HELSINKI UNIVERSITY OF TECHNOLOGY

Department of Electrical and Communications Engineering

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TRANSMISSION OF DIGITAL TELEVISION CHANNELS IN WAVELENGTH DIVISION MULTIPLEXED FIBRE OPTIC NETWORK.

This thesis has been submitted for official examination for the degree of Master of Science in Electrical Engineering in Espoo, 3.12.2001.

Supervisor: Prof. Erkki Ikonen

Instructors: M. Sc. Tapio Niemi and B. Sc. Timo Rantanen

TEKNILLINEN KORKEAKOULU DIPLOMITYÖN TIIVISTELMÄ

Tekijä:	Pekka Hilke		
Työn nimi:	Digitaalisten televisiokanavien siirto aallonpituusjakoisessa kuituoptisessa verkossa		
Päivämäärä:	15.11.2001	Sivumäärä:128	
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Professuuri:		S-108 Mittaustekniikka	
Työn valvoja:		Prof. Erkki Ikonen	
Työn ohjaajat:		DI Tapio Niemi ja Ins. Timo Rantanen	

Tiivistelmä:

Työn tarkoituksena oli tutkia digitaalisen videokuvan siirtoa aallonpituusjaotellussa kaapelitelevisioverkossa ja saavuttaa käytännön ymmärtämys kyseisistä verkoista. Tutkimuksen kohteena olivat vain optinen- ja koaksiaalikaapeli-siirtoverkko, joista optiset asiat saivat suuremman painoarvon. Kaapelitelevisioverkon päävahvistimet ja digitaaliset runkoverkot rajattiin kokonaan pois tästä työstä.

Tiedonsiirron teoriaa muun muassa modulaatiomenetelmistä, kohina- ja särörajoituksista sekä optisista epälineaarisuuksista esitetään. Aallonpituusjaotelluissa verkoissa käytetyt komponentit käydään lyhyesti läpi. Komponenteille ja tiedonsiirtoa rajoittaville ilmiöille tehdyt mittaukset esitetään ja lopuksi työssä rakennettuja verkkotopologioita analysoidaan.

Perinteiset ohjelmalähetykset rakennetuissa demonstraatio verkoissa hoidettiin ulkoisesti moduloidulla lähettimellä. Uudentyyppisiin segmentoituihin palveluihin käytettiin suoraan moduloituja lähettimiä. Koe verkoissa käytetyt kanavakuormat oli valittu vastaamaan nykyisiä palveluita ja palveluita joita tulevaisuudessa arvellaan käytettävän.

Avainsanat: Tiheä aallonpituusjaottelu, Digitaaliset video lähetykset, Kaapeli-TV, Optinen tietoliikenne

HELSINKI UNIVERSITY OF TECHNOLOGY ABSTRACT OF MASTER'S THESIS

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Abstract:

The object of this work was to study the digital video broadcasting in the wavelength division multiplexed cable television network and to achieve a practical understanding of such networks. Only the hybrid fibre-coaxial networks are included into this work emphasis being on the fibre optic networks. The headends and the digital backbones are left outside the scope of this work.

Theory of the transmission is described including modulation formats, noise limitations, distortion effects and optical nonlinearities. An overview of the components used in the wavelength division multiplexed networks is given. Measurements of the components and the limiting phenomena are presented and discussed. Finally, test networks are realised and analysed.

In these networks, externally modulated transmitter is used for traditional broadcast services, whereas narrowcast services are transmitted using directly modulated DWDM transmitters. Loading of the transmitters in test networks is done with channel allocations that are considered typical these days and with channel plans that seem to be natural for the future evolution.

Keywords: Dense wavelength division multiplexing, Digital video broadcasting, Cable-TV, Optical communication

FOREWORDS

This work has been carried out at HFC Group of Teleste Corporation and has been a part of OPLATE[1] research project. Share in this project provided me with very valuable help from Simo Tammela from VTT and Tapio Niemi from TKK. Tapio Niemi has also been an instructor of this work and has made a great effort in advising me in writing as well as in understanding the theory. I would like to thank Simo and especially Tapio for their help. Timo Rantanen the other instructor of this work deserves many thanks for his help with practical questions of lifelike optical CATV networks as well as assigning me to this work. Working environment in HFC group has always been good and I want to give many thanks to all members of the group, especially Aarne Jaakkola and Karri Vanhatalo who both have helped me in this work. I would also like to thank my supervisor professor Erkki Ikonen.

Turku, November 15th 2001

Pekka Hilke

Pekka Hilke

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[1] Collaborate research project of Nokia, NK-cables,VTT, TKK and Teleste. Partly funded by TEKES.

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TABLE OF ACRONYMS

BER	Bit Error Ratio
CATV	Common Antenna Television / Cable Television
CSO	Composite Second Order distortion
CTB	Composite Triple Beat distortion
DSF	Dispersion Shifted Fibre
DVB-C	Digital Video Broadcasting standard for Cable transmission
DVB-S	Digital Video Broadcasting standard for Satellite transmission
DVB-T	Digital Video Broadcasting standard for Terrestrial transmission
DWDM	Dense Wavelength Division Multiplexing
EAM	Electro Absorption Modulators
FWM	Four-Wave Mixing
IIN	Induced Intensity Noise
MPI	Multi Path Interference
MER	Modulation Error Ratio
NICAM	Near Instantaneous Companded Audio Multiplex
NTC	Negative Temperature Coefficient
NVOD	Near Video On Demand
PMD	Polarisation Mode Dispersion
QAM	Quadrature Amplitude Modulation
RBW	Resolution Bandwidth
SBS	Stimulated Brillouin Scattering
SMF	Single Mode Fibre
SPM	Self-Phase Modulation

SRS	Stimulated Raman Scattering
TEC	Thermo Electric Cooler
VOD	Video On Demand
XPM	Cross-Phase Modulation

TABLE OF SYMBOLS

A_{eff}	Effective area of the fibre
B	Band width
c_0	Velocity of light
D	Dispersion parameter
e	Charge of an electron
f	Frequency of an electrical signal
g_B	Brillouin gain coefficient
h	Planck constant
I_b	Bias current of the laser diode
I_{DC}	Average current
I_r	Noise current density of the receiver [$pA/(Hz)^{1/2}$].
I_{th}	Threshold current of the laser diode
k	Boltzmans constant or number of number of bits per symbol in Eq. 38.
L	Length of the fibre
M	Number of constellation points
m	Optical modulation index (OMI)
q	Charge of an electron
R	Resistance
S	Fraction of scattering captured by fibre
T	Temperature [K]
α	Attenuation coefficient of fibre
α_s	Proportion of signal scattered per unit length
$\Delta\nu_B$	SBS gain linewidth

$\Delta\nu_L$	Broadened effective laser linewidth
$\Delta\omega$	Linewidth of the frequency
γ_b	Signal to noise ratio per bit.
η_{FM}	Chirp efficiency of the light source [MHz/mA]
μ	RMS modulation index
ν	Frequency of the light
ζ	Polarization state parameter

1 INTRODUCTION

Common Antenna Television (CATV) networks are used to deliver video services to homes. Early networks were based on coaxial cable, but in late 80's fibre optic links were introduced to achieve longer distances. Today these Hybrid Fibre Coaxial (HFC) networks may be connected to form large systems via digital backbones using ATM/SDH based solutions. One service provider may operate a network connecting several cities and the service may be available to more than one million households.

In this study, the focus is on the transparent HFC network. Transparency means that video, voice or data with many different modulation formats can be transmitted over the network. The network itself makes very little limitations to equipment connected to each end.

Services that can be offered are almost unlimited. However, bandwidth per subscriber is chosen by the network segmentation, by the number of subscribers per node. Information can typically be transferred both ways. The return channel is commonly called the reverse path or upstream. Older systems were used only for broadcasting without any return path. The service was very similar to terrestrial broadcasting but the channel count was higher. The return channel allows interactive services and the other new technique called Dense Wavelength Division Multiplexing (DWDM) makes narrowcast targeted services to certain areas or to individual users possible. These services include video on demand, cable phone and fast Internet access. These new services use digital modulation schemes, out of which the Quadrature Amplitude Modulation (QAM) is often the most suitable one.

Dense Wavelength Division Multiplexing (DWDM) is a technique for launching several wavelengths of light into one fibre. Each wavelength can be separately modulated and detected thus allowing several virtual links to be transmitted over one fibre. This can be helpful in segmentation or simply a way to reduce fibre count.

DWDM technology has been exploited for several years in optical telecommunication. In CATV networks, this new technology is still taking form and seeking possible applications. Some DWDM solutions have been installed into CATV networks as well. The first ones were installed in the USA where the lack of available fibres has been the main driving force. In Europe, the interest in the use of several wavelengths in CATV networks is rising rapidly and a few installations have been made. DWDM products made by Teleste exist, partly as a result of this work, and they have been proven to work by both me (with help of my colleagues) and our customers.

Slowly but unavoidably television broadcasting is changing to digital format. This means different requirements for the CATV network. The existing networks may actually be too good for the new signals and this may enable the implementation of cheaper techniques and longer transmission distances. The digital modulation format also allows more program contents to be transmitted in the same bandwidth.

This work is organized as follows: After the introduction the CATV networks will be briefly explained in chapter 2. Chapter 3 takes a look at the optics used in CATV systems, noise and distortion problems in the network are discussed in chapter 4. Chapters 5 and 6 present measurements made for components and systems. In the last chapter, some example networks are presented and analysed.

2 CATV NETWORKS

Modern CATV networks vary in size and complexity. An example of a large CATV system is shown in Fig. 1. The heart of the system is the main headend, which can be called super headend or playout in the very large systems. This is the point where most of the program contents are inserted into the network. Sources for signals are for example terrestrial transmission, satellite receivers, video recorders and digital video-on-demand servers. The main headend is connected to sub headends with a digital fibre backbone. The digital backbone allows flexible routing and high transmission capacity. Conversion from digital to analogue typically takes place in the sub headend. An analogue fibre optic network carries the signal from the sub headends to the fibre nodes. A fibre node is a point connecting the coaxial cable network and the optical link. Together the fibre link and the coaxial network are called Hybrid Fibre Coaxial (HFC) network.

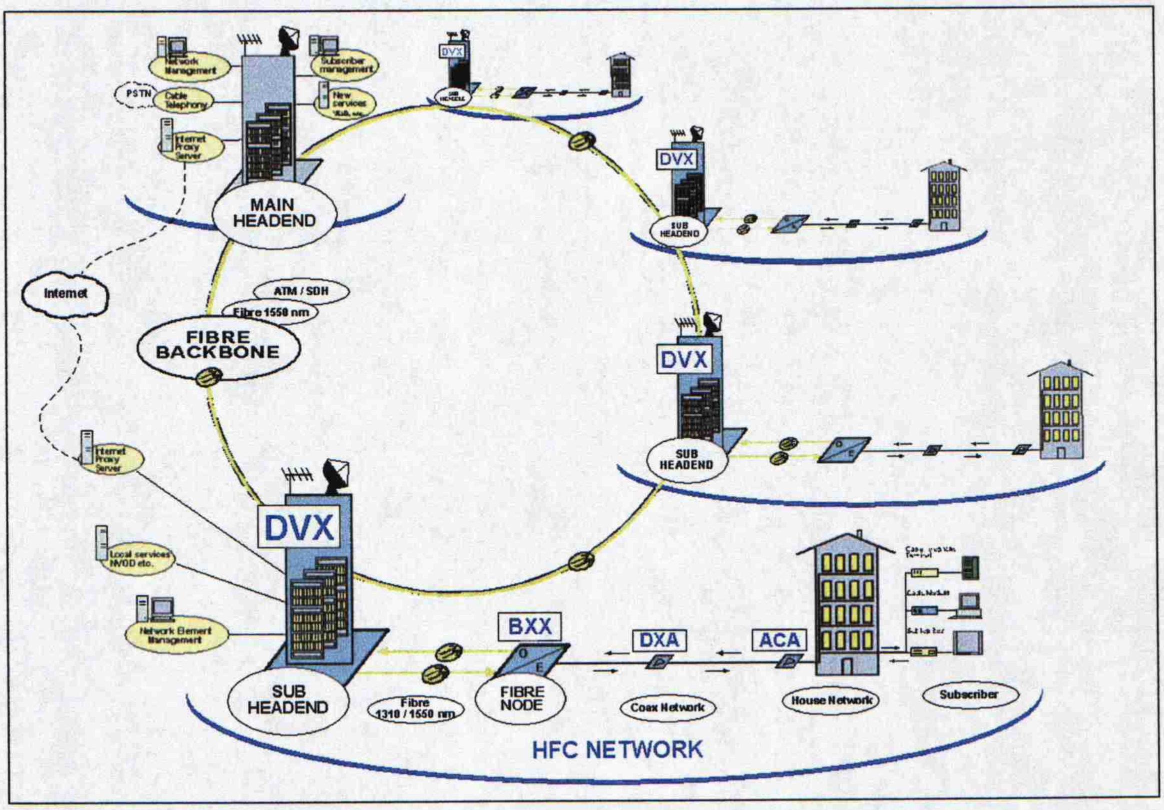


Figure 1. Example of a large CATV network.

The headend and the digital backbone are left outside of the scope of this work and everything is seen from the perspective of the HFC the emphasis being on the effects of the fibre transmission on the signal.

Optical links in the HFC networks are mainly point-to-point or point to multi-point links. The links have quite often a back up route and both spare transmitters and receivers for redundancy. Even though fibre is getting closer to the homes, optical links are typically followed by a coaxial cable network. Fibre to the home is still too expensive and coaxial cable networks are able to deliver all current services.

The management systems of the CATV networks are quite new and they are one of the biggest factors differentiating networks built by different manufacturers these days.

The basic idea of the HFC network is to be transparent. The signal injected in to a network should appear to each connected household untouched. Quite high signal levels needs to be used to avoid problems with noise. The network uses frequency multiplexing and the number of the electrical carriers can be as high as 110. In Europe, the typical number of channels is smaller mainly from 40 to 50. The total power of a large number of carriers is quite high and the linearity requirements of the active devices are strict. (Linearity and noise issues will be treated in detail in chapter 4.)

The HFC network is not only used for broadcasting but the signals go both ways. The return channel is needed for interactive services such as Internet data or cable telephony. In the coax network the return channel is separated by diplex filters and low frequencies are used for return signals and high frequencies for broadcast. From node to the headend, return channel mostly uses a separate fibre though wavelength division multiplexing could be used.

A typical RF spectrum of a HFC network load is shown in Fig. 2. Return channels are allocated between 5 and 65 MHz, FM radio channels from 87 to 108 MHz, analogue TV channels from 120 MHz up to 500 - 700 MHz and QAM modulated DVB-C channels on the high end of the frequency spectrum.

