore, the total carrier-to-noise-and - distortion ratio (CNDR) can be expressed as :

$$1 / CNDR = (1 / CNLD) + (1 / CNR) \quad (5.1)$$

where CNLD is the carrier-to-NLD ratio, and CNR is the ratio of the carrier to the total noise power in 5 MHz bandwidth, excluding NLD.The formula used for CNR is the same as equation (4.2). Our approach is first to determine the maximum value of m which satisfy the CNLD requirements for a digitally modulated CATV trunk system. Then, by substituting the calculated value of m in Eq (5.1), the maximum allowable optical link budget is determined for a given CNDR.

Static NLD can be neglected compared to the dynamic NLD, both of which are determined by the same procedures for the AM-VSB system in chapter 4. Fig. (5.2) shows the numerically simulated CNLD for channel 99 (worst CNLD distortion), as a function of the OMD, m, per channel. As can be seen from Fig. (5.2), the CNLD for the worst-case channel is > 35 dB through the entire system as long as m $\leq$ 2%. The choice of 35 dB as a lower limit for the CNLD is believed to be a very conservative specification [1]. Accordingly, when considering the link CNDR at the receiver, a value of 2% will be taken as an upper limit for m. It should be pointed out that no clipping-induced distortion was observed in our simulation and that the total NLD was mainly due to signal - induced carrier density modulation.

For the link shown in Fig. (5.1), the CNDR at the receiver for a single channel is calculated in a similar way for the AM-VSB system in chapter 4.For the digital system the electronic bandwidth B_e is 5 MHz and RIN = -140 dB / Hz. All other parameters are the same as those used in Eq (4.2). Fig. (5.3) shows the simulated CNDR for the 100 channel 16 / 64 QAM system for the worst-case channel (channel # 99 with m = 2%) and limitations on CNDR due to various noise sources (5 MHz bandwidth), versus the total power budget. As can be seen from Fig. (5.3), the RIN is negligible, even if an FP laser with an RIN of -140 dB / Hz is used in the calculations. The signal-spontaneous beat noise is also unimportant due to the high input power coupled to the amplifier. The NLD (at high received power) and receiver thermal noise (at low received power) dominated. At a CNDR of 20 dB, as estimated to be required for 16 - QAM [1], one could transmit up to 400 MPEG2 video channels with an optical link budget of 34 dB. Alternatively, assuming a 20 Mb / s HDTV standard, either 100 HDTV channels or some combination of MPEG and HDTV could be accomodated. At a CNDR of 32 dB, as estimated to be required for 64 - QAM [1], one could transmit up to 600 MPEG2 video channels with an optical link budget of 27 dB.

It is interesting to compare the overall system performance of our proposed scheme with that of the more conventional scheme reported in [3-4] that uses either a directly modulated DFB laser or a 1.3-μm external modulation system and a TWSOA. In this conventional scheme, it was possible to transport 33-channel 64-QAM digital signals provided that the amplifier can be operated in the linear gain region ($P_{in} < -6.5$ dBm), that is as an in-line or preamplifier, but not as a power amplifier. This is because when P_{in} is high, the system performance degraded due to the increase of carrier-density-modulation induced second order nonlinear distortions which set the fundamental system limitation in delivering multiple channel 64-QAM signals. This is in contrast to our proposed scheme where operating the amplifier well into saturation (with high P_{in}), that is a power amplifier rather than an in-line or preamplifier, is necessary to satisfy both the CNR and distortion requirements for a digital CATV trunk system. It should be emphasized that the present analysis is applicable to short-distance links, where the high-power budget is used to achieve large splitting ratios for high network distribution capacity. Therefore, fiber nonlinearities, that may be important in long-distance systems, are not considered in this work.

5.3 Proposed M-QAM CATV Network Architectures

With a power budget of 34 dB, as obtained for the 16-QAM link, 128 subscribers can be served using a PON, Fig (5.4). For the FTTC system, 8192 subscribers can be accomodated, Fig (5.5).

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Chapter 6

Hybrid AM-VSB / M-QAM Compressed Digital Video CATV Distribution Network Using Travelling Wave Semiconductor Optical Amplifier as an External Modulator

6.1 Introduction

Today's cable television industry has the opportunity to utilize the 1 GHz bandwidth made available with systems based on fiber-trunk/coax-distribution architecture [1]. Current cable TV systems occupy the 50-500 MHz frequency band for 60-80 channel traditional AM-VSB services. To exploit frequencies up to 1 GHz cost-effectively requires more technical innovation, since the stringent requirement on AM-VSB channels limits the total modulation that can be applied to the modulator (direct or external), thus making it extremely difficult to further increase the channel capacity of a pure analog system using a single modulator. Given the technical limitations and the need for creating more service opportunities, delivering new digital services in addition to conventional analog services over cable networks then becomes very attractive and desirable [2].

This chapter proposes a hybrid AM-VSB/M-QAM CATV distribution network using 42 AM-VSB channels (300-546 MHz), ten compressed digital video M-QAM channels (600-654 MHz) and a travelling wave semiconductor optical amplifier as an external modulator at the transmission end of the system. Each channel, whether analog or digital, occupies a 6 MHz bandwidth. This system is compatible with current cable networks and promises delivery of new digital services using spectrally efficient multilevel RF techniques. In addition, because the digital signals are more robust than the analog signals with respect to noise and nonlinearities, the digital signal band can be de-emphasized to the level at which they contribute little to the overall load on the modulator and the whole system. Therefore, the system's channel capacity can be enhanced without increased cost of a second modulator and fiber, making one-modulator (laser or external modulator) broad-band fiber/coax or fiber-to-the-home networks very feasible and attractive..

6.2 Proposed Hybrid AM-VSB/ M-QAM Link Performance

The AM-VSB video subcarriers are the same as those used in chapter 4 i.e. 42 channels from 300-540 MHz, each with a 6- MHz bandwidth. Ten digital subcarriers from 600-654 MHz, each also occupying a 6 MHz bandwidth are also generated. For 16 / 64- QAM, each subcarrier can carry 4 / 6 compressed digital channels, giving a total of 40 / 60 channels. The entire system can therefore transport 82 / 102 video channels. Both AM and QAM signals are simulated by CW carriers. Biasing conditions are the same as those used in chapters 4 and 5 for the AM-VSB and M-QAM systems respectively i.e. P_{in} = 4 mW, and I_{bias} = 100 mA. Again, static nonlinearities can be neglected, and only dynamic nonlinearities are considered. Dynamic nonlinearity is determined using the same procedure as in chapters 4, 5. Fig. (6.1) shows the effect on AM channel # 41 (worst channel) as the modulation index per QAM channel is varied. In order to maintain the power budget derived in chapter 4, the modulation index per QAM channel must be ≤ 1%. Fig. (6.2) shows how the presence of the AM channels affect QAM channel # 9 (worst channel). To maintain CNLD of at least 35 dB, m per QAM channel must be at least 1%. Therefore, to determine the power budget for the QAM channels, m of 1% is used. Power budget for the M-QAM system (for channel # 9) is shown in Fig. (6.3).To determine CNDR versus power budget, the same formula is used as for the M-QAM system in chapter 5. All parameters used in the calculations remained the same.

6.3 Conclusion

Robust QAM signals do not change the distortion levels of the AM system. Therefore, the stringent requirements of cable industry on AM signal quality can be met and very good digital channel performance can be achieved by operating this system within limits imposed by noise and distortion. The system channel capacity has been greatly enhanced without increasing the cost from deploying a second laser and fiber.

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Chapter 7

Multichannel FM-FDM CATV Distribution Networks Using Semiconductor Optical Amplifier as an External Modulator

7.1 Introduction

Although SCM FM-FDM (frequency modulated frequency division multiplexed) transmission has been used in supertrunking since 1985, this market is relatively small compared to that of CATV trunking and distribution [1]. In the latter market, lightwave transmission products have already been deployed and are expected to become a major component in the 1990s. In trunk and distribution systems, because of the reduced capacity for cost sharing as compared to supertrunking, SCM-AM systems, with direct compatibility with the AM receivers built into the conventional TV set, are preferred over FM by CATV operators [2] i.e. FM signals, compared to AM signals, require processing between the optical receiver and the appliance [3]. To be rendered compatible with conventional televisions, an FM signal must be downconverted in frequency, demodulated (FM-to-AM conversion), and remixed into the appropriate frequency band expected by the video appliance. The required mixers, oscillators, and FM demodulators are all available in highly integrated form, but nevertheless represents an additional cost compared to the case of

AM. Since the FM channels must be demodulated individually, the cost increment is multiplied by the number of simultaneous channels to be accessed at the receiving end. FM signals exchanges a wide bandwidth for a considerable signal-to-noise ratio (SNR) improvement. The SNR at the output of an FM detector is much larger than the carrier-to-noise ratio (CNR) at the detector input, provided that the CNR is above a threshold value that is a characteristic of the detector. As a result, FM systems have the potential to become upgrades for current NTSC video transport. An approximate value for the required bandwidth B is given by Carson's rule, $B = \Delta f_{pp} + 2f_m$, where Δf_{pp} is the peak-to-peak frequency deviation of the modulator, and f_m is the top audio subcarrier frequency The corresponding output SNR improvement over the input CNR is given by:

$$(S/N)_o = (C/N)_i + 10.\log[3/2.B/f_o.(\frac{\Delta f_{pp}}{f_v})^2] + w \qquad (7.1)$$

where f_v is the top video bandwidth, and w is a weighting factor. The SNR in (7.1) can be further improved by preemphasis. The total improvement value varies among different systems, but normally in the range 36-44 dB. For example, for a 525- line NTSC C-band downlink satellite signal, $f_m = 6.8$ MHz,

f_v = 4.2 MHz, Δf_{pp} = 22.5 MHz, B = 36.1 MHz, w = 13.8 dB, and the total SNR improvement over CNR, without taking preemphasis into account is 39.5 dB. Therefore, CNR of 16.5 dB is large enough to reach a studio quality SNR of 56 dB.

Current CATV systems normally provide video SNRs on the order of 40 to 45 dB, and studio quality NTSC video and High Definition Television (HDTV) typically have SNRs of 56 dB or higher. Therefore, AM transmission of NTSC video will require a received CNR in excess of 40 dB, while FM-based transmission can offer higher picture quality with a CNR of just 16 dB.

The FM advantage translate to clear benefits in power budget on the fiber link. Higher video SNR and/or a larger menu of channels are two possibilities. The additional power budget could allow the output power of the modulator to be split and shared among a large number of subscribers, decreasing the per-subscriber cost of the Central Office (CO) or head-end transmitter. The additional budget also afford greater system robustness to noise degradations.

7.2 Proposed FM-FDM Link Performance

At P_{in} = 4 mW and I_{bias} = 100 mA, the bandwidth of the TWSOA is about 2.7 GHz. This is insufficient to accomodate the FM signals, each of which requires a bandwidth of about 30 MHz. To overcome this problem, the bias current was shifted from 100 mA to 140 mA. Even though static and dynamic nonlinearities are now increased, because FM requires CNR of about 16.5 dB, good quality signals can still be delivered.

We propose a 120 channel FM-FDM CATV link as shown in Fig. (7.1). The channels are from 60 MHz-2.82 GHz, spaced 40 MHz apart, each occupying a bandwidth of 30 MHz. To minimize the disturbance due to second order intermodulation distortion (IMD), the frequency of the FM carriers are set to half the frequency interval W plus the integer multiple of W. The frequency of the ith FM signal f_i is [4] :

$$f_i = \frac{W}{2}(2i - 1) \quad (i = l, l+1, \ldots\ldots h) \qquad (7.2)$$

With this frequency plan, the second order IMDs fall in the band between the carriers. As for the previous systems considered thus far, CW carriers are used. The numerical simulation of the static and dynamic NLD is done in a manner similar to that of the AM-VSB system in chapter 4. Again, static NLD does not present a problem. The dynamic C / NLD for channel 119 (worst channel) is shown in Fig. (7.2), as a function of OMD, m, per channel. As can be seen, the C/ NLD for the worst channel is > 28 dB throughout the entire system as long as m ≤ 4%.The choice of 28 dB is considered to be a safe value to use [6]. For this system no clipping induced distortion was observed. To calculate the CNR versus power budget, formula (4.2) was used. In this case, B_e is 30 MHz, G is 10.97 dB and RIN = -140 dB/Hz. Fig. (7.3) shows the simulated CNR for the 120 FM-FDM system for the worst case channel (channel # 119, m = 4%), versus the total power budget. For a CNR of 16.5 dB at the receiver, an optical link budget of 37 dB is obtained.

7.3 Proposed FM-FDM CATV Network Architecture

For FM-FDM, only an FTTH system is considered. This is because FM-FDM needs a lot of bandwidth, which means much more expensive coaxial cables would have to be deployed. Using a PON, and the power budget of 37 dB found in section (7.2), 256 subscribers can be served, Fig.(7.4)

7 .4 References

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Chapter 8

Summary

This thesis addressed both the technical and economic performance issues of several novel local distribution fiber/coaxial and fiber-to-the-home CATV network architectures that are capable of supporting low-cost broad-band services including analog/digital entertainment video, and at the same time meet the demand of achieving a high optical power budget, required for deeper fiber penetration. Specifically, this work proposed and analyzed several novel local distribution CATV network architectures that utilized a 1.3 μm Semiconductor Optical Amplifier (SOA) based transmitter as well as a combination of fiber and coaxial cable as an upgrade to narrowband fiber-to-the-curb (FTTC) network architectures. Fiber-to-the-home (FTTC) architectures were also investigated. The main characteristic of the proposed network architectures was the innovation of using a 1.3 μm SOA as an external modulator at the transmitter end of the proposed network architectures. This represented a major milestone in achieving the above mentioned objectives since, as was shown, in addition to performing the modulation function, the required high optical power budget was obtained through 1.3 μm amplification and consequently there is no need to dismantle the already deployed 1.3 μm zero-dispersion single mode fibers.