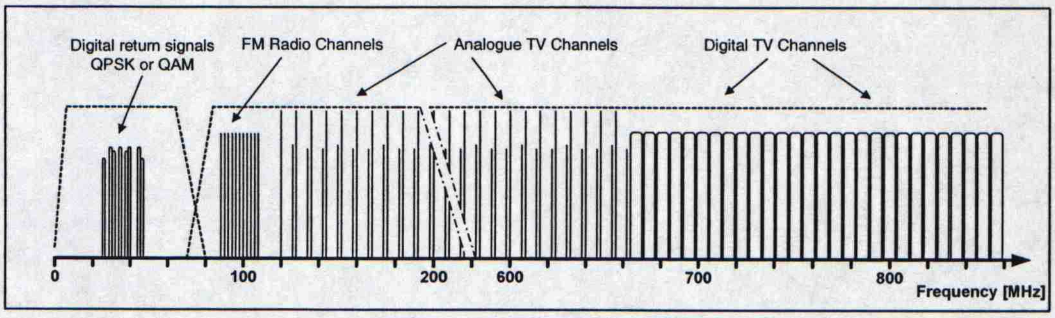


Figure 2. Typical RF spectrum of a CATV network.

2.1 Analogue TV signals

Analogue TV transmission has survived quite a few years. Colour was added later and this was done in such a way that old black and white receivers could still be used. Sound of the TV-signal was first upgraded from mono to stereo and later an extra digital sound (NICAM) was added. The actual video signal bandwidth around the carrier varies from 4 to 6 MHz. The channel spacing is somewhat larger, from 6 to 8 MHz including the sound carriers and safety margins. The lowest carrier frequency used is commonly 112 MHz and the highest frequencies in the network may be designed up to 550, 600, 750 or 860 MHz. All the new networks in Europe are built up to 860 MHz.

2.2 Digital TV signals

Digital format has many advantages over the analogue format in terms of storage and transmission quality. Main advantages are quasi error free reproduction of the signal and the packing. Packing allows more contents to be transmitted in a given bandwidth. Digital picture can be packed quite efficiently. This means that information of picture contents is reduced. With moderate packing ratios very little harm to picture is done. With high packing ratios deterioration of the picture quality is remarkable. Digital television is commonly promoted by better picture quality but this is not the whole truth. Quality of analogue picture is better but transmission always adds distortion and noise to analogue picture. Level of the impurities in analogue transmission and level of the packing in digital picture are the factors affecting subjective judgement of which format has the better picture in reception.

Packing of the picture is mainly done in two different ways. Firstly, all details that are judged not to be important are removed. Secondly, only the changes in the picture are reported. Typical example of the latter is a car moving across the view, practically the only information receiver needs is the change in the place of the car. This approach can be taken a little bit further and not anything that can be predicted in reception is sent. The next position of the car is quite easy to guess over a relatively long period of time and therefore no information needs to be sent, the receiver makes the picture all by itself. The transmitter uses the same rules for predicting the picture as receiver and reports only when the error in the predicted picture is large enough compared to original. Weakness of this technique is often demonstrated by tennis. In a digital picture, the ball flies through the racket with every stroke but finds the correct position quite quickly by discontinuous path. Different and less effective packing techniques can of course be used for sports.

2.2.1 Digital modulation (M-QAM)

M-QAM stands for Quadrature Amplitude Modulation and the M denotes the level of the modulation. Level means here the maximum value of the transmitted symbol. In other words, each transmitted symbol represents 4,6 or 8 bits for 16-, 64- or 256-QAM respectively.

The idea of IQ modulation or QAM is in the orthogonal carriers. Two sinusoidal signals at the same frequency having 90° phase shift can be amplitude modulated and detected without interference. This doubles the data rate that can be transmitted in a given bandwidth. The principles of QAM modulator and demodulator are shown in Figs. 3 and 4.

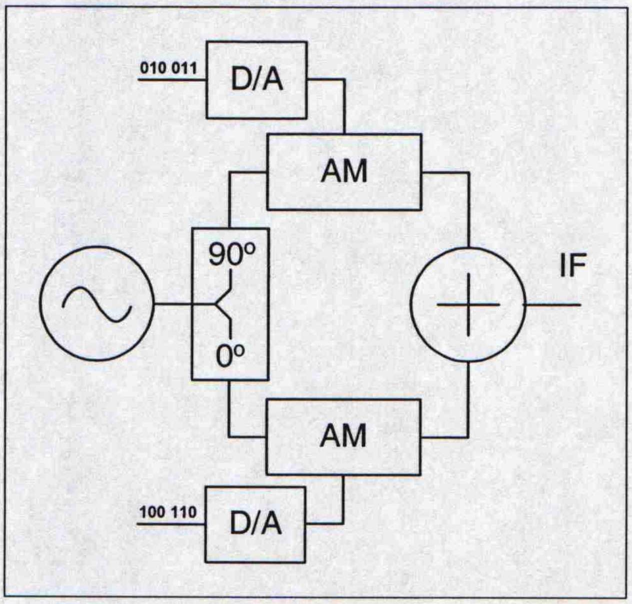


Figure 3. Principle of an IQ modulator.

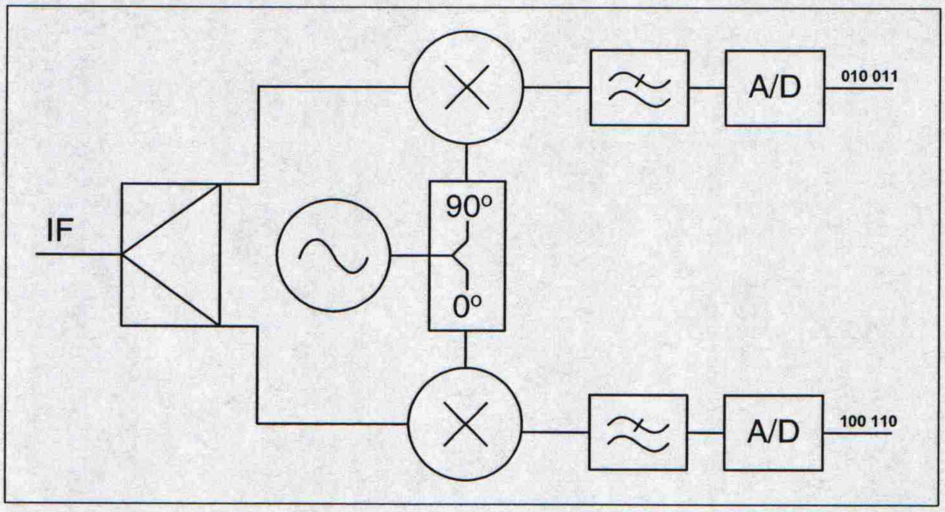


Figure 4. Principle of an IQ demodulator.

In practice, these devices are not made of analogue mixers anymore. Analogue techniques suffer from carrier leakage, compression and amplitude imbalance. Cheap digital electronics have replaced the difficult analogue designs and have much better performance.

The orthogonality of the carriers can be explained with geometry as follows:

Let A(t) and B(t) be the modulating functions. Modulation is digital and therefore functions have only discrete values. Values are in a range {0, … ,n-1, n}, where n depends on the order of modulation n=$\log_2$(M). For actual modulation these values are coded to {(n-1)/2, (n-3)/2, …(-n+1)/2} in order to keep the peak power as small as possible. These two functions are then separately used to modulate the orthogonal IF carriers. The resulting QAM modulated function is

$$g(t) = A(t)\sin(t) + B(t)\cos(t)\ . \tag{1}$$

In the receiver, this function is multiplied separately by 2sin(t) and 2cos(t). In-phase signal I is detected using 2sin(t). This can be presented as

$$\mathrm{I(t)} = \big(A(t)\sin(t) + B(t)\cos(t)\big)\cdot 2\cdot\sin(t) \tag{2}$$

$$\mathrm{I(t)} = A(t)\big[\cos(0) - \cos(2t)\big] + B(t)\big[\sin(2t) + \sin(0)\big] \tag{3}$$

$$\mathrm{I(t)} = A(t) + A(t)\cos(2t) + B(t)\sin(2t)\,. \tag{4}$$

As $A(t)$ is very low frequency signal compared to $A(t)$cos(2t) and $B(t)$sin(2t), $A(t)$ can easily be restored using low pass filtering.

Quadrature-phase signal (Q) is detected similarly using 2cos(t) and result is

$$\mathrm{Q(t)} = B(t) + A(t)\sin(2t) + B(t)\cos(2t)\,. \tag{5}$$

Again, $B(t)$ can be restored using low pass filtering.

Figures 5 through 8 show a 16-QAM simulation made with the Mathcad software. Simulation is done using equations presented above thus signals in the figures are not real band limited signals. Amplitudes of the modulating functions are ±{0.5, 1,5} as explained earlier and the symbol rate is 1/10Hz, one tenth of the IF frequency 1Hz.

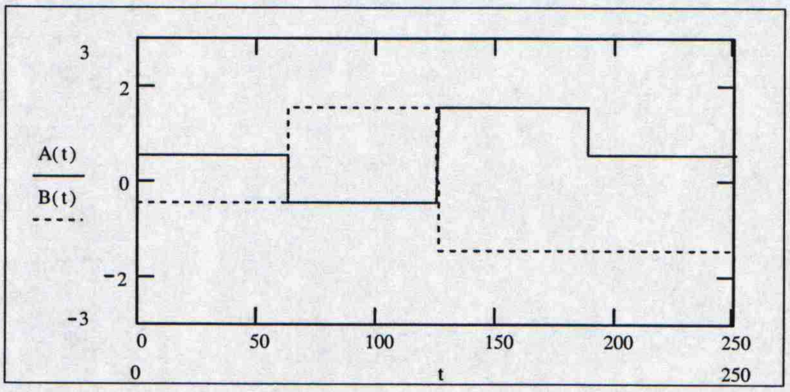


Figure 5. Modulating baseband signal.

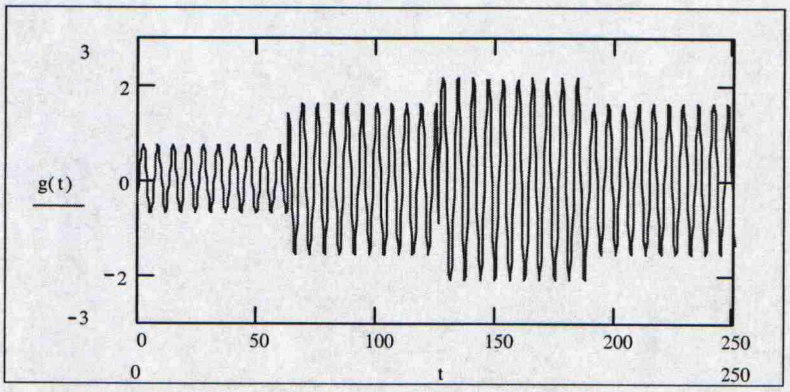


Figure 6. 16-QAM signal

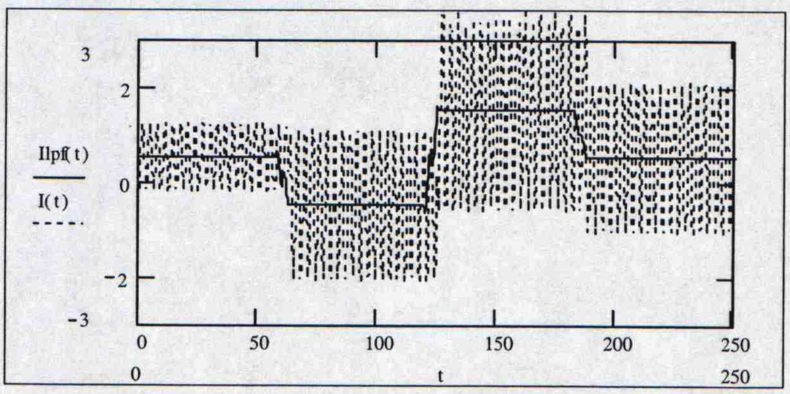


Figure 7. Demodulated in-phase signal and the same signal with low pass filtering.

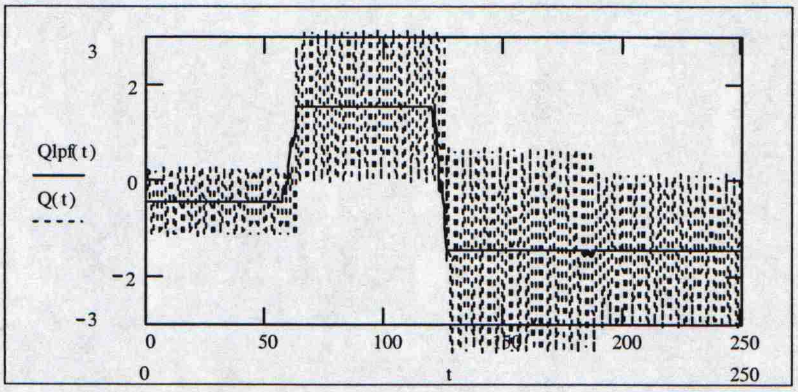


Figure 8. Demodulated quadrature-phase signal and the same signal low pass filtered.

QAM modulated signals are often viewed as constellation diagrams. Constellation diagram is an x-y coordinate presentation of demodulated in-phase and quadrature-phase signals where a certain number of detected voltages at decision points are presented. This is an informative way of looking signal quality of a QAM modulated signal. All points should ideally be in the centre of the squares. Example of a constellation diagram is presented in Fig. 9.

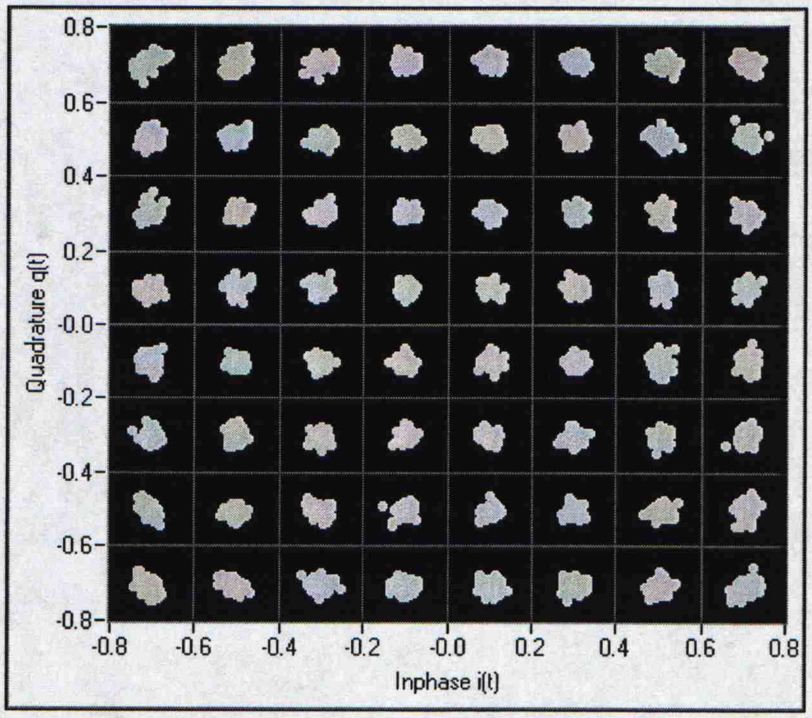


Figure 9. Example of 64-QAM constellation diagram.

Each constellation point corresponds to a certain bit sequence. Constellation diagram with corresponding bit sequences for 64-QAM is shown in Fig. 10.

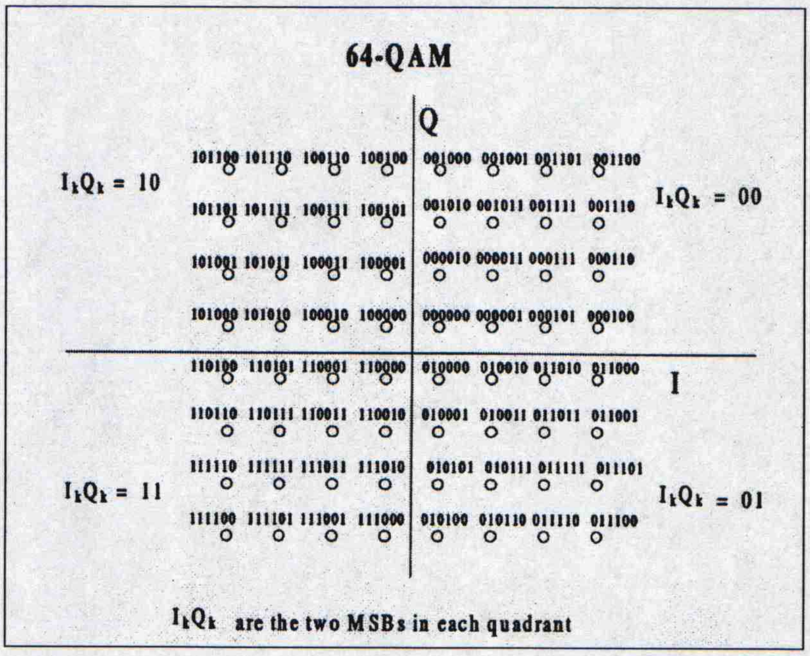


Figure 10. Bit to symbol mapping in 64-QAM. /1/

3 OPTICAL NETWORK COMPONENTS

The use of optical fibres enables huge improvements in the transmission distances over the coaxial cables due to low loss. Reliability of the network is improved also with the lower count of active devices. Attenuation of the coaxial cable is much larger than that of an optical fibre. In a cable network an amplifier is needed after every 400m, whereas in fibre even 60 km transmission without amplification can be realised. Cost of installing long fibres is of course high but can be justified, as expensive signal processing doesn't have to be done locally. The cost of the headend can be divided with large areas and a large number of subscribers.

3.1 Optical fibre

Optical fibre consists of a core made of glass that is surrounded by a cladding. The cladding has a lower index of refraction than the core and therefore light travelling in the core will be totally reflected back into the core in the core-cladding boundary. This simple ray theory explains the propagation of light in step-index multi-mode fibre with a large core.

3.1.1 Types of fibre

Numerous types of fibre exist for different applications. Optical CATV networks are almost completely built of standard single-mode fibre (SMF).

Multi-mode fibres allow light to travel at various speeds through the fibre introducing distortion to the signal. Different types of multi-mode fibres exist and they can be studied in any book on fibre optics, for example in /2/.

In standard single-mode fibre the chromatic dispersion minimum is at 1310 nm and has not been shifted to any other wavelength. Earlier transmission at 1310 nm was a standard, as fabrication methods of 1550 nm lasers had not been developed yet. Nowadays, 1550 nm transmission is increasingly popular because of low losses and possibility of optical amplification without conversion into electrical domain.

Dispersion shifted fibre (DSF) has the minimum dispersion in 1550 nm window but 1310 nm transmission is then difficult and the DWDM applications are impossible due to optical nonlinearities.

Non-zero dispersion shifted fibre has low dispersion in 1550 nm region for suppressing the dispersion effects but still allowing DWDM applications. Dispersion compensating fibre has negative dispersion and therefore compensates the dispersion of SMF with a cost of extra attenuation.

Special fibres like polarisation maintaining fibre and holey fibres, can have tailored properties such as dispersion compensation, dispersion shift or extended pass band.

As almost only SMF is used in the considered networks the other types of fibre are not given any more attention.

3.1.2 Properties of the single-mode fibre

Propagation of light in a SMF cannot be explained with the simple ray theory. Geometry of the fibre allows light to propagate through exactly one path and each photon travels the same distance in the fibre. In SMF actually two modes of propagation exists. In an ideal case, these two modes travel with the same speed and are separated only by the polarisation state. However, phenomena called polarisation mode dispersion (PMD) exist due to the errors in the geometry of the fibre. Because of the errors the two polarisation states may propagate at slightly different velocities in different parts of the fibre and some part of the light may be coupled from one polarisation state to the other at some points.

Attenuation of the fibre is a function of wavelength as shown in the Fig. 11. The attenuation has two minima one at 1310 nm and the other at 1550 nm. These minima are called transmission windows and both 1310 nm and 1550 nm windows are used in CATV links. Typical attenuation at 1310 nm is 0.35 dB/km and 0.25 dB/km at 1550 nm, but can be lower than 0.3 dB/km and 0.2 dB/km, respectively. Attenuation of the fibre is mainly caused by scattering and absorption. Attenuation, in the 1310 and 1550 nm regions, is mainly caused by the intrinsic Rayleigh scattering of silica class but also by scattering from imperfections of the materials and geometry. Absorption is caused by impurities and the dominant factor is the hydroxyl ion (OH^-) that causes high absorption peak at ~1383nm.

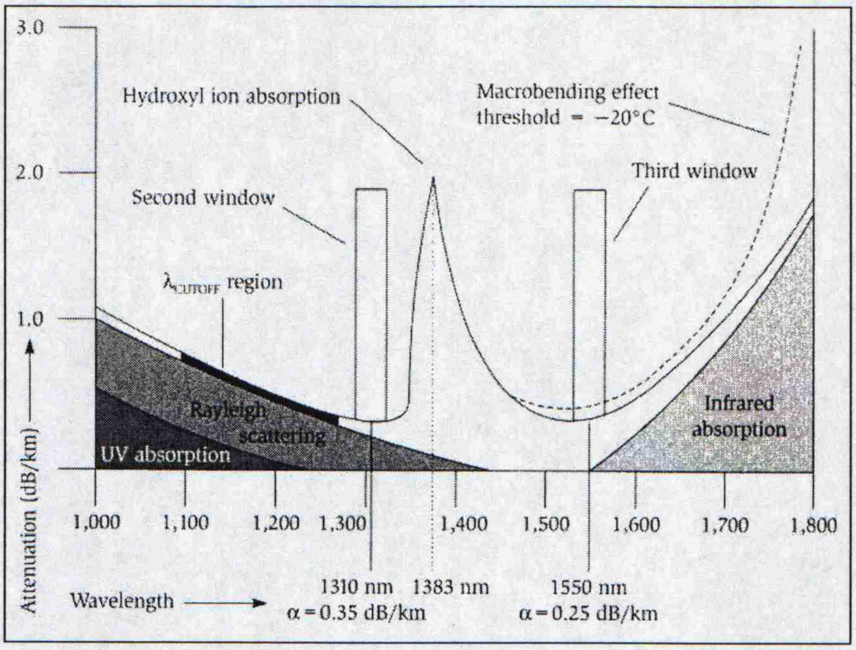


Figure 11, Spectral attenuation of SMF fibre /3/.

Different wavelengths propagate with different velocity through the fibre. This is called chromatic dispersion. Dispersion is caused partly by the material used (material dispersion) but also by the geometry of the fibre (waveguide dispersion). The dispersion of the fibre can be modified by changing the geometry of the fibre and by tailoring the refractive indices of the core and the cladding. Figure 12 shows the chromatic dispersion in SMF and DSF.

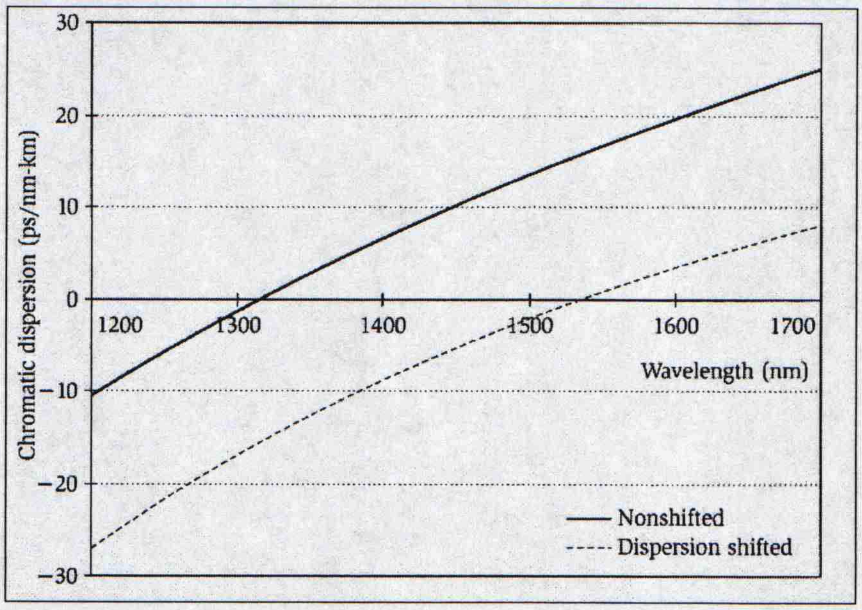


Figure 12, Chromatic dispersion of SMF and DSF fibre /3/.

3.1.3 Fibre connections

Connecting fibres is done by splicing or with connectors. Splicing is permanent but reliable and adds only very little loss to fibre. Splice loss is typically less than 0.05 dB.

Several types of fibre connectors exist. Some of them snap, others attach with a thread or a bayonet. All connectors however cause some attenuation that varies from connection to connection. Typical connection loss is from 0.1 to 0.5 dB. Ends of the fibres in the connectors are polished to minimise reflections. Depending on the type of the polishing the connectors are called Polished Connectors (PC), Ultra Polished Connectors (UPC) and Angle Polished Connectors (APC). Reflection loss is of a PC is typically better than 40 dB, for UPC better than 50 dB and over 60 dB for APC connector. Reflection in APC is actually equal to PC but reflected light does not couple back into the fibre because of the angled cut of the fibre end.

3.2 Splitters/couplers

Splitters or couplers are used to split one signal to many signals or to combine many to one. These devices work both ways. Splitting of the signals causes attenuation as energy of the signal is divided between several fibres. Theoretically attenuation A is

$$A = -10 \cdot Log(N) \tag{6}$$

Where N is number of fibres power is divided to. Extra attenuation is often very small. Split ratio can also be uneven such as 30/70 or 5/95.

3.3 Transmitters

Optical transmitters convert electrical signal to light. Light source in the CATV networks is always a laser diode but two different approaches to modulate light exists. At wavelength of 1310 nm directly modulated transmitter is a cost effective choice in short point-to-point links. With longer links over ~40km, the 1550 nm externally modulated transmitter is the only option.

3.3.1 Directly modulated transmitters

Directly modulated transmitters are the simplest type of the optical transmitters. Modulation of the laser is applied directly to the bias current, allowing a quite simple construction. The drawback of the direct modulation is that it causes frequency chirping. Chirp broadens the spectrum of the laser more than the intensity modulation itself does. This is discussed further in chapter 4.2.3. When operating in 1310 nm window this is no problem, but at 1550 nm, dispersion in fibre causes problems especially for analogue transmission. For shorter distances, nowadays up to 40 km, 1310 nm directly modulated transmitters can offer high signal quality. For longer distances or for point-to-multipoint links an externally modulated transmitter at 1550 nm is a better choice. In this study, the directly modulated transmitter at 1550 nm is examined for use in digital TV transmission.

3.3.2 Externally modulated transmitters

Externally modulated transmitters consist of a continuous wave laser and a modulator. This construction doesn't broaden the spectrum of the laser more than the modulation does and allows longer transmission distances in the 1550 nm window. However, stimulated Brillouin scattering (SBS, see 4.3.2) limits the power density that can be launched into the fibre and for this reason it is necessary to artificially broaden the line width of the laser. This is done with GHz phase modulation and if more than one tone is used the inter modulation products have to be outside the modulation bandwidth. Modulators used in the analogue CATV transmitters are exclusively Mach-Zehnder interferometers. The modulator itself is not very linear and a pre-distortion circuitry is always used in the externally modulated transmitters.

Electro absorption modulators (EAMs) are not linear enough to be used in analogue CATV links but the use of EAM has been successfully demonstrated in long distance DVB-C transmission/4/. In this work, however, EAMs are left outside further treatment.

3.4 Receivers

Purpose of the optical receiver is to convert the optical signal back to electric form. Key component of the receiver is a photodiode. There are two types of photodiodes, PIN diodes and avalanche diodes. PIN diodes are the only type used in CATV networks. Avalanche diodes cannot be used due their poor linearity. Structure of the PIN diode is presented in Fig. 13. In theory, each photon arriving to I-region of reverse biased PIN diode allows one electron to pass the region. In the diode, the optical signal is converted to current, which again, is converted to voltage in the resistor of the receiver.

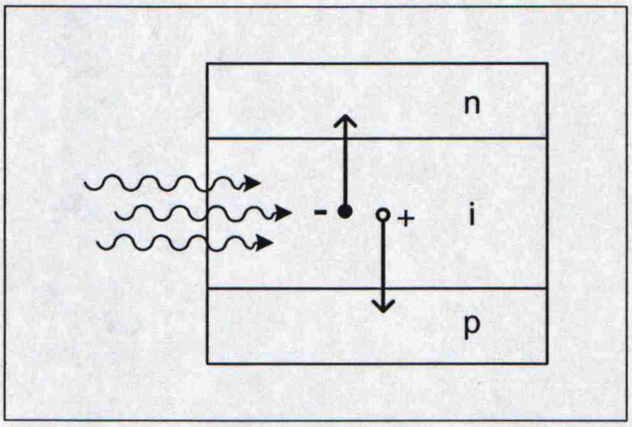


Figure 13. PIN diode structure.

Important parameters of the PIN diode are the linearity and the capacitance. Linearity can be affected by bias voltage and improved with push-pull receiver configuration. However, linearity is mostly a property of a diode and the selection of the proper type of the diode is important. Impedance of the diode is high and it can be seen as a current source at low frequencies. Capacitance of the diode, however, lowers the impedance at higher frequencies and makes the coupling to practical impedance, typically 75Ω, over several octaves tricky. Small capacitance helps in realising good coupling, flat frequency response and good noise performance.

Responsivity of the diode describes the ability of the diode to convert the light to electric current. Typically, responsivity is expressed in [mA/mW], or as percentage of the theoretical maximum called the quantum limit. Maximum responsivity can be written as

$$R = \frac{e}{h\nu} = \frac{e}{hc} \cdot \lambda \, [A/W] \tag{7}$$

where e is the charge of an electron, h is the Planck constant, ν is the frequency of light, c is the velocity of light and λ is the wavelength of the light.

At longer wavelengths, the photons have smaller energy and therefore there are more photons arriving to receiver in the longer wavelength light having the same power. Maximum responsivity of a receiver diode is therefore higher in longer wavelengths. Upper wavelength boundary of receiver diode is limited by energy of photon, which falls below the energy gap. The lower boundary is set by UV absorption of the receiver diode material or by the cut off wavelength in fibre. In practise efficiency of the diode is typically 70 ...80% of this theoretical maximum (Fig. 14).