The overall objective of this work was to investigate and analyze, through computer simulation and modeling, the performance of all the critical elements necessary for the implementation of high-capacity, high-performance, cost-effective local loop distribution CATV networks based on subcarrier-multiplexed fiber communications technology. We implemented a flexible, powerful computer modeling tool for evaluating the end to end performance of the proposed local distribution CATV network architectures. The model allowed several candidate modulation formats: Amplitude Modulation (AM), Frequency Modulation (FM) and Quadrature Amplitude Modulation (QAM), to be compared, identified the most important sources of performance degradation, and allowed overall estimation of the proposed networks' capacities. These simulation models were interconnected into a system level simulation model of the proposed subcarrier-multiplexed architectures. We demonstrated some of the capabilities of the simulation model through a number of examples at both device and system level.

Following is a summary of the main results of this thesis :

- Characterization of the TWSOA in terms of its static nonlinearity, resulting from its P - I characteristic curves, and its dynamic nonlinearity, due to carrier density modulation.
- Determination of the frequency response characteristics of the TWSOA, showing the effects of both optical input power and DC bias on its bandwidth.
- To achieve acceptable linearity, and provide high output power, the TWSOA has to be operated deep into saturation. For our simulations, P_{in} = 4 mW, and I_{bias} is at least 100 mA.
- Using AM-VSB modulation, we were able to transmit 42 channels from 300-546 MHz, at 3% modulation index per channel, and obtain a 17 dB power budget for CNR of 50 dB and CTB of -60 dBc.
- For QAM modulation, using 16 / 64 QAM, 100 subcarriers from 139.25 - 733.25 MHz were conveyed. Each channel had a modulation index of 2%, and a power budget of 34 / 27 dB was achieved while maintaining a CNDR ratio of at least 20 / 32 dB. A CNLD ratio of 35 dB was conservatively estimated. With current compressed digital video standards, 400 / 600 digital channels can be transported in this system.
- A hybrid AM-VSB and M-QAM hybrid system was also demonstrated.

We used the same AM-VSB scheme discussed above in combination with ten digital carriers from 600-654 MHz. With 16/64- QAM, a total of 40/ 60 compressed digital channels can be delivered. The entire system provided 82/102 video channels to choose from.

- A 120 channels FM-FDM system from 60 MHz - 2.82 GHz, having a modulation index of 4% per channel, CNR of 16.5 dB (SNR of 56 dB), and a power budget of 37 dB was simulated.
- Using the above results for AM, QAM, and FM systems, various design examples, employing the PON architecture and FTTH design, were illustrated..
- FTTC systems, were implemented for the AM-VSB and 16-QAM formats.

Eventually, we were able to model the end-to-end performance of any signal, irrespective of modulation scheme, that passed through the network. This permitted us to compare the feasibility of different modulation formats in both FTTC and FTTH systems. In addition, this has enabled us to determine how to use the capabilities of each architecture to best advantage and provide insight into which subcarrier multiplexed architecture should be stressed by providing estimates of comparative networking costs.