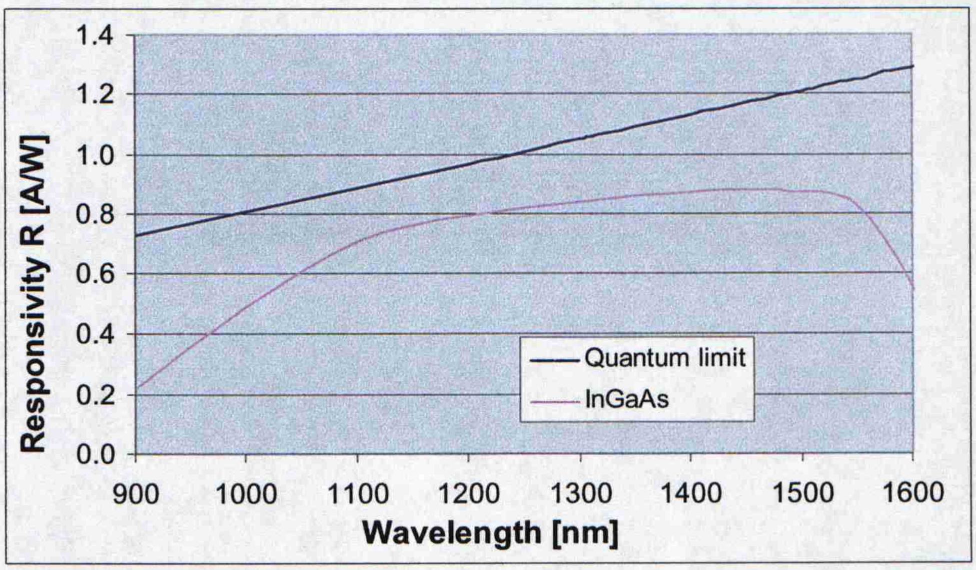


Figure 14, Quantum limit of the responsivity and responsivity of an InGaAs PIN diode /5/.

Two simplified receiver topologies are presented in Fig. 15. Advantages of the push-pull receiver compared to the simple one are the doubled received voltage and enhanced second order distortion performance.

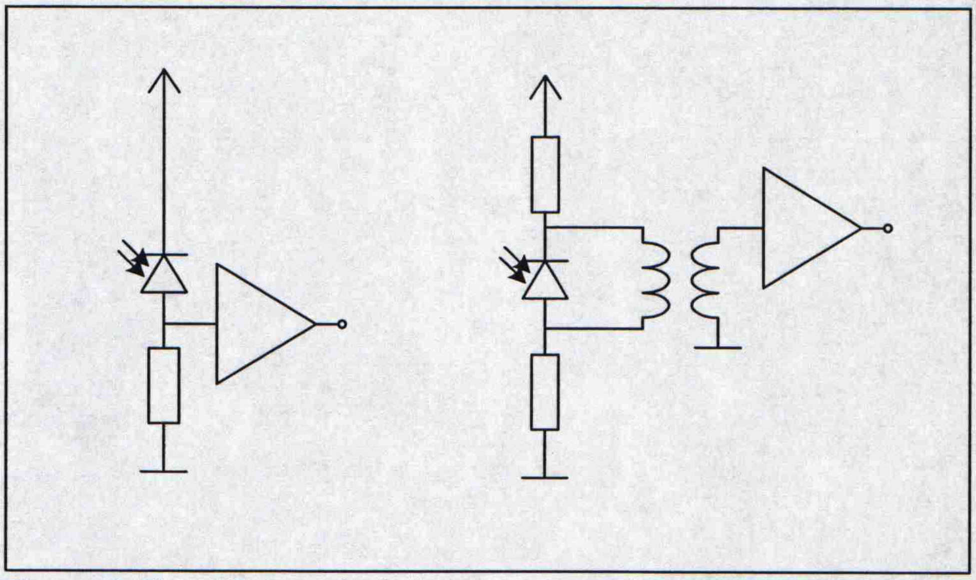


Figure 15. Simplified topologies of a simple receiver and a push-pull receiver.

3.5 DWDM Filters

In the DWDM system, the filters play a very important role. These components allow several wavelengths to be multiplexed and de-multiplexed.

Important parameters of the DWDM filters are attenuation, adjacent channel attenuation, width of the pass band, and flatness of the pass band, group delay, polarisation dependent loss and reflection loss. Devices used in this work were built using thin-film filter technology and allowed multiplexing and de-multiplexing of light with low enough attenuation and crosstalk. Distortion added to signal in these devices is also minimal. Thin-film filters have lower attenuation than arrayed waveguide (AWG) filters when channel count is low. With 32 or more channels, AWG filters are probably a better choice. In this work, only 8 channels were used and AWG filters were left outside of the scope of this work.

DWDM channels are specified by ITU /6/ and typically, this recommendation is used. Channel number in ITU Grid (Appendix B) is the third and the fourth digit in light frequency, for example 19<u>53</u>00GHz is the frequency of the channel 53, which corresponds to a wavelength of 1535.04nm in vacuum. Wavelength in vacuum is more commonly used than the optical frequency.

3.6 Optical Amplifiers

Optical amplifiers are used to amplify optical signals. Earlier repeaters were needed to achieve longer distances.Even though other solutions have been reported, erbium doped fibre amplifier (EDFA) is the only commercially used solution. EDFA has gain only in 1530 – 1560 nm range and therefore is only operable in the 1550 nm window. In the 1310 nm window, no commercial amplifier solution is currently available.

Most important parameter of an EDFA is the saturated output power. In CATV networks, the EDFAs are always operated in saturated region. The number of fibres fed by the EDFA and the optical loss in the fibres determines the suitable output power. In DWDM networks, the output power is split between the wavelengths. For example, if an EDFA with +16 dBm output power is used +7 dBm output power is available for each of the 8 wavelengths. Output powers commonly used range from +13 dBm to +22 dBm.

The gain of an EDFA can be chosen by the design and it is always high enough to ensure operation in saturated region with input powers used. For a +16 dBm EDFA, for example, the gain must be over 20 dB to keep amplifier in saturation even when the input power is around 0 dBm.

Amplifiers add noise to transmitted signals but deterioration of signal quality is small as long as input power to amplifier is large enough. Typical noise figure of a CATV EDFA is 5-6 dB. With low input powers of -10 dBm the noise figure is lower, only about 4,5 dB. However, overall noise performance requires input powers of 0…+7 dBm to be used.

Gain shape of the amplifier means that the gain varies with wavelength. Gain shape of an EDFA is also a function of the input power. However, with many wavelengths the total input power is always relatively high and the problem is not very severe. With typical link lengths, one amplifier is enough and the gain shape problem can be tolerated. The connectors and the DWDM filters often cause bigger variations to received powers of a DWDM link than the EDFA does.

Because the EFDA is used in saturation, the output power of the other wavelengths will increase if one wavelength is dropped. This may be a problem if summing of narrowcast and broadcast is done optically or automatic gain control is not used in electrical summing. Change from 8 to 7 wavelengths is not very severe as the change in electrical signal level is ~1.2 dB but a change from 3 to 2 wavelengths would cause a 3.5 dB change in received electrical level. An increment over 3 dB in narrowcast signal level may deteriorate the broadcast signals.

4 NOISE AND DISTORTION IN CATV SYSTEMS

Two different phenomena, noise and distortion, limit the purity of transmitted signal. To minimise noise problems high signal levels need to be used but to achieve good distortion performance signal levels should be kept low. Signal level in transmission is always a compromise between these two limiting factors and clever design of networks is about making right compromises and selecting the proper components.

4.1 Noise limitations

Noise in optical link can be divided into three different components: intensity noise, shot noise and thermal noise. Intensity noise is noise associated with the transmitter, shot noise appears in the receiver and thermal noise is noise mechanism of the electrical amplifiers.

It is convenient to derive formulas for noise current density as no assumption of receiver topology or impedance is needed and a practical equivalent noise current density parameter can be used to characterise the noise in receiver. From absolute noise power in the link with help of the optical modulation index (OMI) analysis continues to a formula for Carrier-to-noise ratio (C/N).

4.1.1 OMI

Optical modulation index (OMI) is one of the most important parameters of an analogue optical link. Definition of OMI is very similar to definition of modulation index in ordinary AM-modulation. An illustrative explanation of this definition is shown in Fig. 16. OMI is defined as

$$m = \frac{P_{MAX} - P_{MIN}}{P_{MAX} + P_{MIN}} = \frac{P_{MAX} - P_{MIN}}{2 \cdot P_{AVG}}, \qquad (8)$$

where m it the optical modulation index.

OMI is defined for one channel and for sinusoidal signal. Same definition can be used with QAM signals but equivalent power is used to replace the peak value of the sinusoidal modulation.

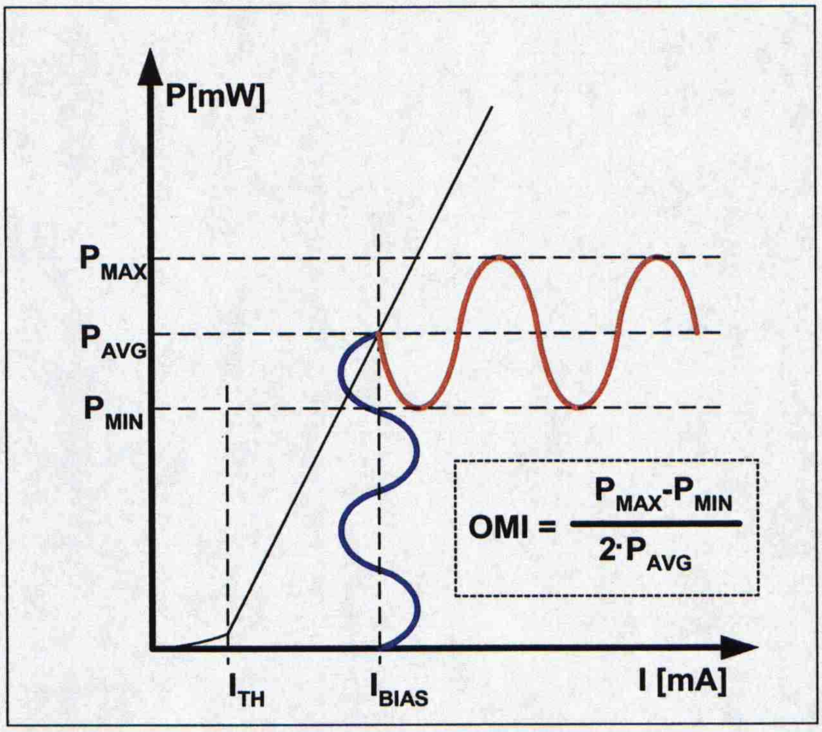


Figure 16. Definition of OMI for an optical transmitter.

Total OMI for several channels is calculated by summing the powers of individual carriers

$$m_T = \sqrt{m_1^2 + m_2^2 + \ldots + m_N^2}\ . \tag{9}$$

Provided that all the channels have equal OMI the formula simplifies to

$$m_T = m\sqrt{N}\ . \tag{10}$$

Definition of total OMI is a practical value in making estimations of the maximum channel counts or C/N calculations for certain number of channels. Peak-to-average ratio for relatively small number of channels can be surprisingly high and clipping may occur at smaller total OMI values than in case of more than 10 channels. For analogue carriers total OMI of 30% is a typical value for 1310 nm directly modulated transmitter and 25-28% for the externally modulated 1550 nm transmitter.

4.1.2 Intensity noise

Intensity noise is a parameter of the transmitter and it determines the maximum achievable C/N in the link. Intensity noise is originated from the optical interference between the stimulated laser signal and spontaneous emissions generated within the laser cavity /1/. Intensity noise is often specified as relative intensity noise (RIN) value and often expressed in [dBc/Hz], optical noise power density compared to the power in the optical carrier. In the electrical domain, this is expressed as a ratio between noise current density and the average current /8/.

$$RIN = \frac{<i^2>}{I_{DC}^2} \quad \left[\frac{1}{Hz}\right] \tag{11}$$

For the bandwidth of interest relative intensity noise current is then given by

$$I_{RIN} = I_{DC}\sqrt{RIN \cdot B} \quad [A]. \tag{12}$$

Three other intensity noise like mechanisms exists: amplifier spontaneous emission (ASE) in EDFA, PM/AM noise and multi path interference (MPI).

ASE noise originates from spontaneous emission of exited erbium atoms in the amplifier. Noise mechanisms in EDFA are quite complicated and this topic is not given further treatment in this work.

Calculation of EDFA noise component is derived from the definition of the noise figure. It is possible to make calculations with assumptions closer to physics but these calculations are not as practical as this one. In addition, the calculations based on measured noise figures give results that are more accurate.

Noise figure of the EDFA can be measured in electrical domain and calculated using the following expression /9/

$$F = 10\lg\left(\left(\frac{m^2}{2B \cdot 10^{\frac{C/N_{out}}{10}}} - \frac{m^2}{2B \cdot 10^{\frac{C/N_{in}}{10}}}\right)\frac{P_{in}}{2h\nu} + \frac{P_{in}}{P_{out}}\right) \tag{13}$$

where m is the optical modulation index, B is the noise bandwidth, h is the Planck constant, ν is the frequency of light, P_{in} and P_{out} are the optical powers measured at the input and output and C/N_{in} and C/N_{out} are the carrier to noise ratios measured at the input and output.

Because the interest is on the noise added by the EDFA C/N_{in} is assumed to be infinite. With this simplification the equation becomes

$$F = 10\lg\left(\frac{m^2}{2B\cdot 10^{\frac{C/N_{out}}{10}}}\cdot\frac{P_{in}}{2h\nu}+\frac{P_{in}}{P_{out}}\right). \tag{14}$$

With m=1 and B=1 the ratio between noise and maximum signal output power can be solved for 1Hz noise bandwidth as

$$\frac{1}{10^{\frac{C/N}{10}}} = \left(10^{\frac{F}{10}}-\frac{P_{in}}{P_{out}}\right)\cdot\frac{4h\nu}{P_{in}}\ . \tag{15}$$

This is equivalent to $2\cdot RIN_{EDFA}$. Dividing by two and using Eq. 12 this can be converted to an equivalent noise current in the receiver, which can be given by

$$I_{EDFA} = I_{DC}\cdot\sqrt{\left(10^{\frac{F}{10}}-\frac{P_{in}}{P_{out}}\right)\cdot\frac{2h\nu}{P_{in}}\cdot B}\ . \tag{16}$$

Dispersion in a fibre results in a phase modulation to amplitude modulation conversion (PM/AM). Phase modulation spreads the optical spectrum and the dispersion causes some parts of the signal propagate with higher velocity than the others. This early or late arrival can be seen as a variation in the amplitude at reception. When the phase modulation is due to the phase noise of the optical carrier the resulting variation in amplitude is noise. Therefore, the narrower the laser linewidth is the smaller is the amount of resulting noise. PM/AM conversion is of course a problem in the 1550 nm transmission window but not a problem in 1310 nm window close to zero dispersion wavelength.

The intensity noise induced by the PM/AM-conversion can be estimated from

$$RIN_{PM/AM} = 2\cdot\Delta\omega\cdot\left(\frac{D\cdot\lambda^2\cdot L}{2\cdot\pi\cdot c_0}\right)^2\cdot(2\cdot\pi\cdot f)^2\ , \tag{17}$$

where $\Delta\omega$ is the linewidth of the laser, D the dispersion parameter of the fibre, λ the wavelength of the light, L the length of the fibre, c_0 the speed of light and f the frequency of interest /10/.

Equivalent noise current in the receiver is then

$$I_{PM/AM} = I_{DC} \cdot \sqrt{2 \cdot \Delta\omega \cdot \left(\frac{D \cdot \lambda^2 \cdot L}{2 \cdot \pi \cdot c_0} \right)^2 \cdot (2 \cdot \pi \cdot f)^2 \cdot B} \,. \tag{18}$$

MPI occurs in case of double reflection. Twice reflected light arrives to receiver later than the main signal. This causes interference between the signals. If the double reflection distance is greater than the coherence length of the laser, the interference occurs as noise. Reflections in the CATV networks are typically very small and double Rayleigh back scattering of ~ -30 dBc in long fibres is a dominating effect. In this case the reflections are not discrete but the resulting noise phenomena has same nature. This noise caused by double Rayleigh back scattering is called induced intensity noise (IIN). As Rayleigh scattering is greater in 1310 nm transmission window the IIN is more of a problem in 1310 nm transmission than in 1550 nm transmission.

Approximation to calculate the effect of IIN is given in Eq. (19). Approximation is based on an infinite number of discrete reflections and is valid for light with Gaussian spectral distribution /8/. Therefore, this approximation should not be used with externally modulated transmitters exhibiting SBS suppression

$$RIN_{IIN} \approx \frac{4}{\pi} \cdot \left[\frac{S^2 \cdot \alpha_S^2}{4 \cdot \alpha^2} \left(2 \cdot \alpha \cdot L - 1 + e^{-2 \cdot \alpha \cdot L} \right) \right] \cdot \frac{1}{\eta_{FM} (I_b - I_{th}) \mu} \,. \tag{19}$$

Equivalent noise current in the receiver can be written as

$$i_{IIN} \approx I_{DC} \cdot \sqrt{\frac{4}{\pi} \cdot \left[\frac{S^2 \cdot \alpha_S^2}{4 \cdot \alpha^2} \left(2 \cdot \alpha \cdot L - 1 + e^{-2 \cdot \alpha \cdot L} \right) \right] \cdot \frac{1}{\eta_{FM} (I_b - I_{th}) \mu} \cdot B} \,, \tag{20}$$

where α_S is the proportion of signal scattered per unit length and S is the fraction of scattering that is captured by the fibre. η_{FM} is the light source chirping efficiency in [MHz/mA], I_b and I_{th} are the bias and threshold currents, and μ is the RMS modulation index.

$$\mu = m_T \cdot \sqrt{2} \tag{21}$$

Where m_T = is the total OMI as described in chapter 4.1.1.

4.1.3 Shot noise

Shot noise is a noise component originating from the statistical variation in the arrival of the photons at the receiver. Magnitude of the shot noise is given by

$$I_{SHOT} = 2 \cdot I_{DC} \cdot q \cdot B \quad [A] \tag{22}$$

4.1.4 Thermal noise.

Thermal noise originates from thermal movement in the structure of the matter. Thermal noise power is constant over any given bandwidth, and it only varies with temperature. The noise power can be given as

$$P_{THERMAL\ NOISE} = 4 \cdot k \cdot T \cdot B \tag{23}$$

and equivalent noise current is

$$I_{THERMAL\ NOISE} = \sqrt{\frac{4 \cdot k \cdot T \cdot B}{R}} \tag{24}$$

Calculating the thermal noise of the receiver as suggested in Eq. (24) results in an optimistic approximation as it assumes that no noise is added in amplifier stages of the receiver. CATV receivers are commonly characterised with a parameter noise current density [pA/(Hz)$^{1/2}$]

4.1.5 Carrier to noise ratio

With definition of OMI and the calculated noise factors it is possible to calculate the carrier to noise ratio (C/N). C/N is usually expressed in desibelsand it can be determined for the detected signal in electrical domain and responsivity of the receiver can be left out. C/N can be defined as currents as

$$C/N = 20 \cdot \log\left(\frac{I_{SIGNAL}}{I_{TOTAL\ NOISE}}\right) \quad [dB], \tag{25}$$

where

$$I_{TOTAL\ NOISE} = \sqrt{I_{RIN}^2 + I_{SHOT}^2 + I_{THERMAL\ NOISE}^2 + I_{EDFA}^2 + I_{AM/PM}^2 + I_{IIN}^2} \tag{26}$$

$$I_{SIGNAL} = I_{DC} \cdot OMI \cdot \frac{1}{\sqrt{2}} \tag{27}$$

An example of C/N calculation is given in chapter 5.1.

4.2 Nonlinear distortion effects

Knowing a few parameters, noise effects are easy to calculate quite accurately. This is not the case with distortion. Mathematical treatment of nonlinear effects is possible, but results are not as accurate as in noise calculations. However, calculations can be made to estimate where to look for problems and the calculations are far from obsolete.

Behaviour of an active element can be defined as a polynomial

$$y = Ax + Bx^2 + Cx^3 + \ldots \tag{28}$$

First order term defines the gain and the higher order terms represent distortion induced by the element. Higher than 3rd order terms are seldom of interest as CATV transmission of multiple carriers requires operation in a very linear region. Mathematical treatment of distortion of analogue TV carriers is explained in detail in /11/ and falls outside the interest of this work. However, the phenomenon is explained at a practical level.

4.2.1 CSO and CTB in analogue CATV systems

In a practical CATV network distortion appears as composite second order distortion (CSO) and composite triple beat distortion (CTB). Second order distortion products appear at sum or difference of two individual carrier frequencies

$$f_{CSO} = f_1 \pm f_2 . \tag{29}$$

As carriers are located at frequencies xxx.25 MHz the CSO beats can be found at frequencies 0.75 MHz from the carrier.

Correspondingly, third order distortion products appear at frequencies

$$f_{CTB} = f_1 \pm f_2 \pm f_3 . \tag{30}$$

CTB beats appear at the carrier frequency and due the fact that in many rasters lower frequency channels are spaced by 7 MHz and upper frequency channels by 8 MHz some beats can appear at 1.5 MHz from the carrier, see (Fig. 17).

Knowing the frequency plan, often called the raster, it is possible to calculate the distortion frequencies and the number of beats falling on every frequency. By nature the CSO beats are mostly located near the ends of the raster and the CTB beats in the middle. Beat counts of a common test raster, CENELEC 42chs, are shown in Appendix A. Cenelec raster has only 42 channels and a typical CATV raster with 50 channels (Europe) or 100 channels (USA) generates even more beats.

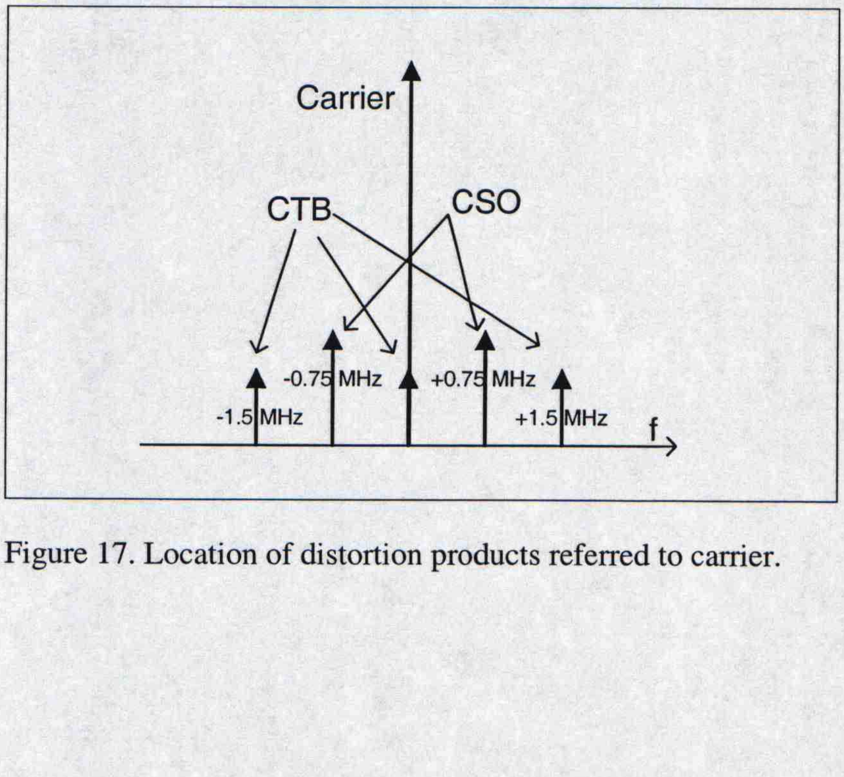


Figure 17. Location of distortion products referred to carrier.

As long as the operation of the device is within the compliance meaning that the device is not clipping the distortion ratio of arbitrary output level can be calculated if the ratio is known at certain output level. As CTB beat is a product of tree carriers, the amplitude of the beat increases 3 dB if output level is increased by 1 dB. Likewise, the change of 1 dB in output level makes a 2 dB change in the level of CSO. Thus 1 dB change in amplitude effects the CSO ratio by 1 dB and CTB ratio by 2 dB as illustrated in the (Fig. 18).

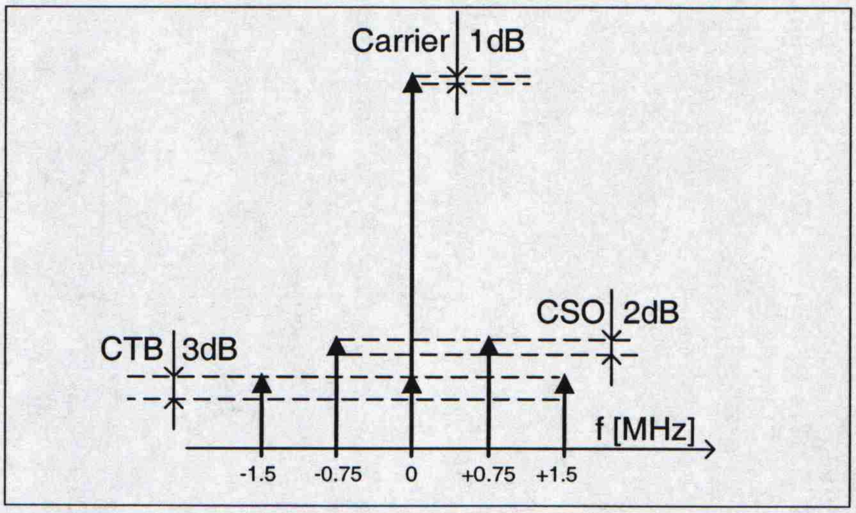


Figure 18. The effect of changing output level on the distortion.

Active devices used in CATV networks are specified by the output level that gives distortion ratio of 60 dB in the worst distortion frequency. From the specification, it is quite simple to calculate the actual distortion performance for the selected output level.

4.2.2 CSO and CTB in digital CATV systems

QAM channels frequency spectrum resembles a block of noise which 3 dB bandwidth is the same as the modulation baud rate. Also the distortion in QAM links is like noise. Randomness of the distortion is quite large as many independently modulated carriers have contribution to it. Still it is not safe to make any separation between noise and distortion induced noise. In the system QAM carriers are equally spaced and all of them are modulated with the same baud rate. It is possible that all the carriers are locked to the same reference or all the modulators can be in phase. Typically C/N needs to be 1 dB higher if it is limited by the distortion induced noise than in case of the pure thermal noise to achieve the same error rate.

For sinusoidal analogue carriers, it is relatively easy to calculate the distortion products in the time domain, using trigonometry. This is not the case with DVB-C signals, but convolution can be used to solve the distortions in frequency domain. In theory, it is possible to link the distortion behaviour of analogue and digital loads together. This could be done by first testing the device against an analogue load, then calculating the corresponding polynomial (Eq.28) and finally estimating the distortion behaviour of the QAM loading by convolution. No attempt to do this is taken in this work. Practice has shown that every analogue raster has to be tested and calculations between rasters only help to make coarse estimates and to locate the difficult spots. One reason for this is that A, B and C in Eq. (28) are not constants, but functions of frequency. The frequency dependency of B and C is not easy to find out. No practical tool for calculation of the convolution is not known to the author and the convolution has to be made with pen and paper.

4.2.3 Chirp and dispersion induced distortion

Distortion caused by chirp and dispersion is the biggest limitation to the number of channels in directly modulated 1550 nm transmitters. Understanding this is essential in succeeding in building DWDM networks with directly modulated 1550 nm transmitters. Use of direct modulation is possible and economical. However the frequencies used must be chosen within one octave bandwidth to avoid the distortion products to fall among the carriers.

Chirp means unwanted frequency modulation of the light signal with the intensity modulation. This phenomenon causes distortion in dispersive fibres, as positive and negative peaks of the signal travel with different speeds in the fibre. An example result of a chirp measurement is shown in Fig. 19. Shape of the transmitted signal is presented after 0 km and after 36 km. A change in the shape of the received signal after 36 km of fibre can be seen. Measured chirp is also presented in the figure.

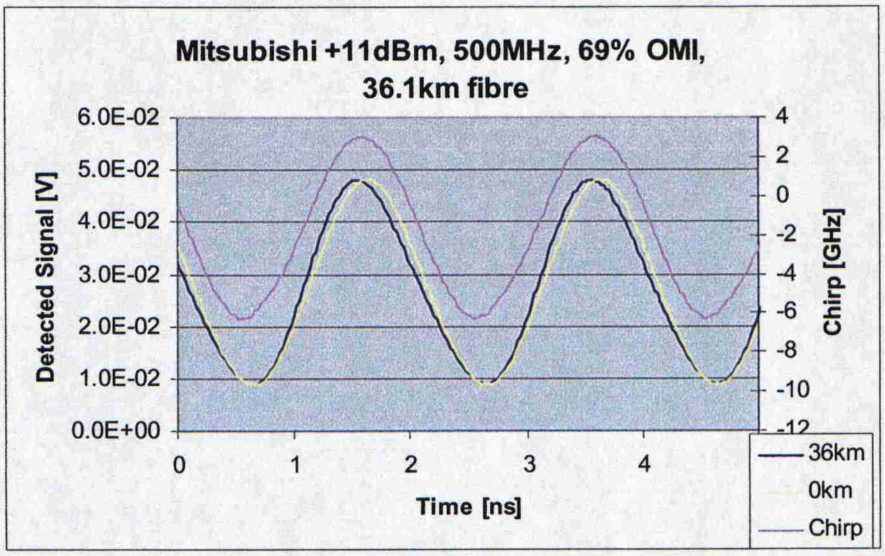


Figure 19. Detected signal after 36 km and 0 km and the chirp in the signal.

Frequency chirp in externally modulated light sources is very small compared to directly modulated sources. In directly modulated lasers the wavelength of the light varies with the changes of the bias current. The chirping effect can be explained by the change in refractive index of the laser chip caused by changes in carrier concentration. The wavelength of the laser is a function of the refractive index of the active area.

Distortion induced by chirp and dispersion is mostly of 2nd order. This is easily understood with an example presented in Fig. 20. Peaks of the signal v(t) have been delayed and the valleys advanced just as in chirp dispersion phenomena, this is plotted with both undistorted sinusoidal signal and the difference between the two. The difference signal is a second harmonic overtone of the original.