BASIC SCM SYSTEM CONFIGURATION ANALOG AND/OR DIGITAL SIGNALS

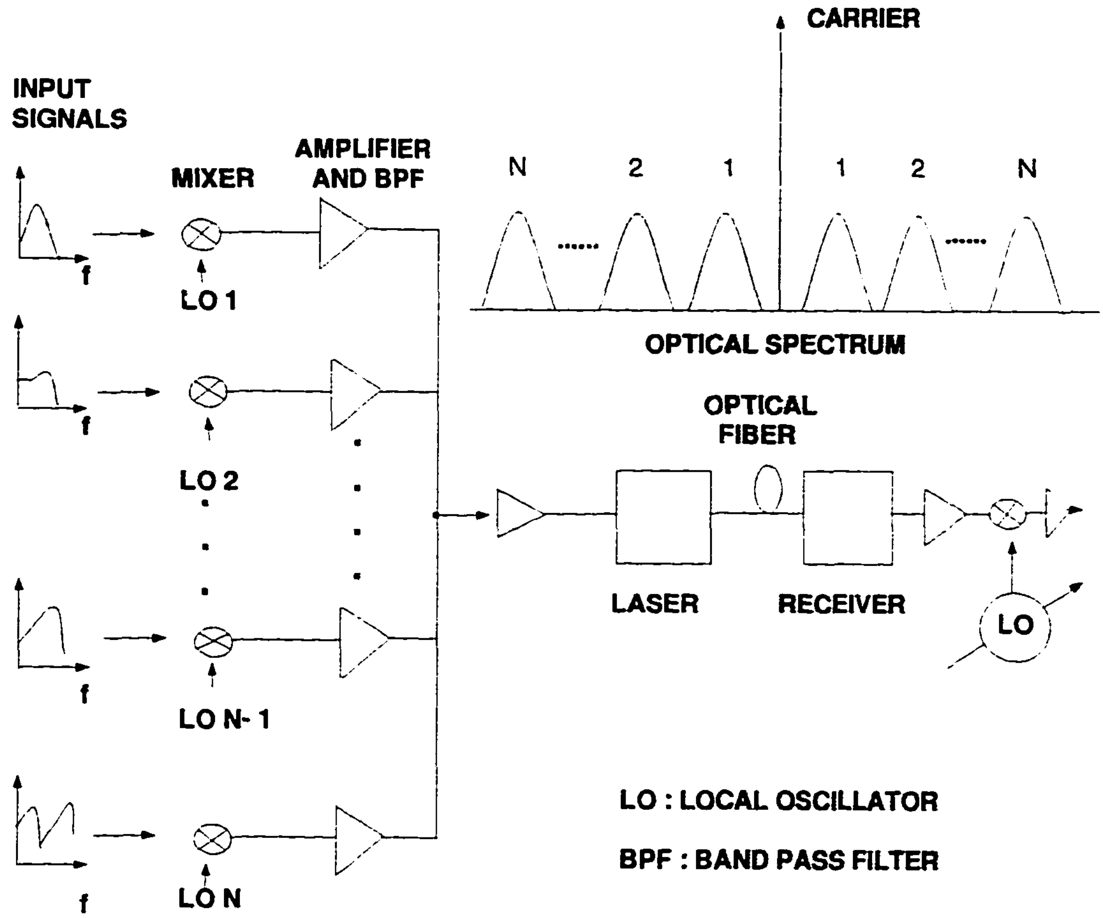

Figure 2.1

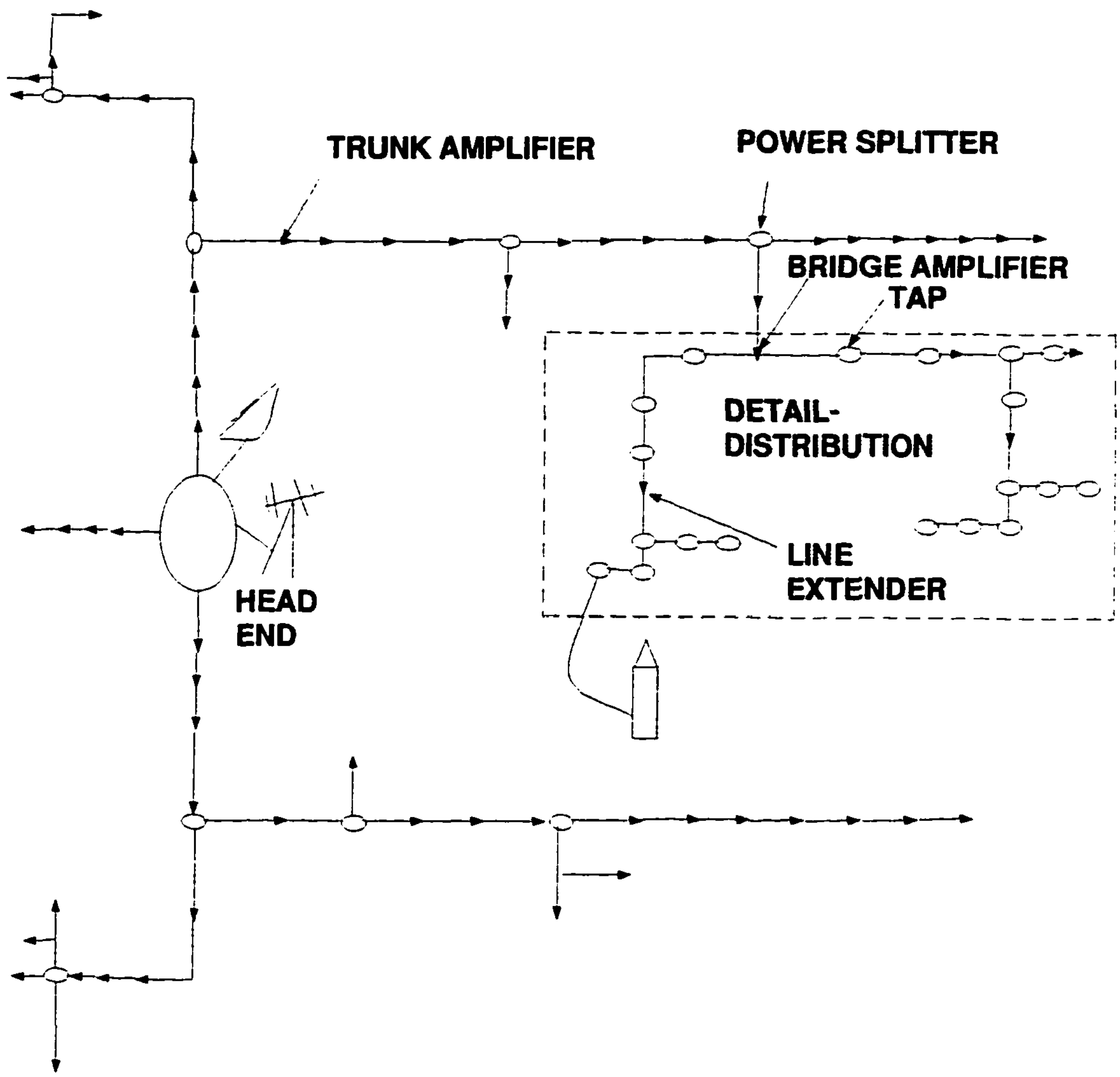

Figure 2.2 TYPICAL CABLE SYSTEM

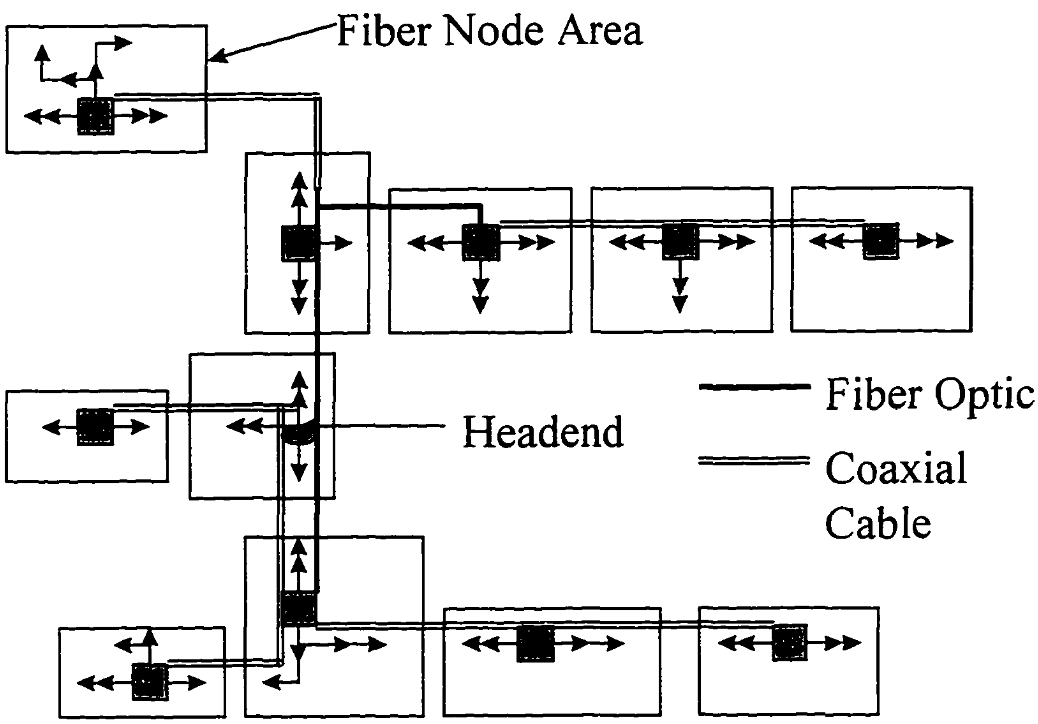

Figure 2.3 Cable System with backbone trunk

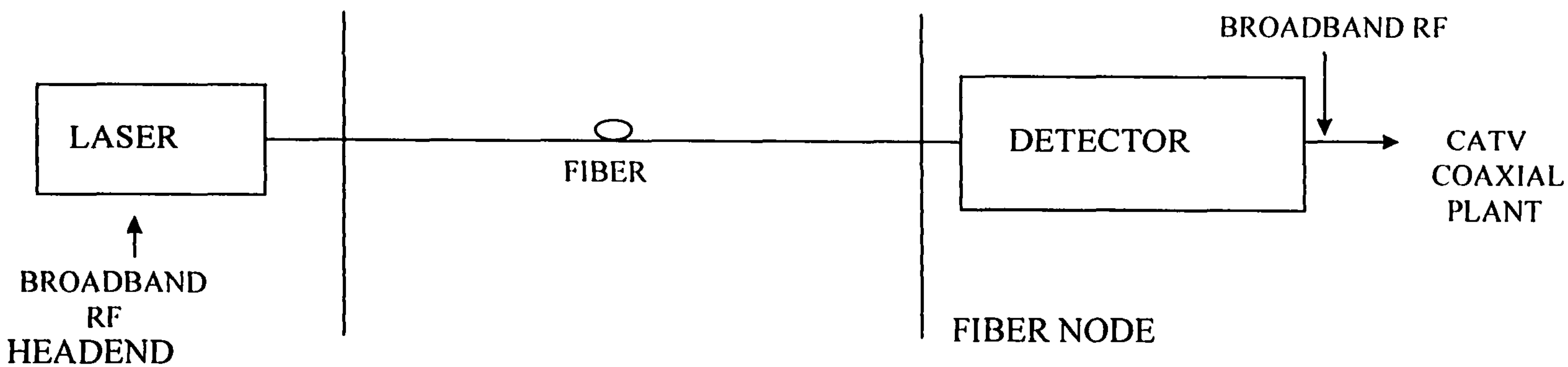

FIG 2.4 Broadband AM on fiber

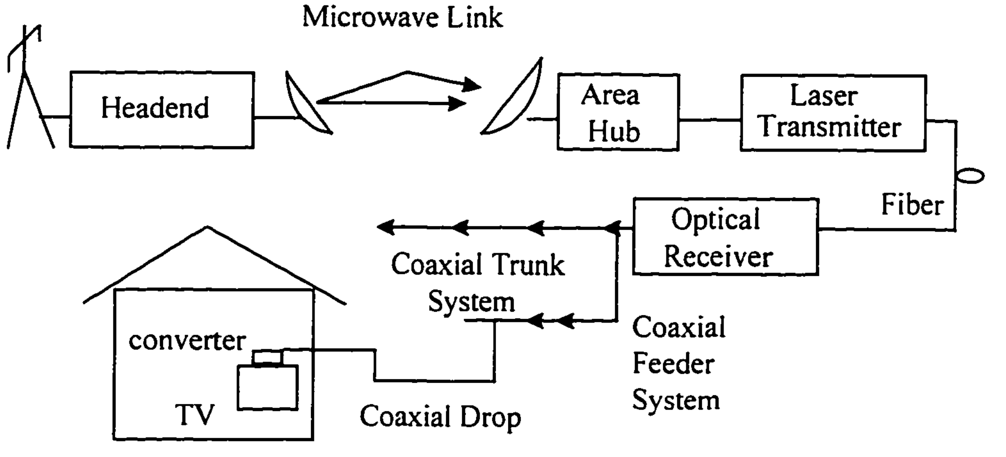

Fig. 2.5 CATV System Elements

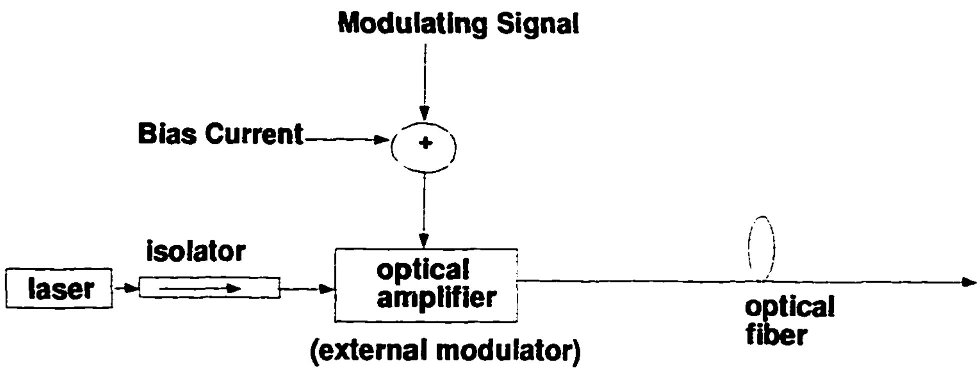

Figure 3.1 CATV system using optical amplifier as external modulator

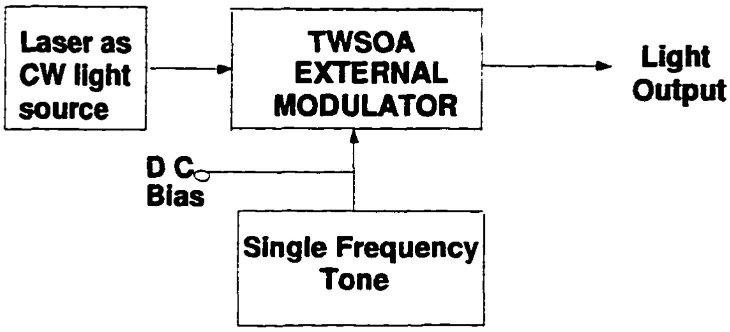

Figure 3.2 System used to find Frequency Response of the TWSOA

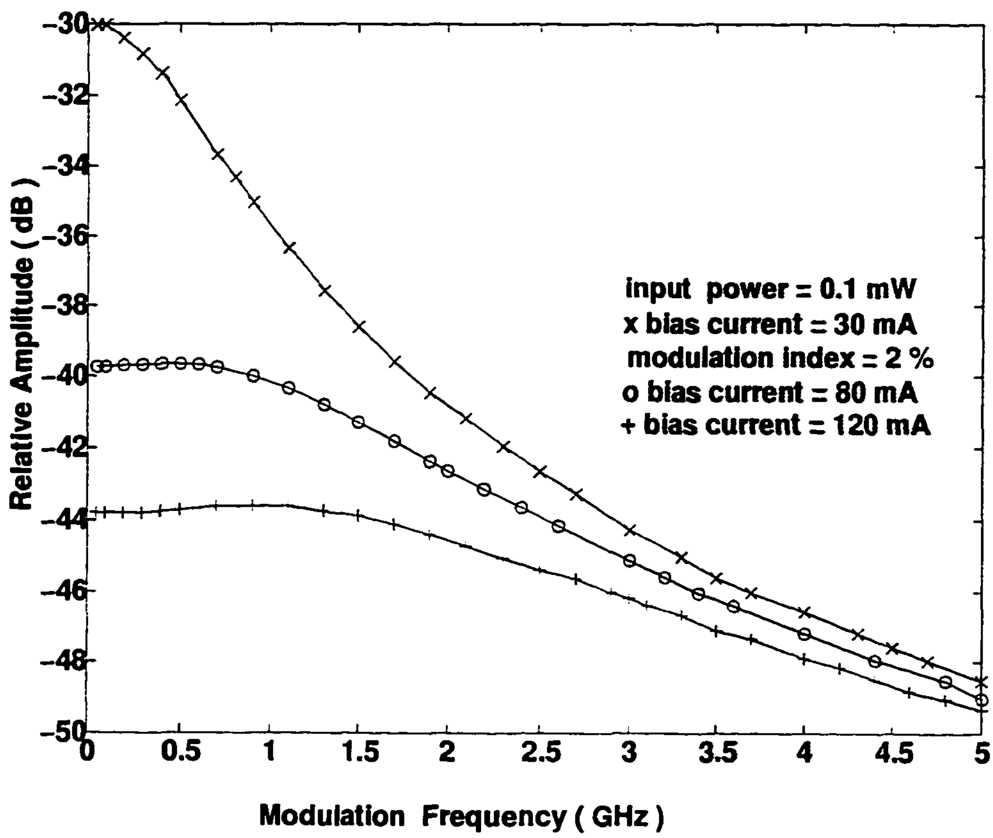

Figure 3.3 Frequency Response Curve

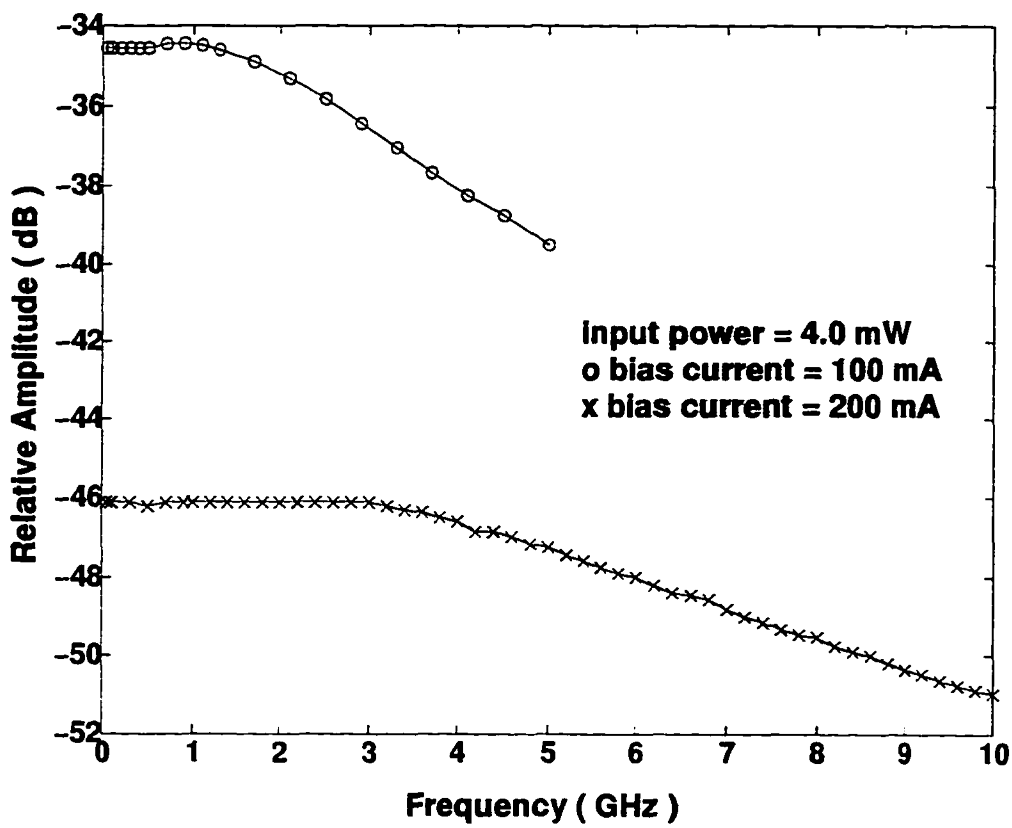