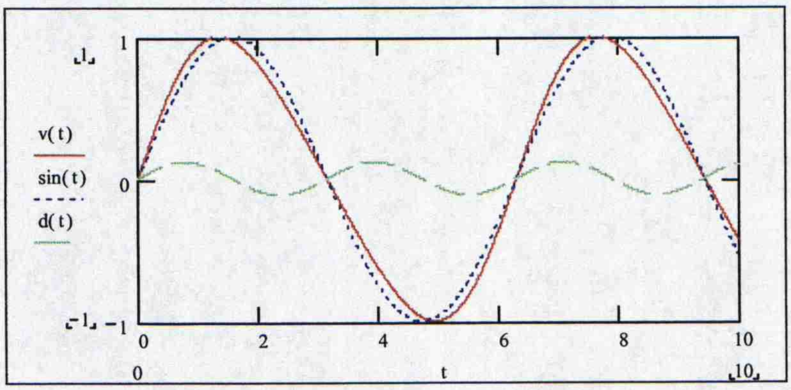


Figure 20. Effect of chirp and dispersion induced distortion. Shift of the phase can be explained by addition of a 2nd harmonic overtone.

In theory chirp can be composed of two parts: adiabatic and transient part. The transient part is proportional to the derivative of the output power, and it is large for the fast changes of the output power. The adiabatic term follows the slow changes of the output power. As adiabatic term is proportional to output power and the transient term to the derivative, there is a 90° phase shift between the two terms of chirp, when modulation is sinusoidal /12/. The chirp can be analytically estimated from

$$\Delta\nu = \frac{\alpha}{4\pi}\left(\frac{1}{P(t)}\frac{d}{dt}P(t) + \kappa P(t)\right), \tag{31}$$

where α is linewidth enhancement factor, *P(t)* is the output power and

$$\kappa = \frac{2\Gamma_f \varepsilon}{V_{act}\eta_{qe}h\nu}, \tag{32}$$

where Γ_f is the optical confinement factor, ε is the gain compression parameter, V_{act} is the volume of the active region, η_{qe} is the quantum efficiency of the laser, h is the Planck constant, and ν is the frequency of the light.

4.2.4 Laser clipping

Laser output power increases quite linearly with bias current. However, if modulation current exceeds the difference between the DC-bias and the threshold current the laser will emit no light for the time that current is below the threshold Fig. 21. This effect is called clipping. In theory 100% OMI can be achieved without any distortion.

Clipping alone is not a severe problem and more than full bandwidth of HFC networks could be filled with 64- or 256-QAM signals if only noise and clipping would be limiting transmission /13/. Soft clipping, before the actual clipping, as threshold is not infinitely sharp, and other nonlinearities limit transmission in real cases, but still with all digital loads it is fare to assume that even a bandwidth of 800 MHz can be filled with 256-QAM. In 1550 nm directly modulated lasers frequency chirp sets the limit to lower channel count in most cases.

In Fig. 21 an example of clipping signal with an OMI of 120% is presented. This means that modulation is increased 20% over the 100% maximum. OMI is defined in chapter 4.1.1, and the definition can be extended to over 100%.

Following the definition, the dependence between the drive power and OMI is no longer linear since clipping adds a DC-component to the signal and the peak-to-peak value only increases with approximately half the rate compared to that of the linear region. Here 120% OMI violates the definition, but is used as it is the linear approximation, and therefore easy to understand. A number of harmonic overtones of a clipped signal are shown in Fig. 22, again for the "120% OMI".

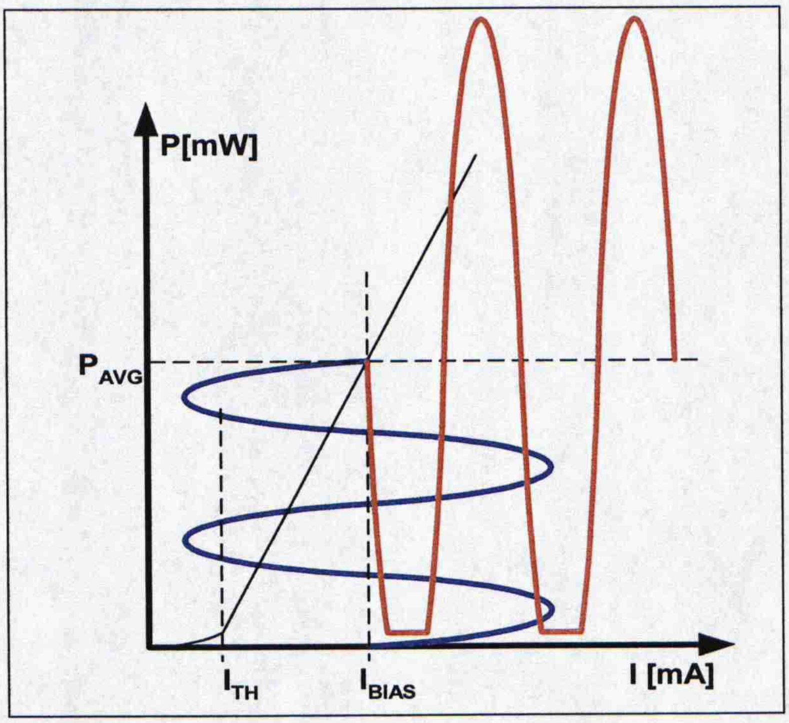


Figure 21. Example of signal clipping in the time domain.

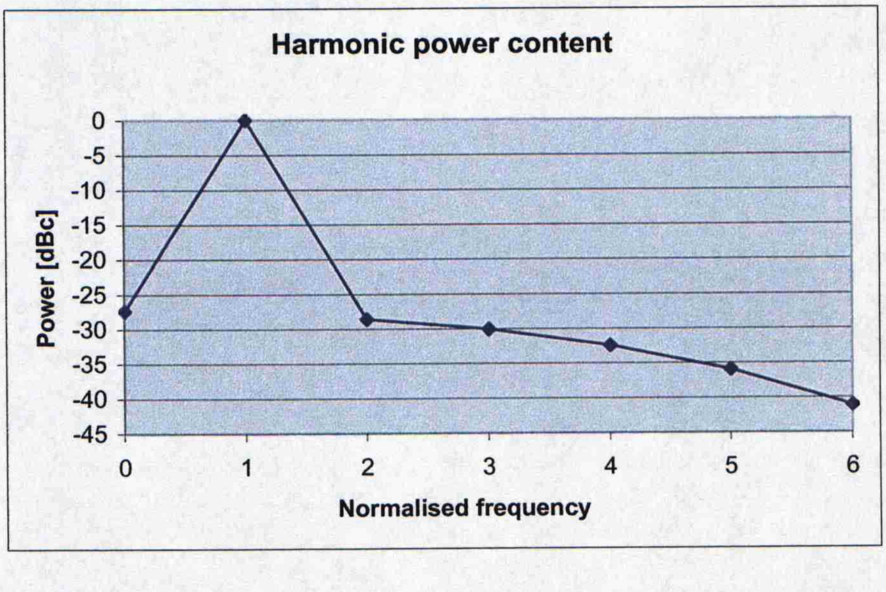


Figure 22. Harmonic power content of the clipped signal having OMI of 120%.

4.3 Optical nonlinearities

The transmission capacity of the fibre is limited by various types of nonlinear effects. These nonlinearities are due to the change of refractive index of the fibre as function of optical power.

4.3.1 Cross phase modulation and stimulated Raman scattering

Cross phase modulation (XPM) is an optical phenomenon where one intensity modulated signal varies the refractive index of the fibre causing phase modulation to other signals in the same fibre.

Stimulated Raman scattering (SRS) causes the optical power to be transferred from lower-wavelength channels to the higher-wavelength channels. Optical power must be present on both channels and in digital systems SRS crosstalk only occurs when there is a “1”-bit on both signals. Amount of Raman scattering is dependent on wavelength difference between the channels and is greater between the channels that are furthest away.

XPM affects mostly the higher frequencies and SRS is a problem in lower frequencies. The effect of SRS is higher but if frequencies used are above 200 MHz this is no longer true. Simulated results where these phenomena can be seen are shown in the figures 23 and 24. Matlab simulation is based on theory of XPM and SRS effects /14 and 15/. Both phenomena are sensitive to the polarisation state, and in this simulation, the worst case is assumed. Powers to fibre used in simulation are the same as in measured case with 8 channels spaced by 1.6 nm at +8 dBm power in chapter 4.3.1. SRS has the strongest effect on the channels that are furthest apart. The Raman-gain increases with the distance between the wavelengths. The crosstalk would be greater with 16 channels with the same spacing. Both phenomena increase crosstalk values by 2 dB with 1 dB increase in optical power. With practical optical power levels and with the frequencies over 300MHz the crosstalk values are better than –50 dB. As the lower frequencies are typically used for broadcast, SRS and XPM don’t limit the narrowcast transmission in a typical situation.

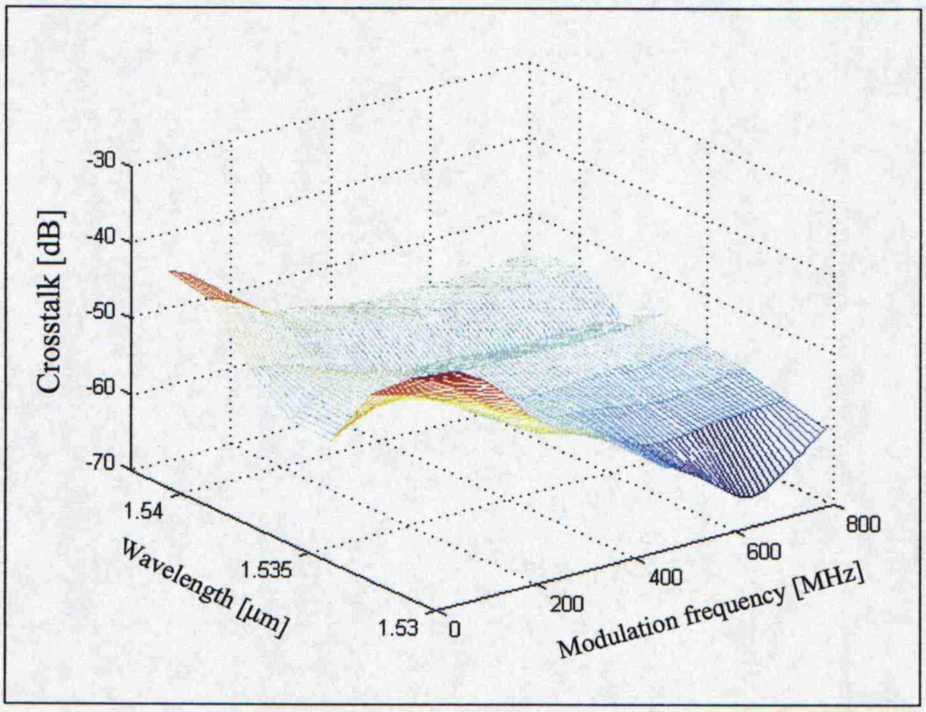


Figure 23. Crosstalk as function of wavelength and modulation frequency.

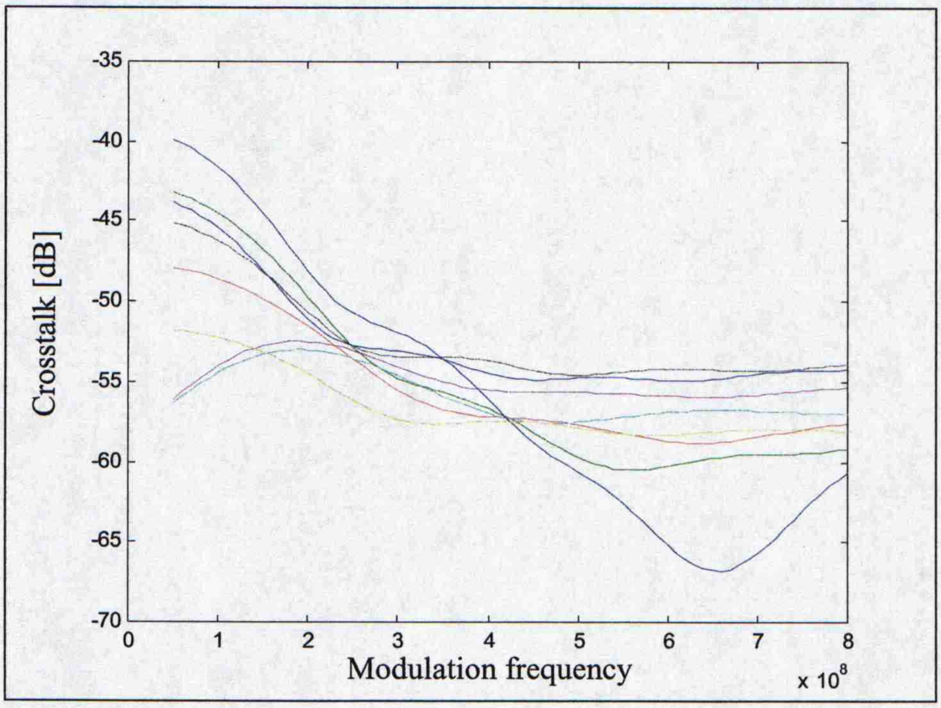


Figure 24. Crosstalk as function of modulation frequency. Projection of the Fig. (23).

4.3.2 Stimulated Brillouin scattering

Stimulated Brillouin scattering (SBS) is a phenomenon where too high spectral density of light causes some light to be scattered. Thus, SBS limits the power that can be inserted into fibre. With broader linewidth laser, more power can be inserted into fibre, but broad linewidth causes distortion due to the dispersion of the fibre. In externally modulated transmitters, the narrow linewidth is intentionally spread to allow higher power to be transmitted.

SBS threshold means the input power at which the reflected power equals the output power of the fibre. The SBS threshold can be calculated from /3,8 and 16/

$$P_{SBS} = 21 \cdot \frac{\zeta A_{eff}}{g_B L_{eff}} \cdot \left(1 + \frac{\Delta\nu_L}{\Delta\nu_B}\right), \tag{33}$$

where the effective length of the fibre is defined as

$$L_{eff} = \frac{1 - e^{(-\alpha L)}}{\alpha} \tag{34}$$

Calculated threshold powers for three different laser linewidths are shown in Fig. 25.

Parameters used are:

$\zeta = 2$, polarization state parameter. 2=random polarization state.

$A_{eff} = 85 \cdot 10^{-12}$ m^2 , Effective area of the fibre

$g_B = 4.6 \cdot 10^{-11}$ m/W Brillouin gain coefficient.

$\Delta\nu_L$ = Broadened effective laser linewidth

$\Delta\nu_B$ = SBS gain linewidth (20MHz)

$\alpha = 0.046$ Fibre attenuation coefficient, the value equals 0.2 dB/km.

L = length of the fibre

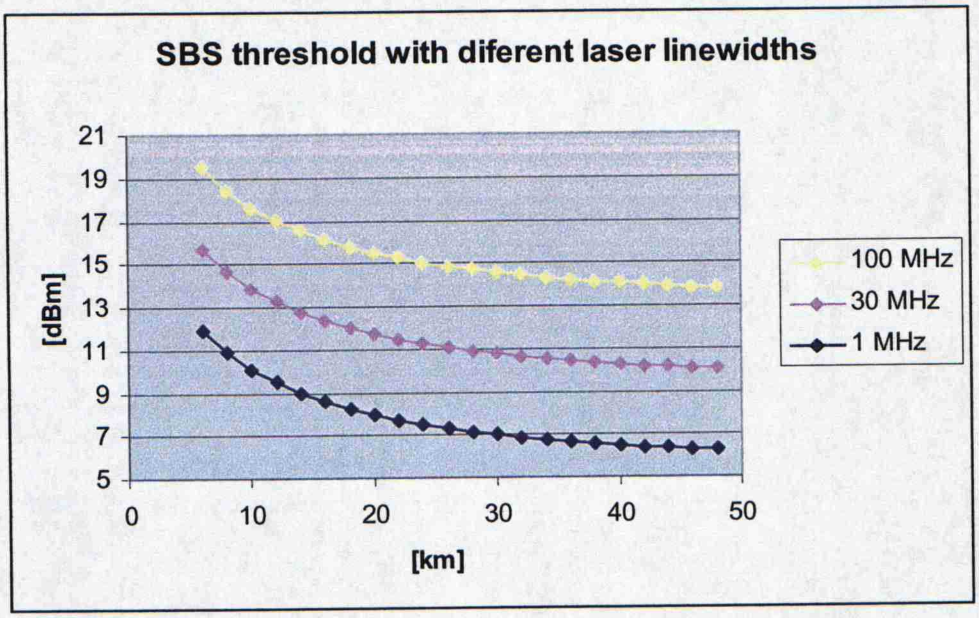


Figure 25. Calculated SBS thresholds with three different laser linewidths.

4.3.3 Self-Phase Modulation (SPM)

Just as in cross phase modulation, an intensity modulated signal varies the refractive index of the fibre causing phase modulation to itself. This is the same phenomenon, but appears as phase modulation to the signal itself and as crosstalk to other signals. No measurements are presented in this work, as this is not a practical problem. The chirp induced CSO distortion has a dominating effect on signal quality.

4.3.4 Four-Wave Mixing (FWM)

Four way mixing is analogous to CTB distortion. Large power light carriers interfering with each other cause some energy of the light to be transferred to other wavelengths.

$$f_{FWM} = f_1 \pm f_2 \pm f_3 . \tag{35}$$

In DWDM systems wavelengths are typically equally spaced which means that FWM products will fall on the other channels. FWM however needs a constant optical phase over relatively long span and therefore is not a problem in standard single mode fibre. Use of dispersion shifted fibre can not be recommended because of FWM in DWDM links.

4.4 Frequency response effect of dispersion.

Amplitude modulation produces side bands to optical carrier on both sides at the distance of modulation frequency. These side bands carry the modulation energy, but as they differ in frequency they travel in fibre with different velocities, and at some point the energy of side bands will be out of phase. This will be seen as a zero in frequency response. Frequency response can be estimated by Eq. 36, /17/

$$H(f_{\mathrm{mod}}) = \cos\left[\pi LcD\left(\frac{f_{\mathrm{mod}}}{\nu}\right)\right], \tag{36}$$

where D is the group velocity dispersion parameter 17ps/nm km in SMF @1550 nm, c is the velocity of light, L is the fibre link length, f_{mod} the modulating frequency and ν the frequency of light. Fig. 26 shows a graph of Eq. (36) for link length of 100 km. From the graph, it is easy to see that this phenomenon limits the frequency response only at very high frequencies above 5 GHz.

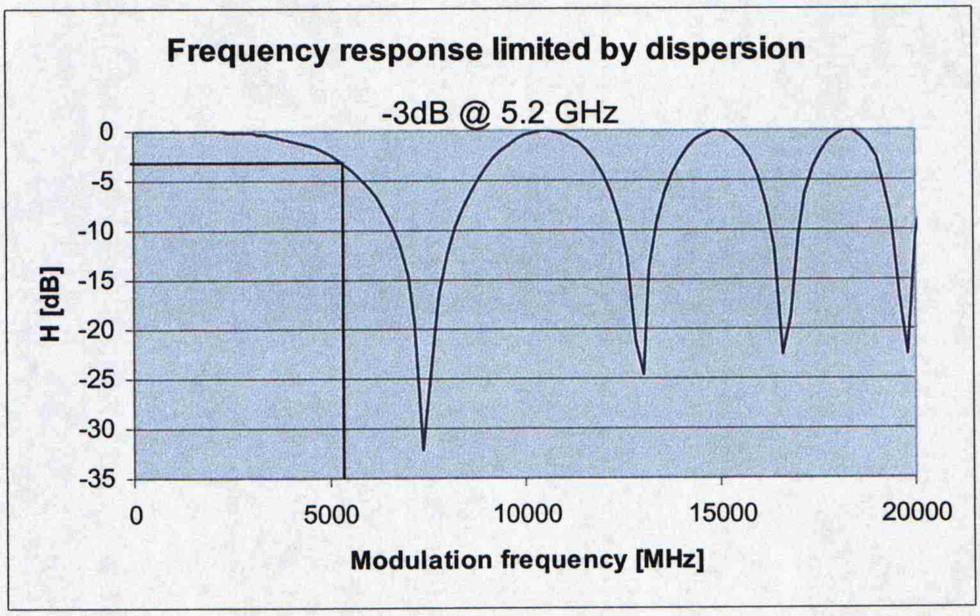


Figure 26. Dispersion limited frequency response of 100 km of single mode fibre.

In externally modulated system, an additional phase modulation is used to spread the spectrum for SBS suppression. This is usually done with two tones at approximately 2 and 3 GHz. This in combination with amplitude modulation causes an effect similar to the one seen in Fig. 26, but the effect can be seen in lower frequencies as well. In direct modulation, the chirp is a dominating effect and this kind of linear distortion cannot be seen.

4.5 Measurement of CATV signal quality

Measurement of the distortion is not easy because of the high dynamics required. Distortion products are typically at least 60 to 65 dB lower than the carriers, and spectrum analyser front end is not as linear device as the measured ones. For this reason band pass filters need to be used to ease the loading of the spectrum analyser. On the other hand, the level into spectrum analyser must be high enough. Practical resolution bandwidth (RBW) for the measurement is 30kHz and this is suggested in /18/ as well. Noise floor of the analyser is typically 7…8 dBμV with 30kHz RBW indicating that minimum measurable level is around 10 dBμV. In order to measure 70 dB distortion ratios at least 80 dBμV should be coming into analyser.

Example of CTB and C/N measurement is presented in Figs. 27 and 28. Level of the carrier has been measured to be 97.5 dBμV. After being measured this individual carrier has been turned off and composite third order distortion can be seen in Fig. 28. Level of distortion is 30.7 dBμV and thus distortion distance is 66.8 dB. C/N is measured at a frequency where no distortion is present. In Fig. 27 the result 46.3 dBμV has been measured with noise marker function and the result given for 4.75 MHz video bandwidth directly. Quite typically, spectrum analysers measure noise only to 1Hz bandwidth. This can be corrected with a factor

$$B_{correction}[dB] = 10 \cdot Log_{10}\left(\frac{Bwanted}{Bmeasured}\right) \tag{37}$$

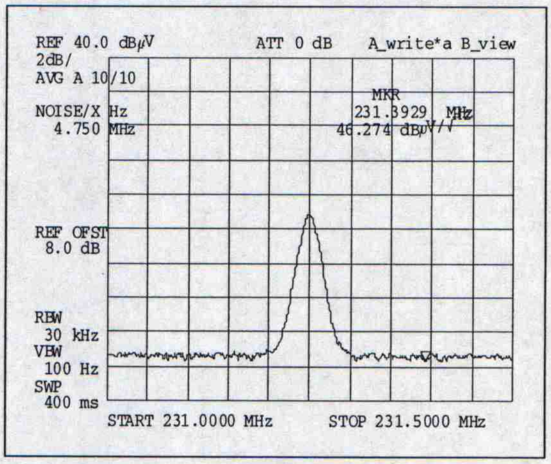


Figure 27. Example of the C/N measurement.

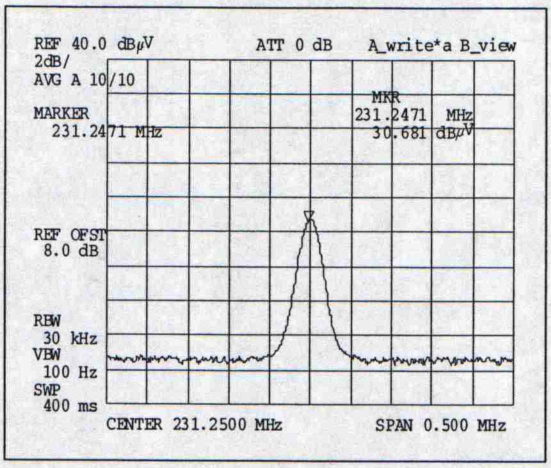


Figure 28. Example of the CTB distortion measurement.

When measuring noise power with spectrum analyser the noise marker should be used and this is valid for measuring the power of the digital channel as well. Peak-to-average ratio is different for noise and sinusoidal carriers and for this reason a correction is needed since the measurement is typically done with peak detector. The correction needed is 1.05 dB. Logarithmic amplifier makes this error larger yet by 1.45 dB. Equivalent noise bandwidth of resolution filter is also different from the ideal. Correction for this can vary but is, for example, -0.8 dB for Advantest R3261A spectrum analyser. The noise marker function automatically makes these corrections.

4.5.1 Measurement of a QAM channel power

Power of a QAM channel is quite evenly spread over a bandwidth of the modulating baud rate. The baud rate or the signal bandwidth typically used in DVB-C transmission in Europe is 6.875 MHz. Noise power density is measured on the flat region in the middle of the channel using the noise marker function as explained in Chapter 4.5 and then multiplied by the channel bandwidth. Large RBW (300kHz – 1 MHz) should be used to smooth the ripple of the channel. Guidelines for measuring DVB signals are given in /19/.

4.5.2 QAM in channel signal quality

QAM signals like all digital signals are judged by bit error ratio BER. In DVB-C transmission Reed Solomon coding is used to detect and correct errors. Reed Solomon coding of the signal ensures BER of 10^{-10} as long as BER is better than 10^{-4} without correction. BER of 10^{-10} means approximately one error in 24 hours per video stream, which is considered quasi error free. The BER or the probability of an error caused by noise in the reception can be estimated with /20/

$$P_M = 2 \cdot \left(1 - \frac{1}{\sqrt{M}}\right) \cdot erfc\left(\sqrt{\frac{3}{2 \cdot (M-1)} \cdot k \cdot \gamma_b}\right) \cdot \left[1 - \frac{1}{2} \cdot \left(1 - \frac{1}{\sqrt{M}}\right) \cdot erfc\left(\sqrt{\frac{3}{2 \cdot (M-1)} \cdot k \cdot \gamma_b}\right)\right] \tag{38}$$

where M = number of constellation points, k = number of bits per symbol, γ_b = signal to noise ratio per bit.

Fig. 29 shows calculated bit error ratios for 16-, 64- and 256-QAM as function of C/N.

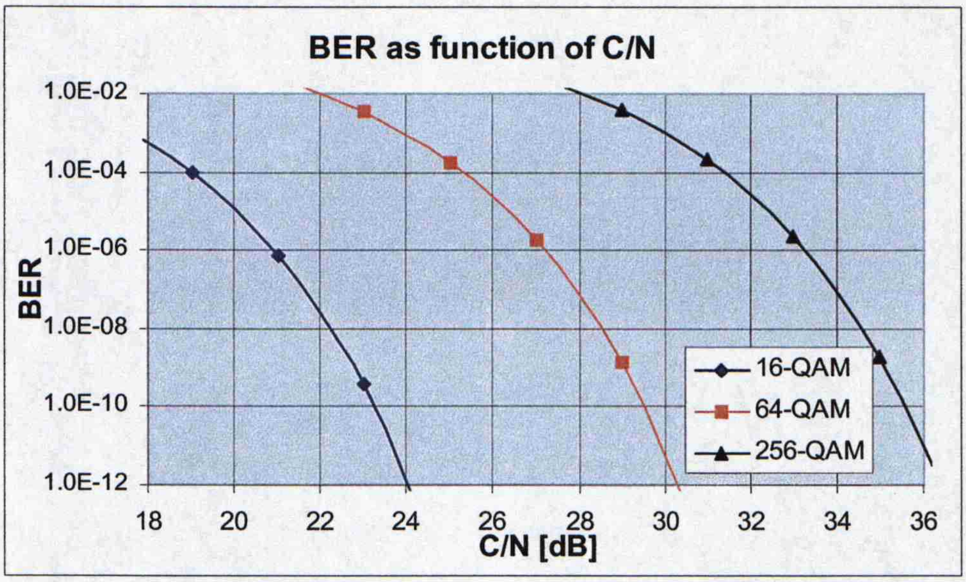


Figure 29. BER as a function of C/N calculated with Eq. (38).

DVB-C transmission is in digital format and QAM modulation is used. Typically, 64-QAM is used but 265-QAM may also be applied. For other than video transmission purposes 16-QAM or QPSK may be used. This work concentrates on 64-QAM but almost anything can be applied to other levels of QAM too if the different C/N requirement can be full filled. As changing from 64-QAM to 256-QAM means halving the level difference between different symbols, 256-QAM needs 6 dB more C/N than 64-QAM. In a similar way 16-QAM will survive with 6 dB less C/N. These theoretical figures assume that Gaussian noise is the only limiting factor in transmission. However, modulated signal is not ideal and this factor cannot be ignored.

Modulation error ratio (MER) describes the purity of the signal in detection. Besides the Gaussian noise, MER includes all other impurities of the received constellation that can not be corrected by the receiver. MER is defined to be the ratio between the ideal signal vector and the average error vector. This ratio is often presented in percents or in decibels in Eqs. (39) and (40). Figure 30 helps to understand these equations.

$$MER = 100 \cdot \sqrt{\frac{\sum_{i=1}^{n} S'^{2}_{i}}{\sum_{i=1}^{n} e_i^2}} \, [\%] \qquad (39)$$

$$MER = 20 \cdot \log \cdot \sqrt{\frac{\sum_{i=1}^{n} S'^{2}_{i}}{\sum_{i=1}^{n} e_i^2}} \, [dB] \qquad (40)$$

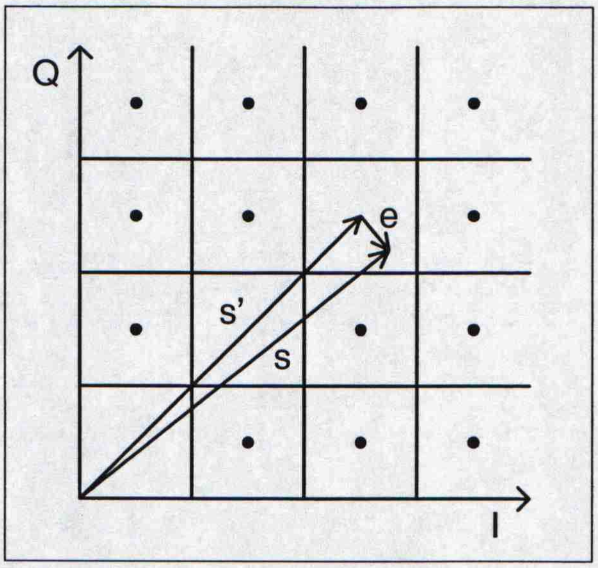


Figure 30. One quadrant of 64-QAM constellation diagram showing the ideal signal vector, error vector and the actual signal vector.