Figure 3.4 Frequency Response curve

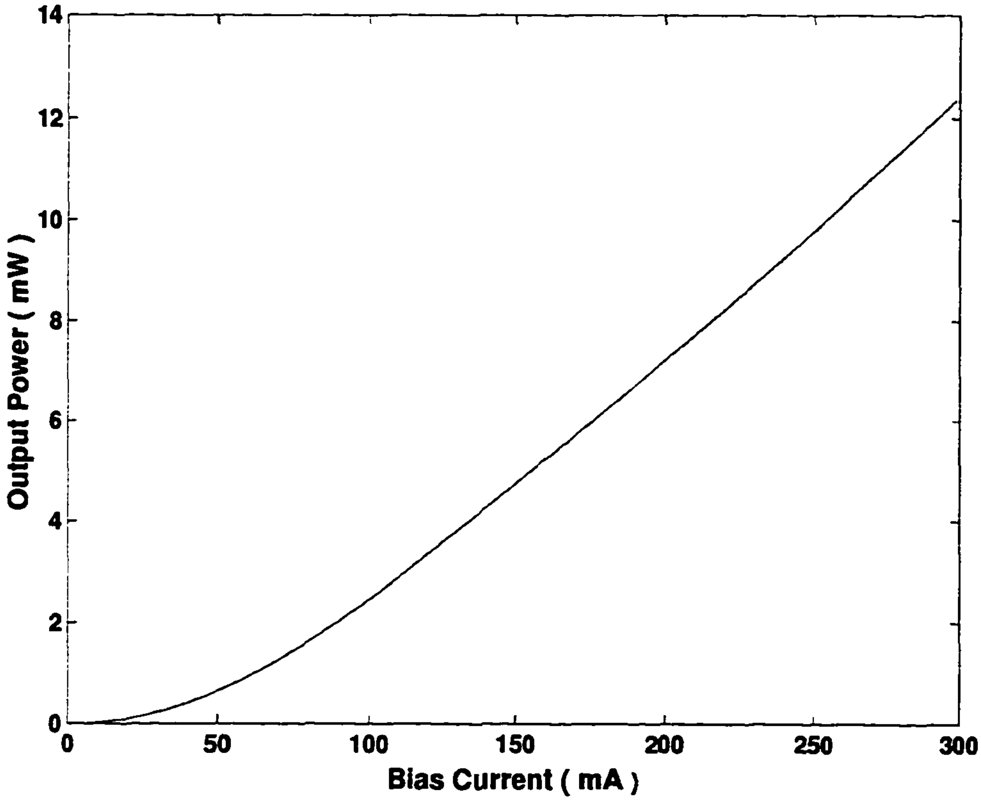

Figure 3.5 Output Power versus Bias Current for input power = 0.01 mW

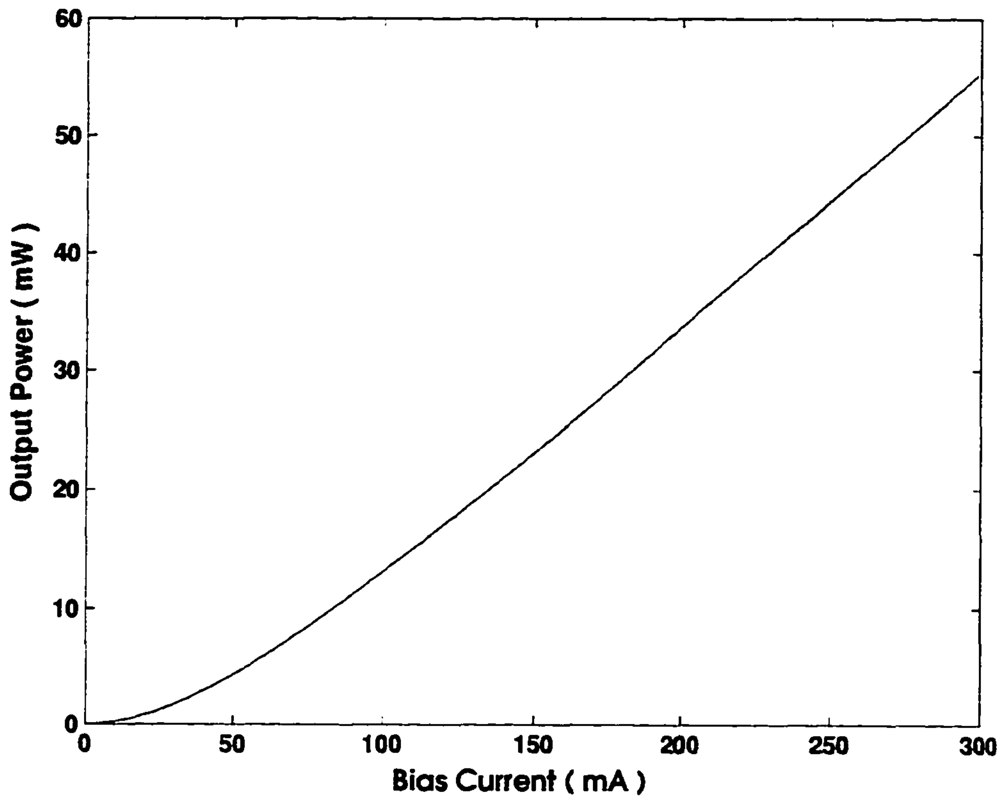

Figure 3.6 Output Power versus Bias Current for input power = 0.1 mW

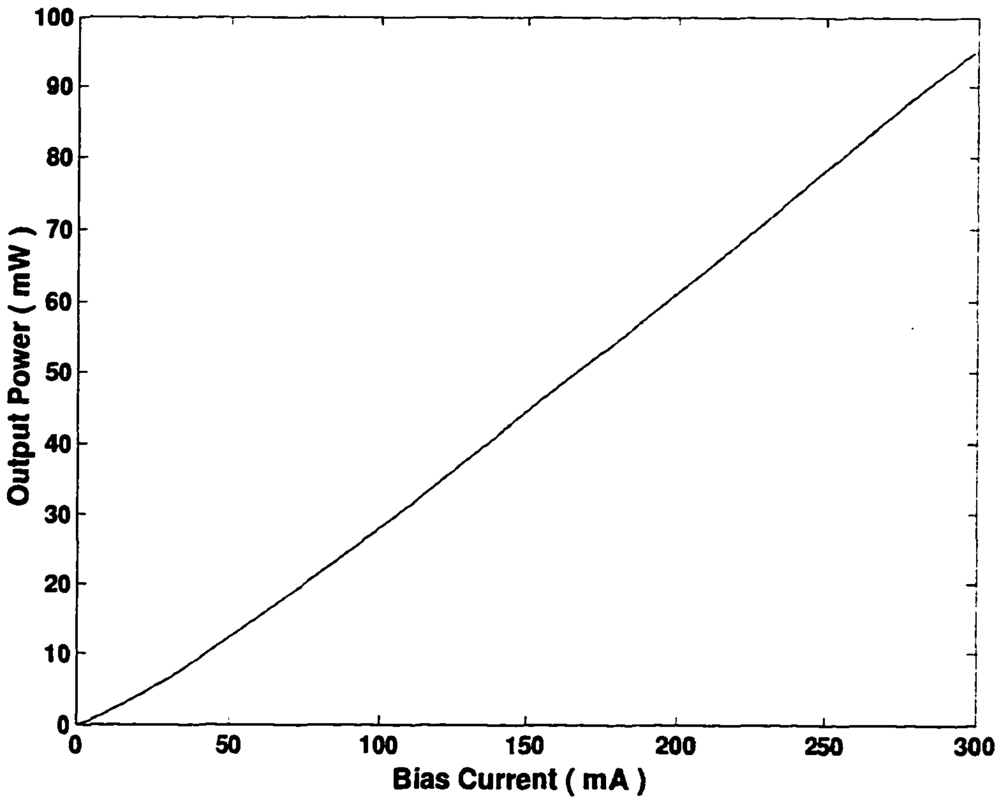

Figure 3.7 Output Power versus Bias Current for input power = 1.0 mW

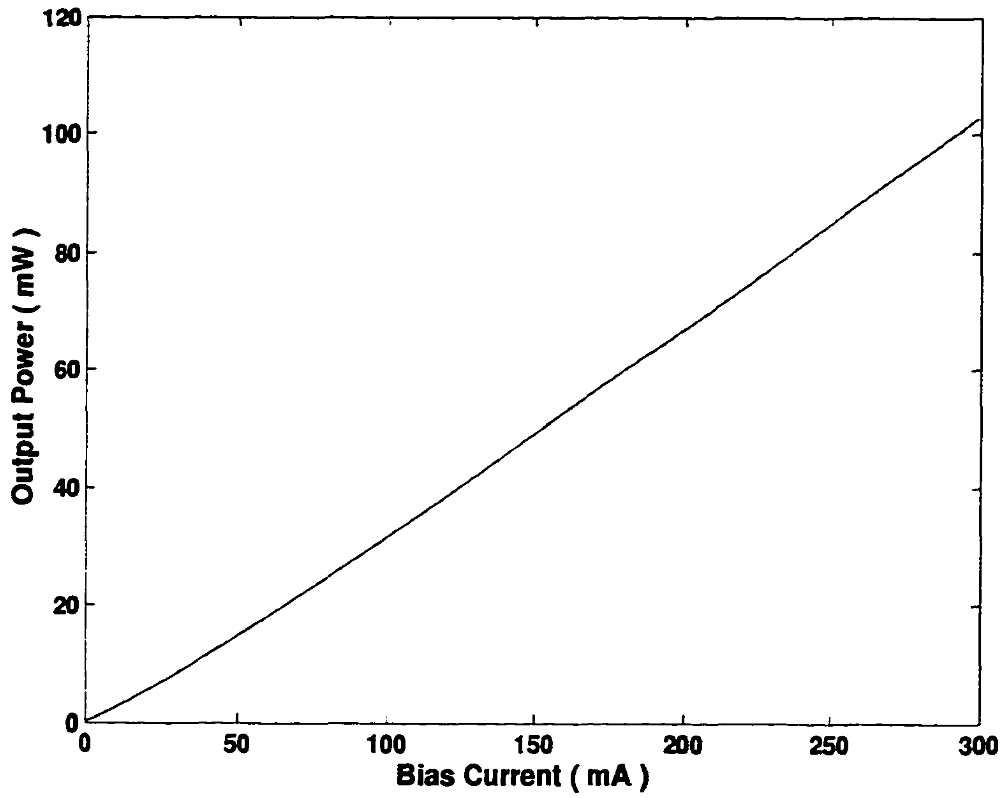

Figure 3.8 Output Power versus Bias Current for input power = 2.0 mW

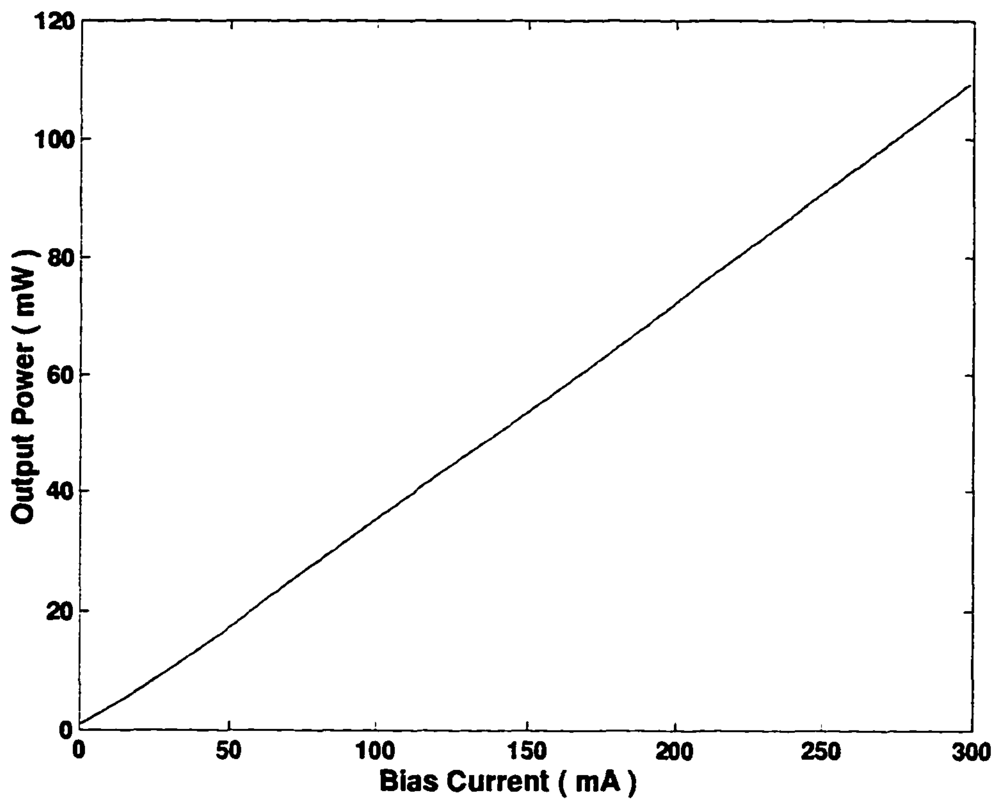

Figure 3.9 Output Power versus Bias Current for input power = 4.0 mW

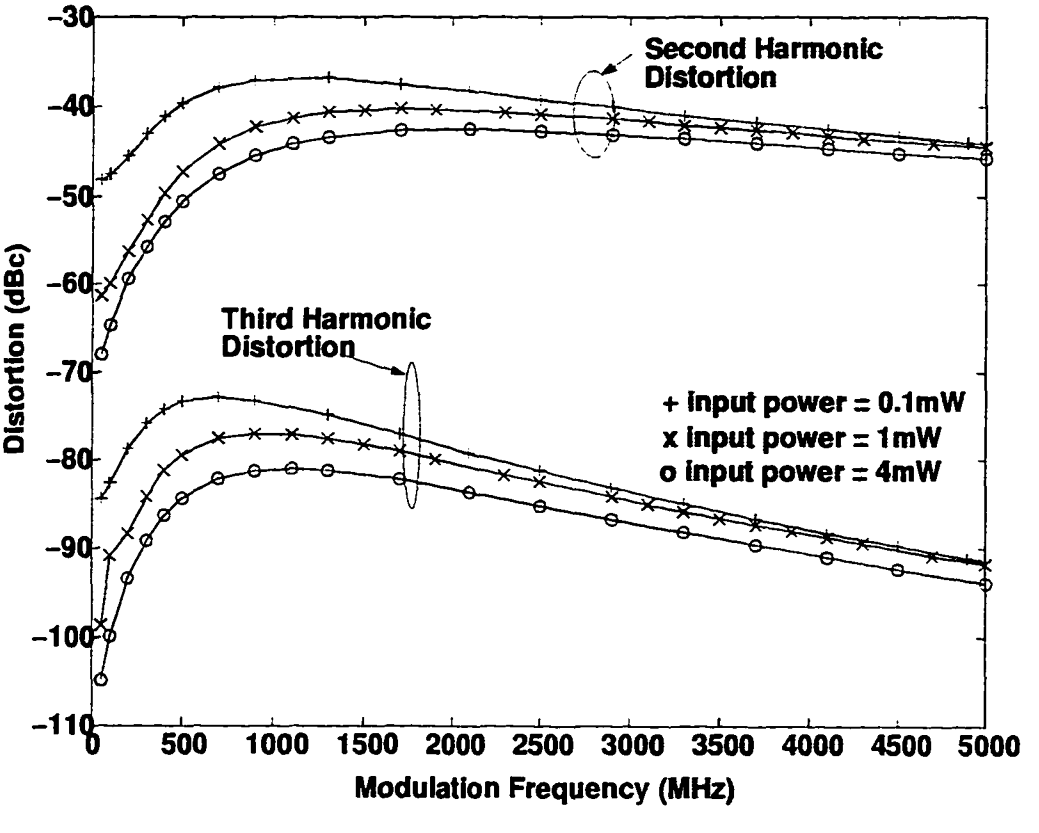

Figure 3.10 Second and third order harmonic distortion versus modulating frequency for various values of input power

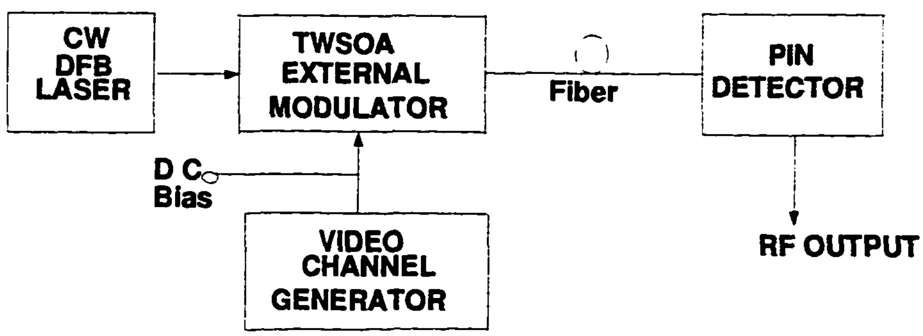

Figure 4.1 Block diagram of the AM-VSB system

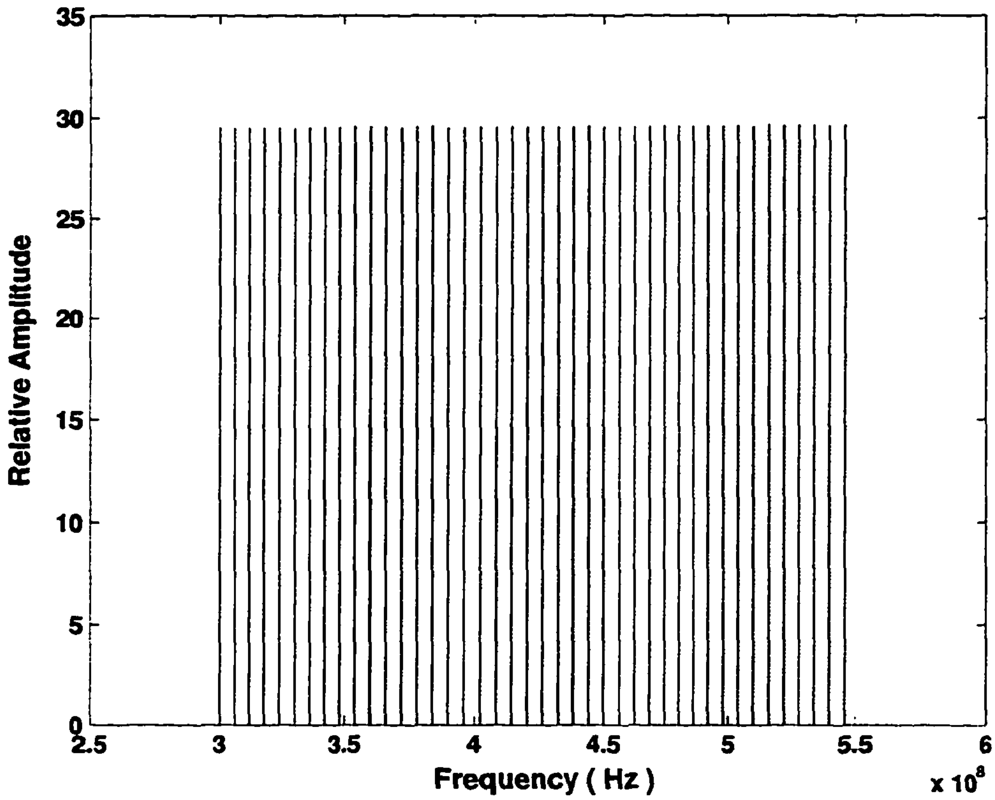

Figure 4.2 42- channels AM - VSB system from 300 - 546 MHz

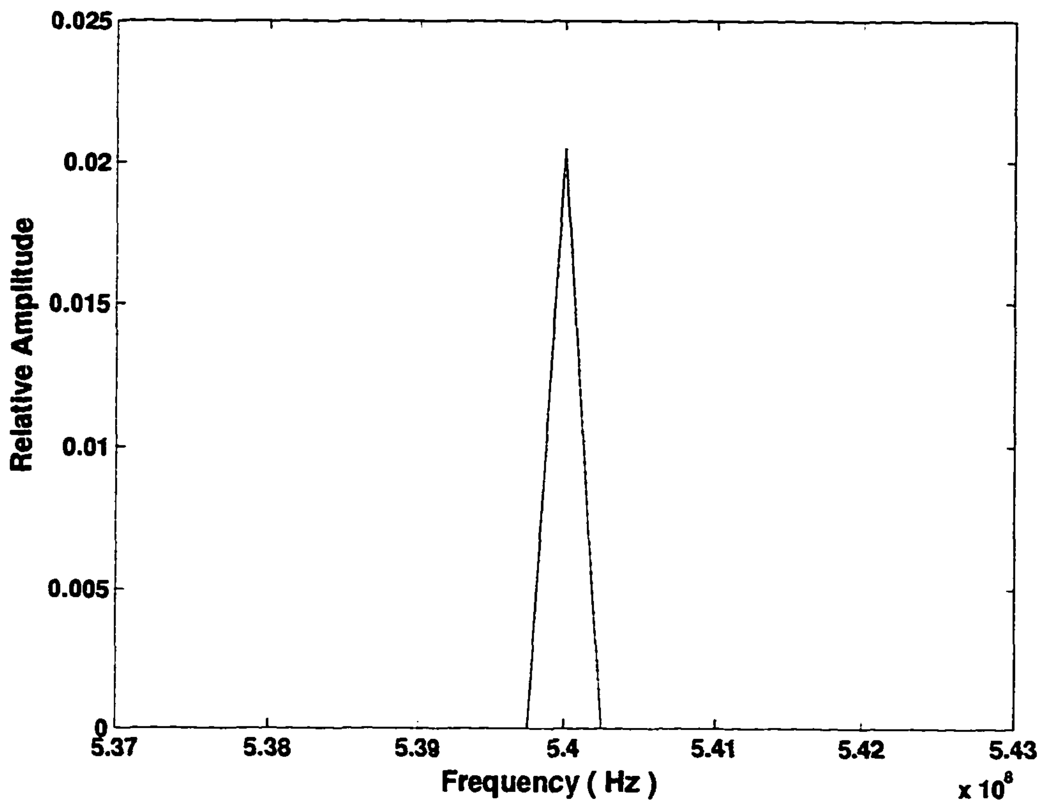

Figure 4.3 Distortion in channel number 41 when the carrier is suppressed

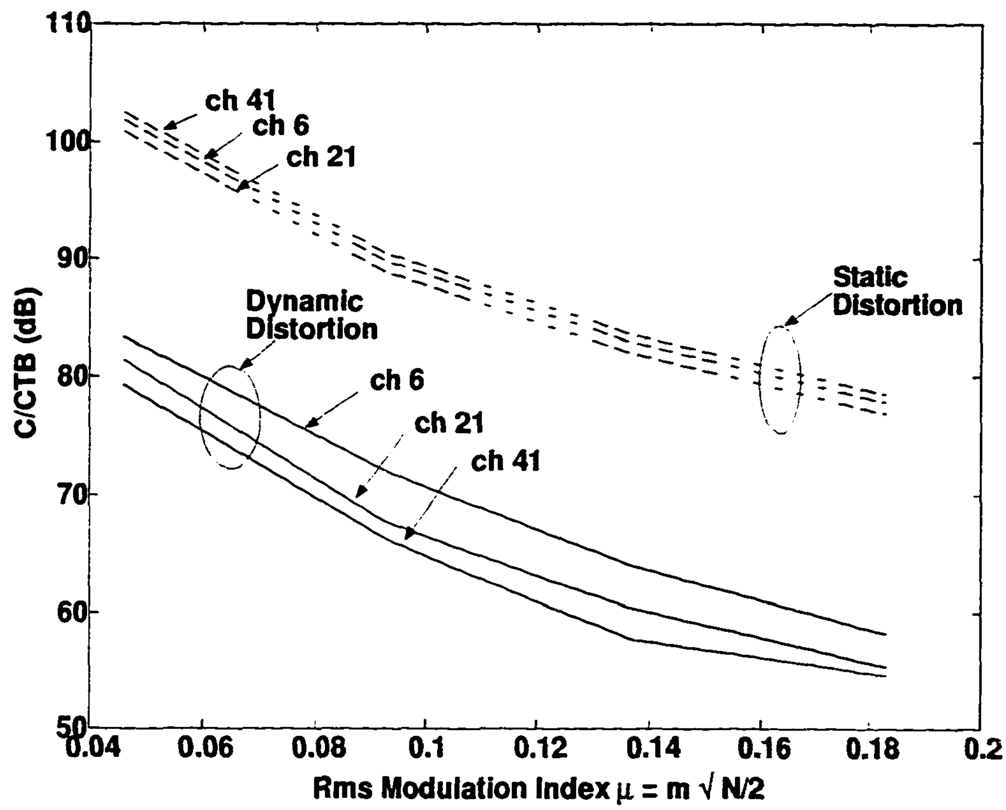

Figure 4.4 Numerically simulated CTB distortion for channels 6, 21, and 41.

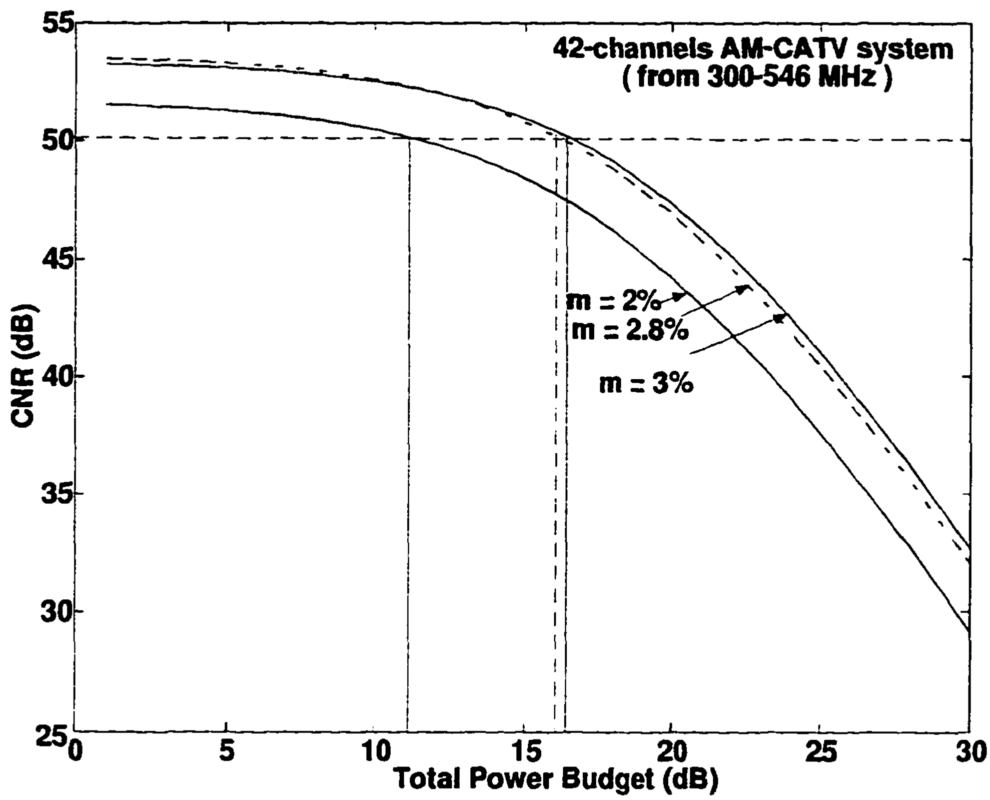

Figure 4.5 CNR versus total power budget for m = 2%, 2.8%, 3%.

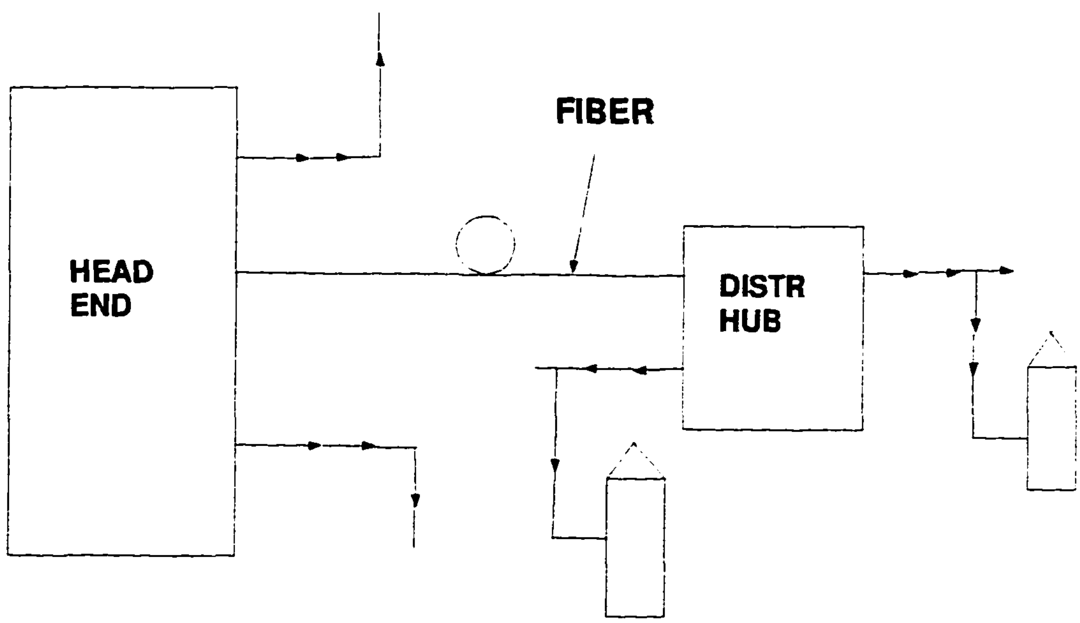

Figure 4.6 Fiber To The Curb Schematic

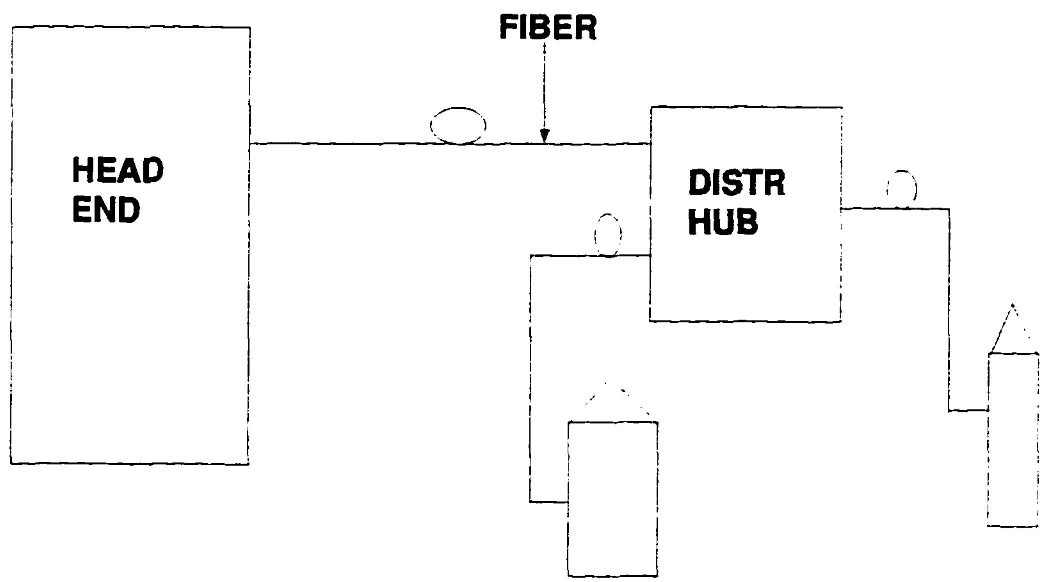

Figure 4.7 FIBER EXTENDED CLOSE TO,OR ULTIMATELY ALL THE WAY TO THE HOME (FTTH).

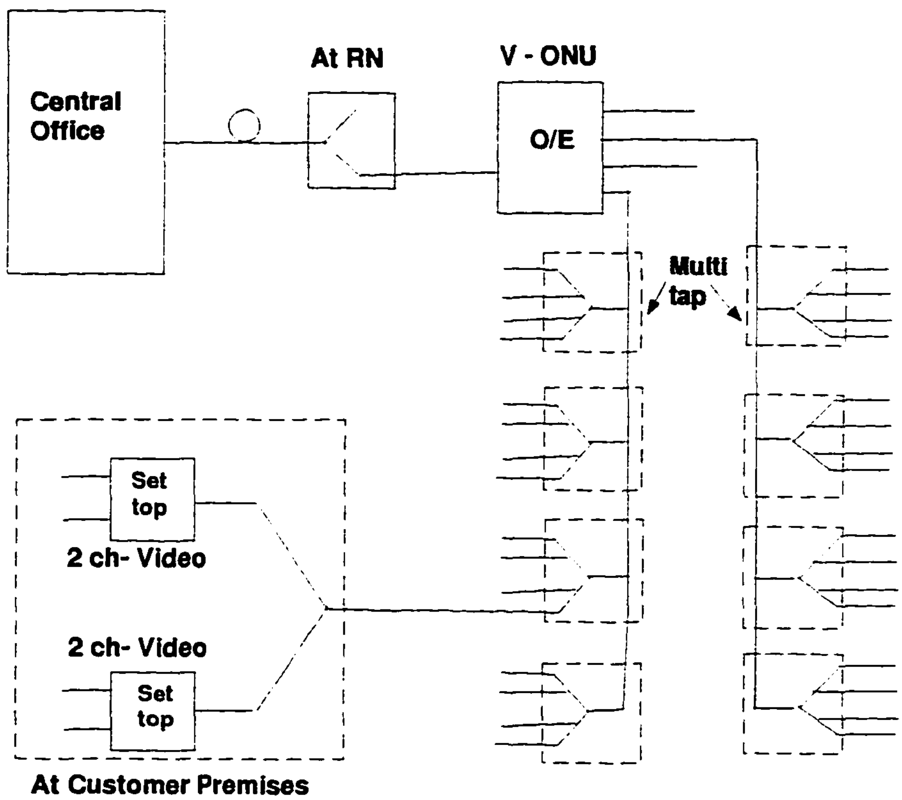

Figure 4.8 Fiber To The Curb System

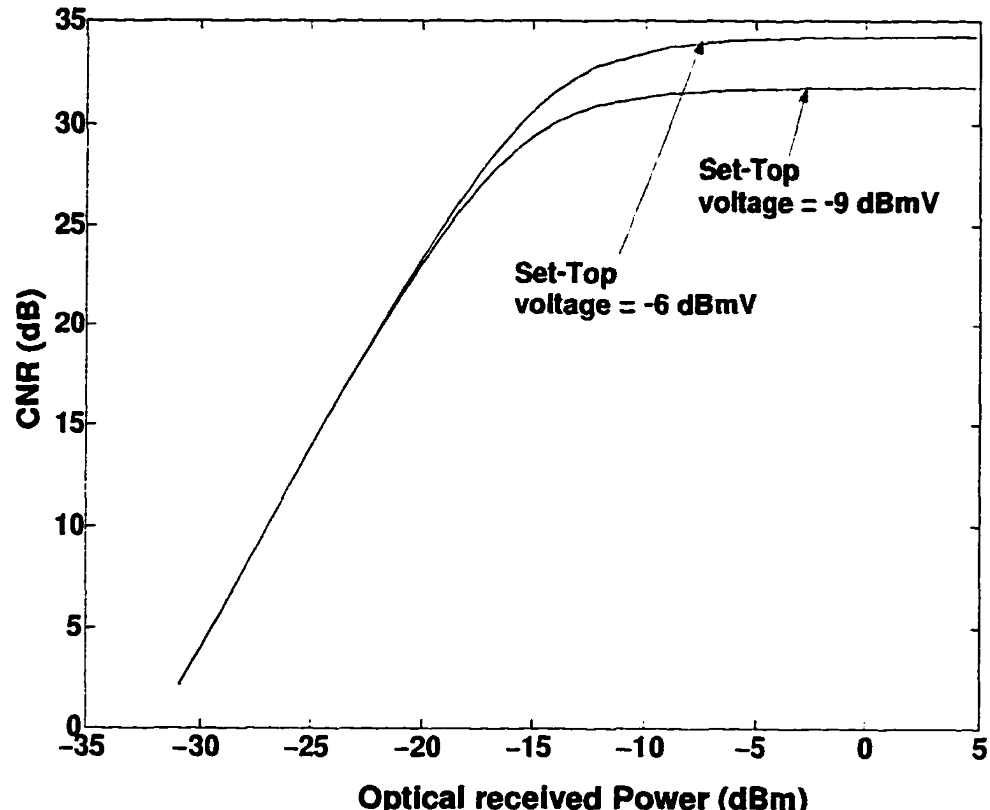

Figure 4.9 Typical curves of CNR vs. received optical power as a function of set-top signal level

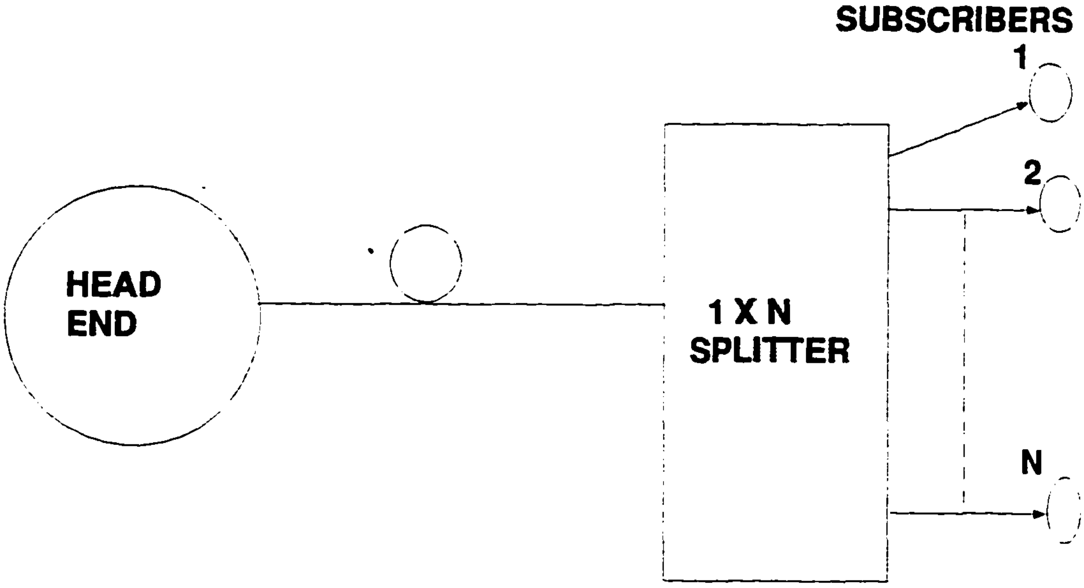

Figure 4.10

PASSIVE OPTICAL NETWORK SERVING N SUBSCRIBERS OVER A SINGLE FEED FROM THE HEADEND.

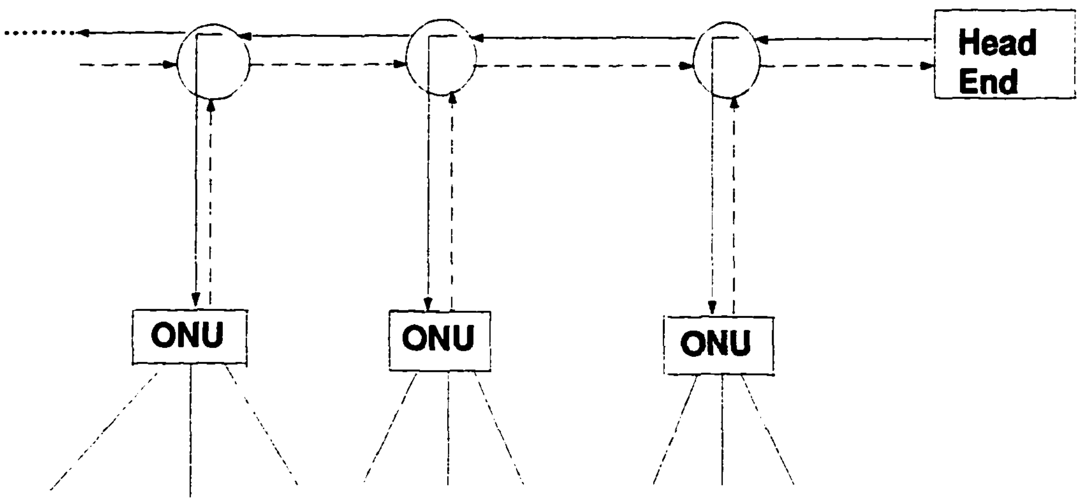

Figure 4.11 BUS TYPE FTTH OPTICAL NETWORK

ONU: optical network unit

: optical splitter

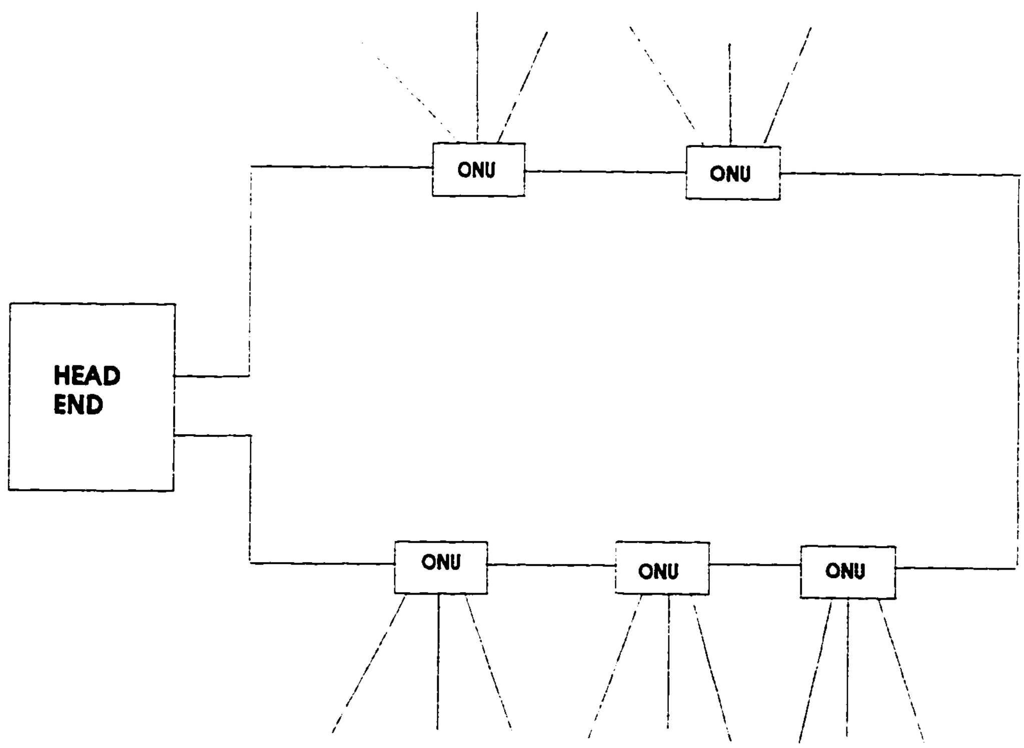

Figure 4.12 RING TYPE DISTRIBUTION SYSTEM

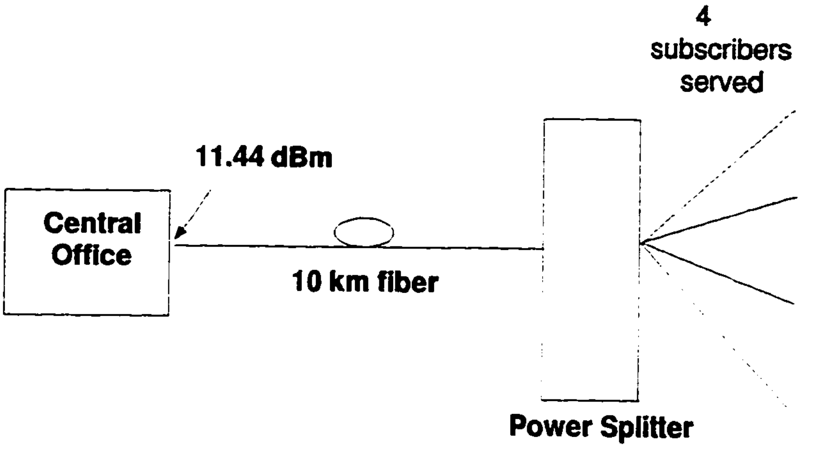

Figure 4.13 FTTH AM - VSB System

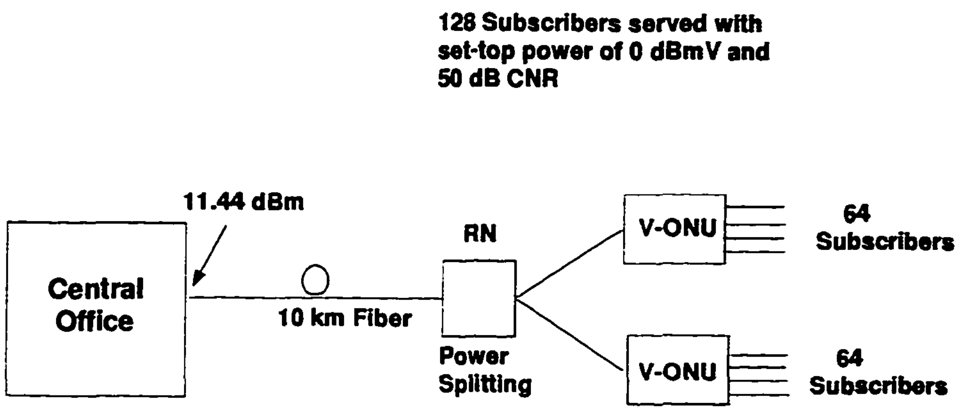

Figure 4.14 FTTC AM - VSB System

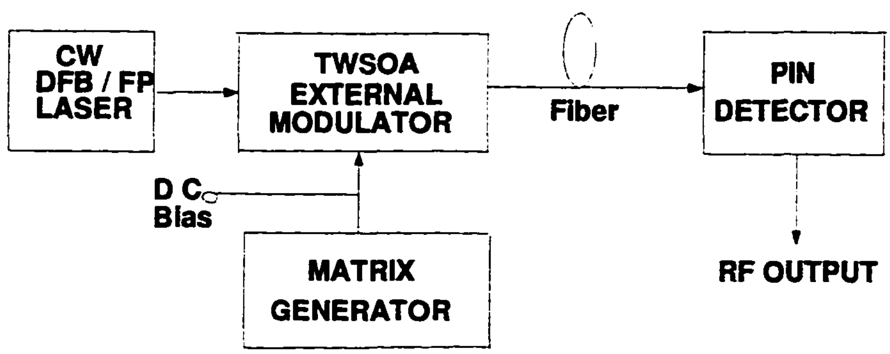

Figure 5.1 Simulation block diagram for M-QAM system

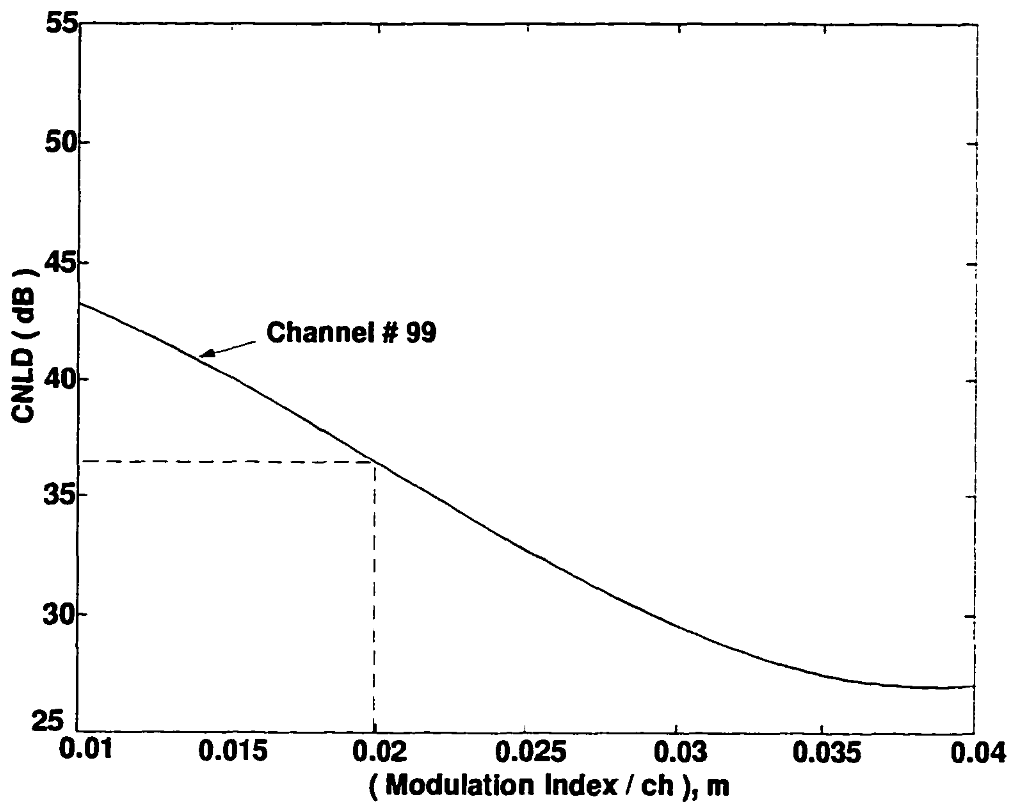

Figure 5.2 CNLD versus m for worst-case channel in a 100 channel system.

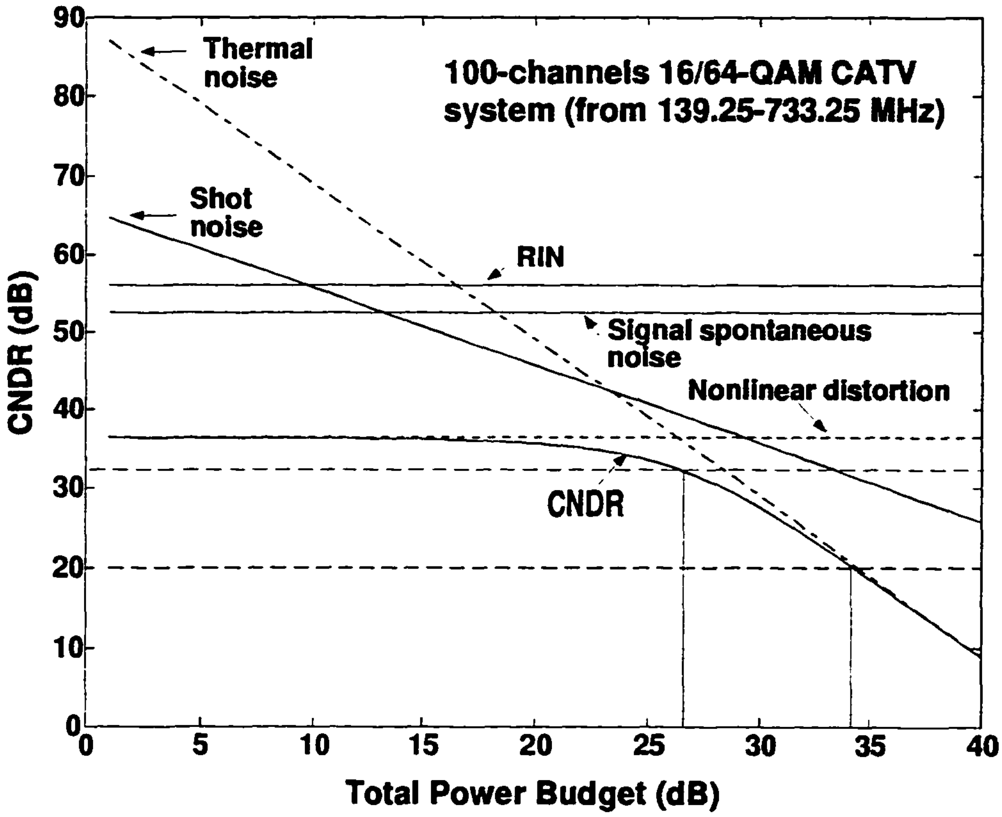

Figure 5.3 CNDR versus total power budget for m = 2%

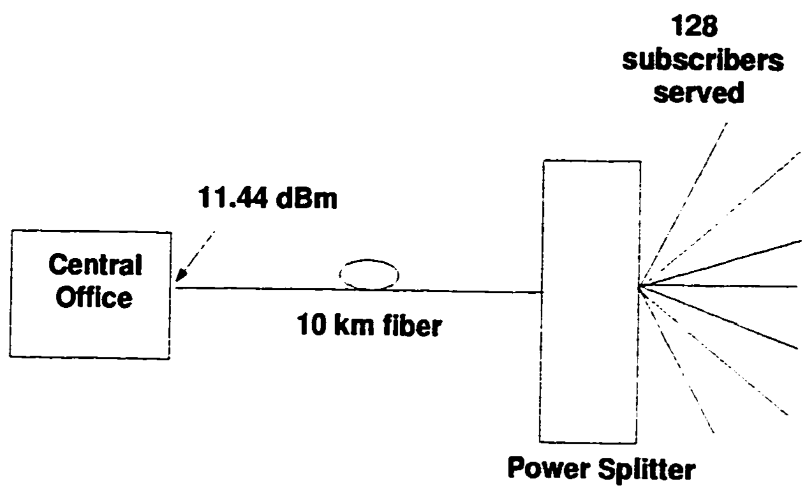

Figure 5.4 FTTH 16- QAM System

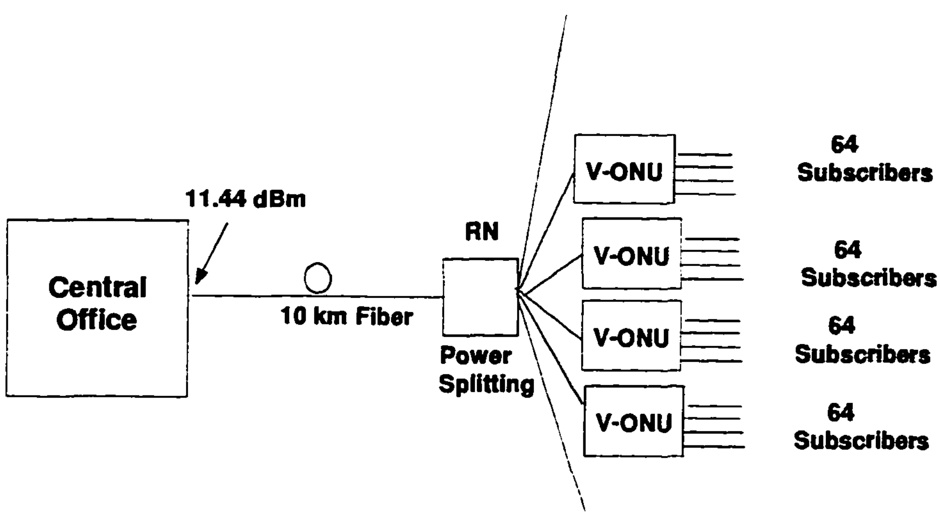

Figure 5.5 FTTC 16- QAM System

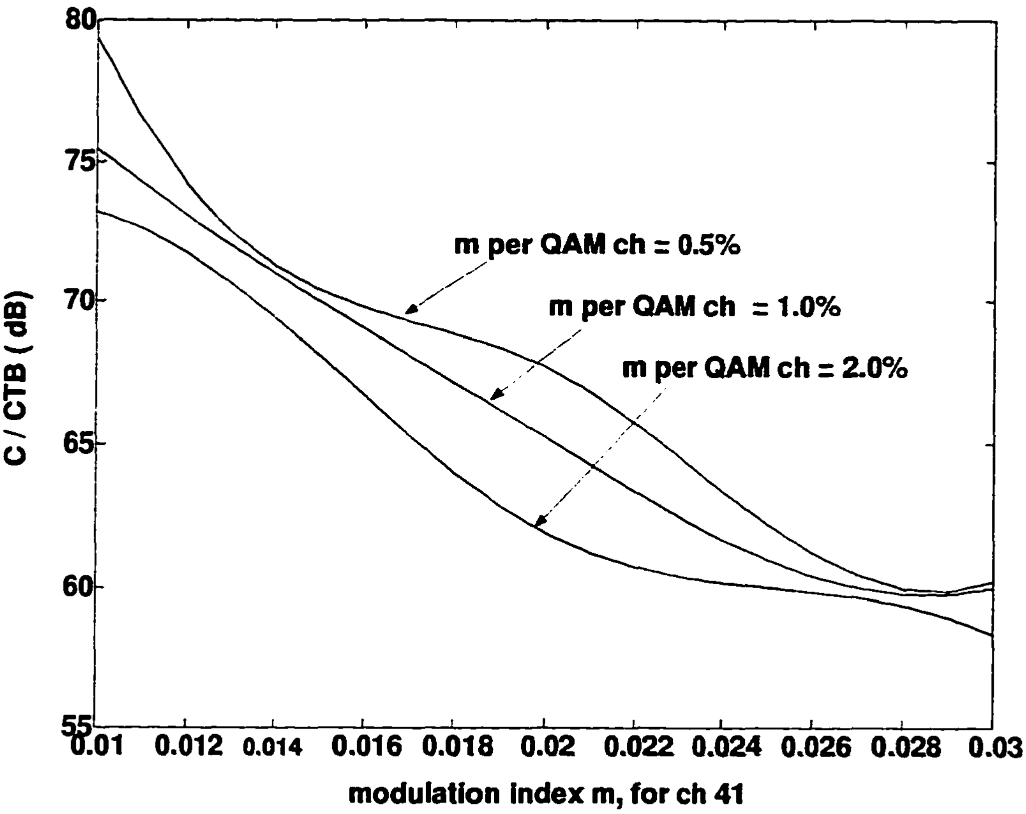

Figure 6.1 Effect on AM channel # 41 due to QAM channels

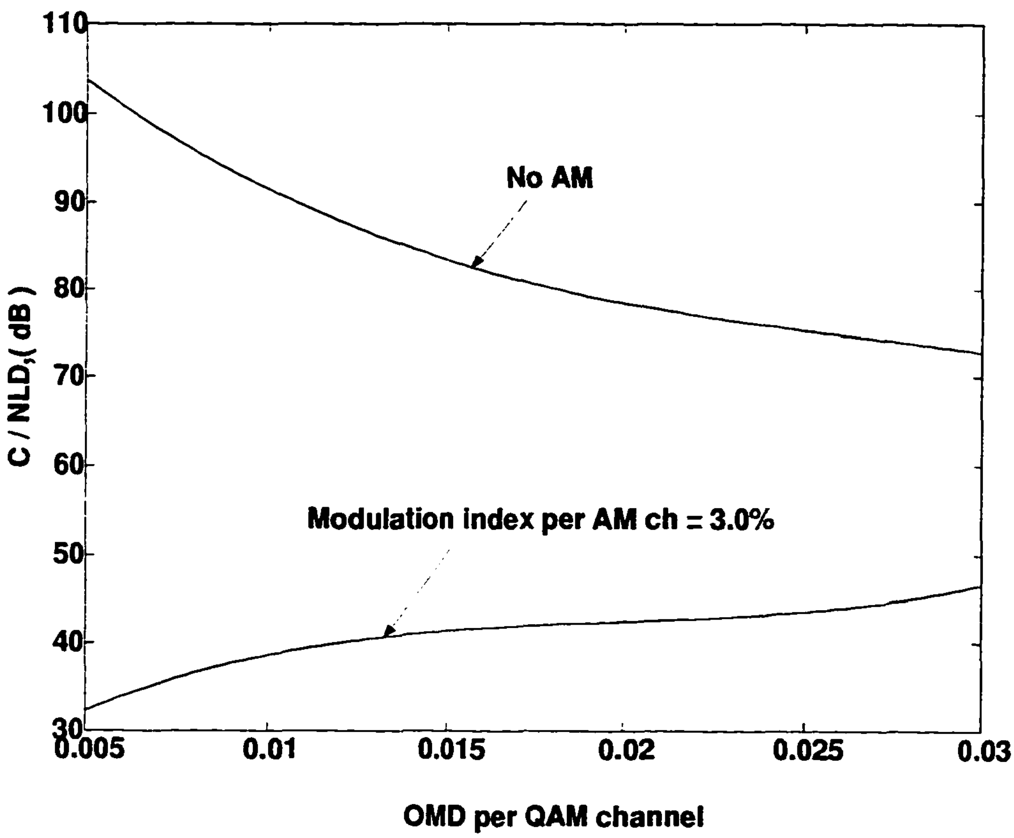

Figure 6.2 Effect on QAM channel # 9 due to AM channels

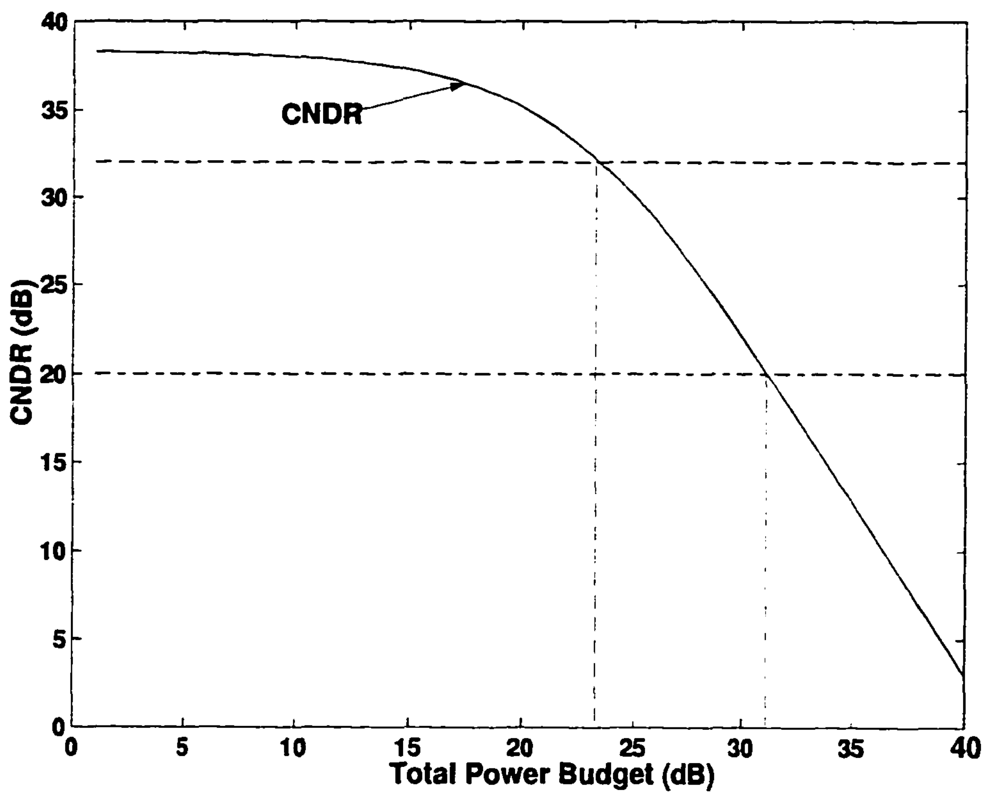

Figure 6.3 CNDR versus Power Budget for QAM system

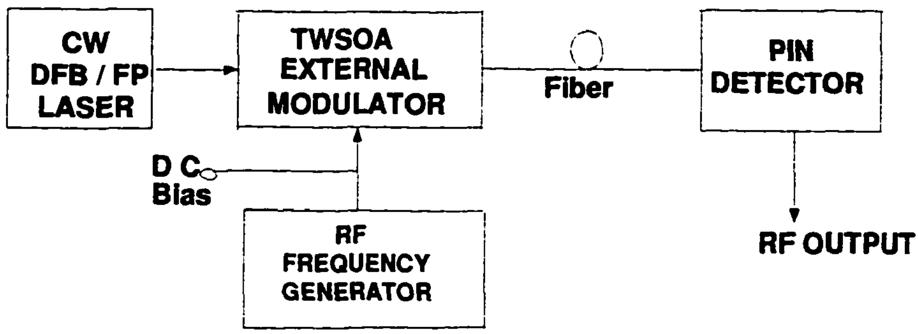

Figure 7.1 FM-FDM Link

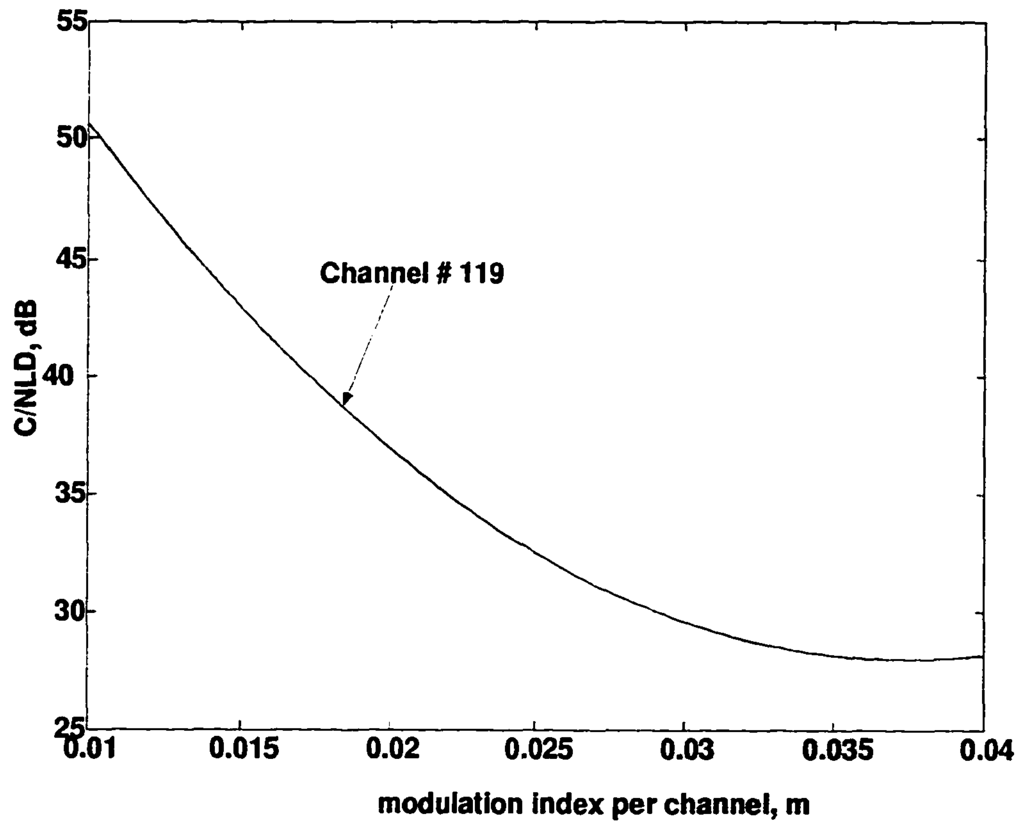

Figure 7.2 C / NLD versus m for FM channel # 119

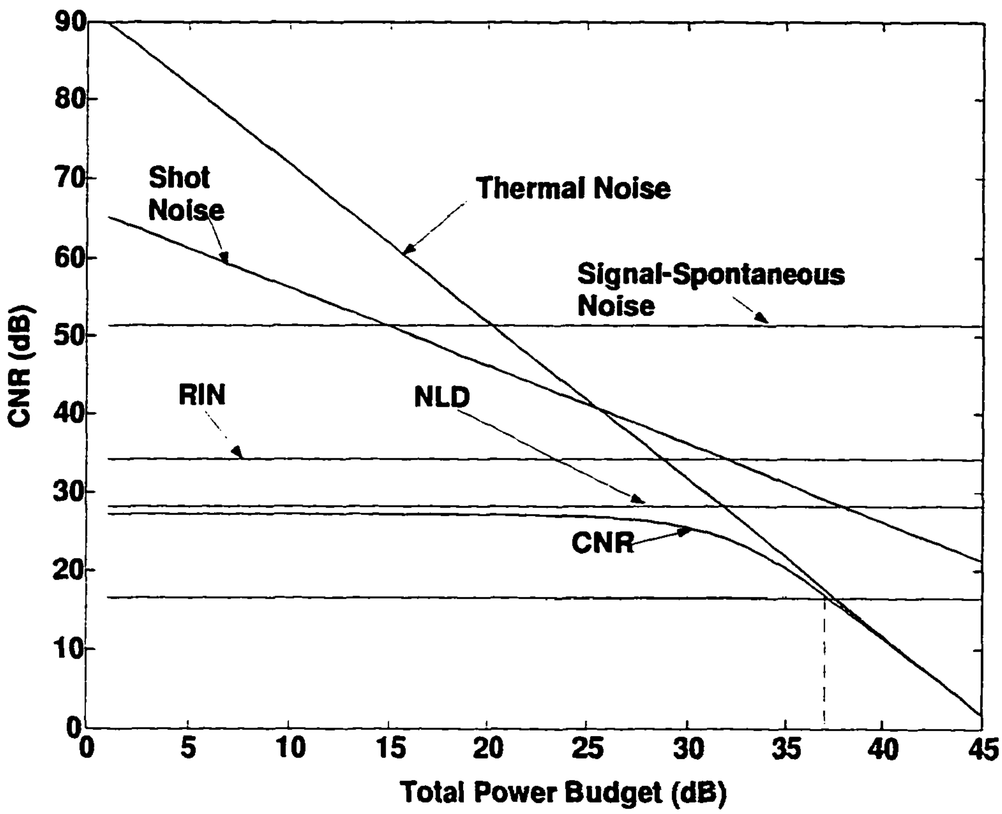

Figure 7.3 CNR and Noise Impairments versus Power Budget for channel # 119 with m = 4.0%

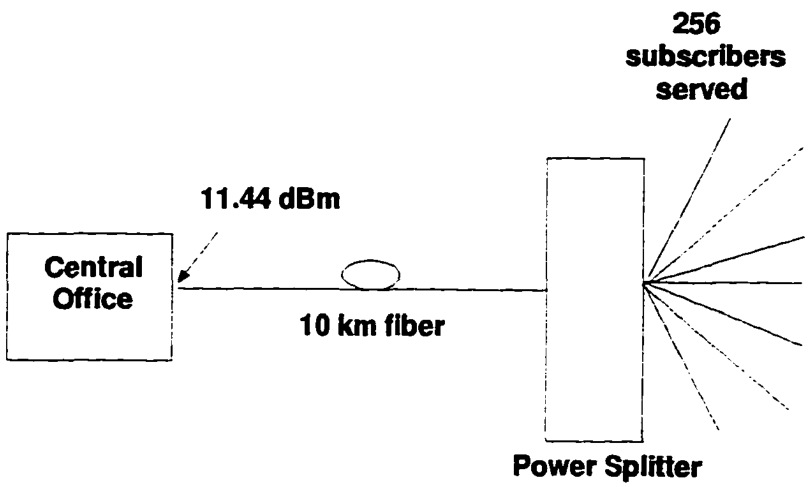

Figure 7.4 FTTH FM - FDM System

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