The measurement DVB standards committee for ETS 300-429 suggests using MER as a single figure of merit for in channel performance of the entire link. This makes sense since any kind of signal impurity effects MER figure. With low C/N values, MER and C/N are equal. With higher C/N values, the MER is limited by spurious tones, or by the modulator.

To ensure fair judgement between different parts of the link MER should be used to evaluate the modulator only and the link itself should be judged by C/N and distortion distance.

MER measurement is easy as it can be done without disturbing the payload and can be used for monitoring purposes. Evaluation of C/N requires turning of the channel measured unless an empty channel space is close enough to assume having the same noise power.

Signal quality can be estimated by looking at the constellation diagram with a test receiver. Some examples of constellation diagrams made with Rohde & Schwarz WinIQSIM™/21/ software are presented below. In the ideal constellation all points are exactly in the middle of the squares as shown in Fig. 31. In Fig. 32 phase noise of either modulator or receiver has rotated the points. An interfering carrier causes constant amplitude error with rotating phase as displayed in Fig. 33. Noise causes random error with a Gaussian distribution illustrated in Fig. 34. Earlier, in Fig. 9, a constellation with noise and phase noise was presented. Constellation of the received signal may look like the last example and still deliver quasi error free video stream.

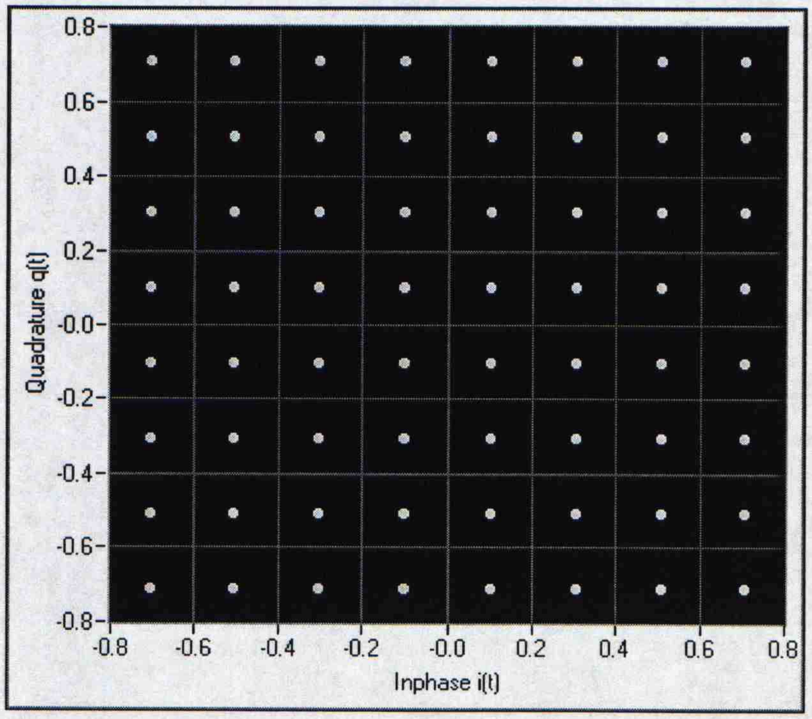


Figure 31. Ideal 64-QAM constellation diagram.

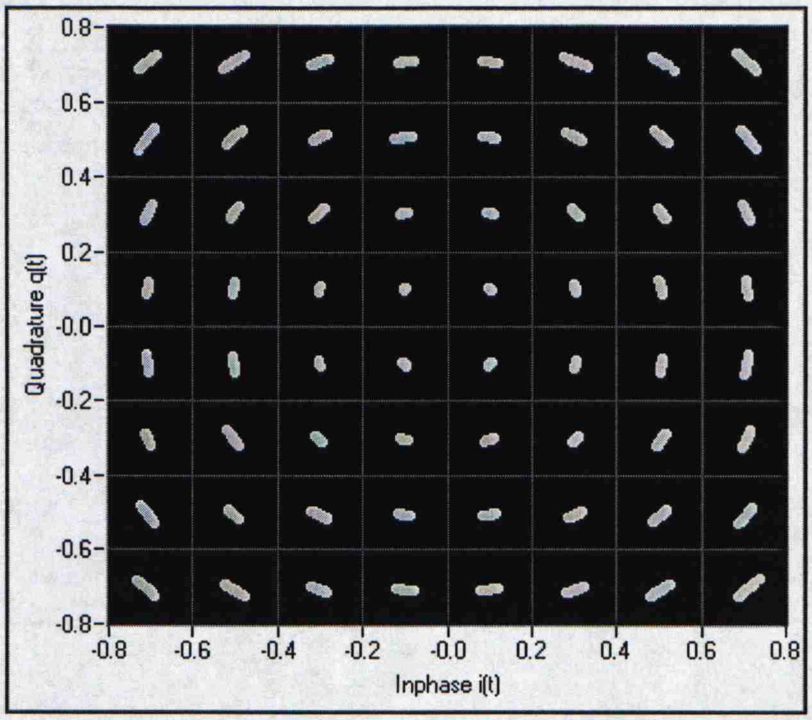


Figure 32. 64-QAM constellation diagram with phase noise.

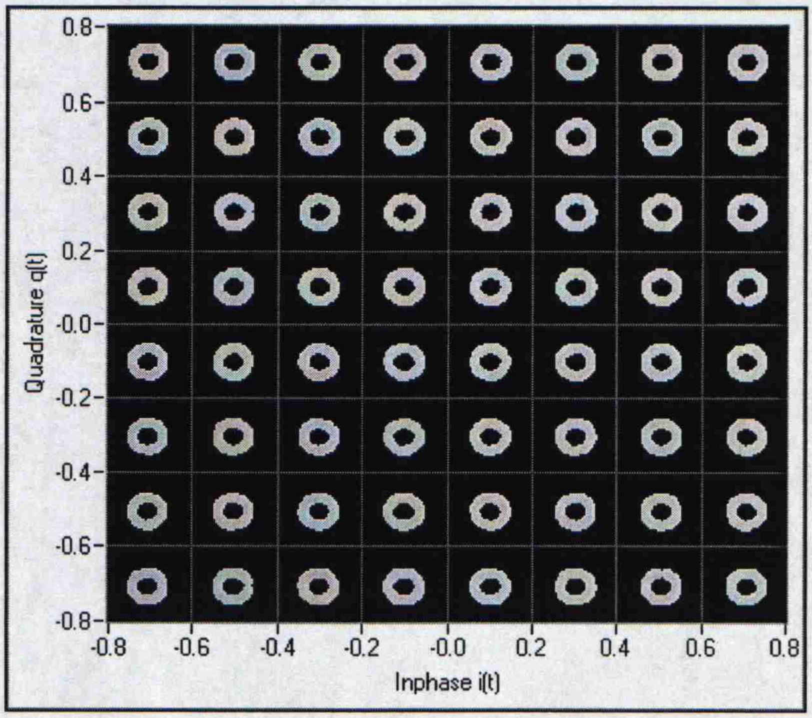


Figure 33. 64-QAM constellation diagram with interfering carrier.

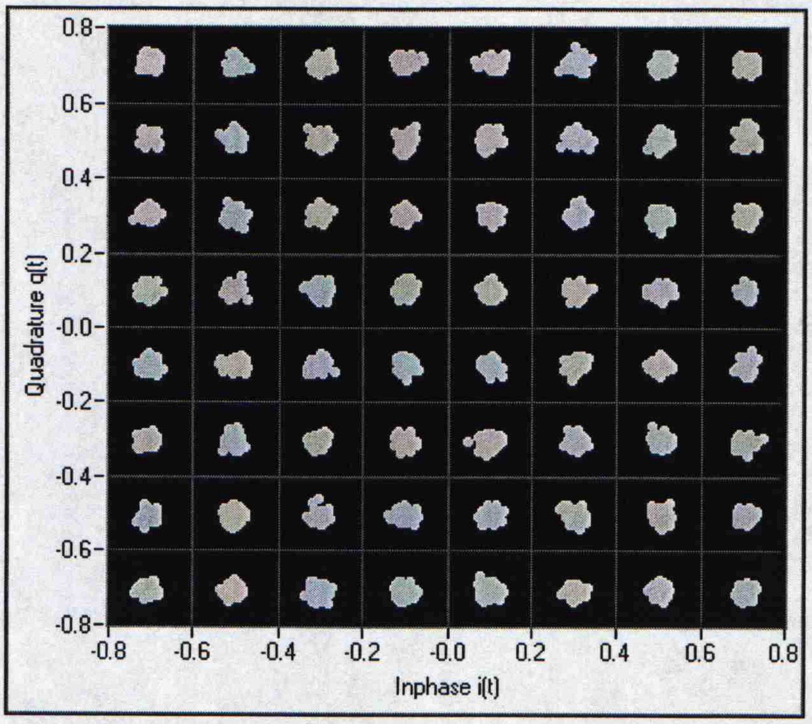


Figure 34. 64-QAM constellation diagram with Gaussian noise.

5 MEASUREMENTS OF THE NETWORK COMPONENTS

5.1 RIN of the Laser

The fact that the RIN of the laser limits the maximum achievable C/N in the link, can be applied in a simple method of measuring the RIN. As power launched to the receiver determines the best measurable RIN value, the receiver should be able to withstand high powers. Both coupling to the amplifier and the noise figure of the amplifier should be good. RIN of the lasers used in this work was specified to be better than -157 dBc at 10 dBm output power (see appendix D).

The RIN of the laser chip was measured with a good CATV receiver that can tolerate 6 dBm input power. C/N was measured at various input powers with constant OMI of 2% and the RIN value was fitted to match the measured data (Fig. 35). Noise current density of the receiver was fitted as well, but this parameter affects the C/N value at low input powers. C/N equations used are introduced in chapter 4.1. Measured RIN of the laser was -161 dBc at 6 dBm output power. Noise current density of the receiver used was 7 pA/(Hz)$^{1/2}$.

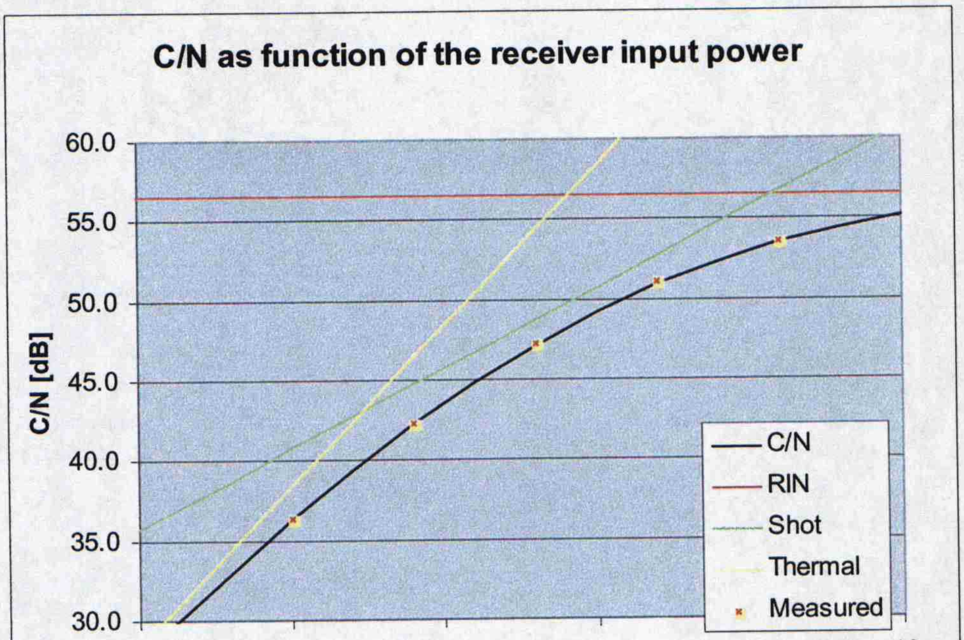


Figure 35. Measurement of laser RIN.

5.2 Maximum Loading of the laser

As output power of a laser chip is limited, the only mean to increase the C/N in a given link is to use higher OMI. In other words, to use higher modulation. Modulating power can only be increased to a point at which the distortion is still acceptable. Maximum loading depends of course on what is considered acceptable quality. As distortion is a nonlinear effect the deterioration of the signal quality is quite fast after some point and typically total OMI is around 40% at that point. Maximum loading of the laser is valid only for very short links using 1550 nm directly modulated lasers. In longer links interaction between chirp and dispersion sets the limit for the loading of the laser. Figure 36 shows a measured C/N of the transmitter. The CTB limited C/N is 45 dB. This value is good enough for most applications.

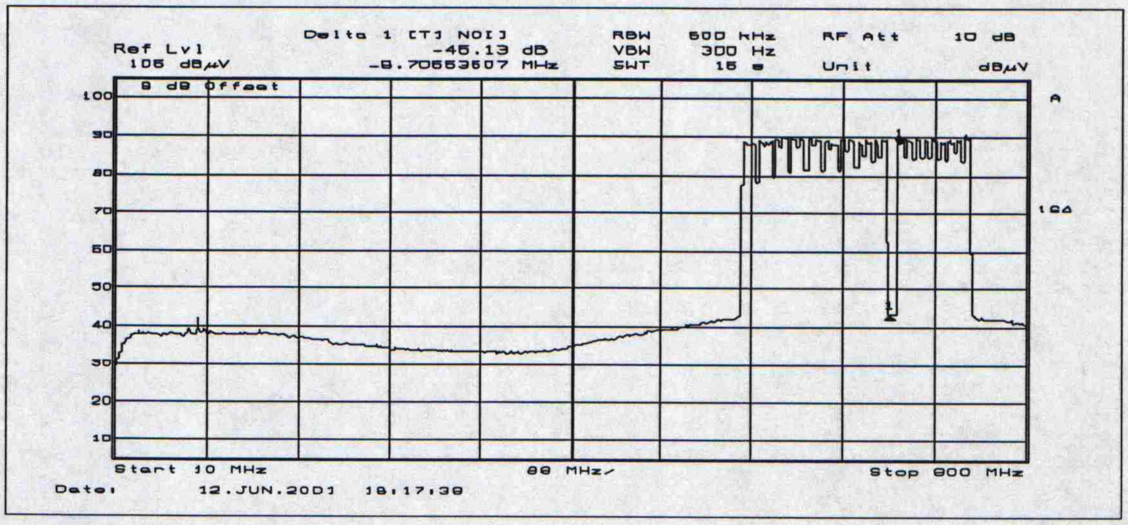


Figure 36. 33 QAM channels at 6% OMI. Total OMI is 39%.

5.2.1 Effect of the drive amplifier

Not all of the distortion is necessarily doe to the laser; some of it may be caused by the drive amplifier. The amplifier used in tested transmitters has very little effect to signal quality. Fig. 37 shows measurement of laser test point C/N. Measured result 50.1 dB needs to be corrected, as noise floor of the spectrum analyser is only about 5 dB lower than measured noise. C/N of the QAM signal coming to transmitter is less than 55 dB. Noise limited C/N of the drive amplifier is around 55 dB as well. With these extra noise sources, it is fair to say that C/N of the drive amplifier is not limited by CTB distortion at all.

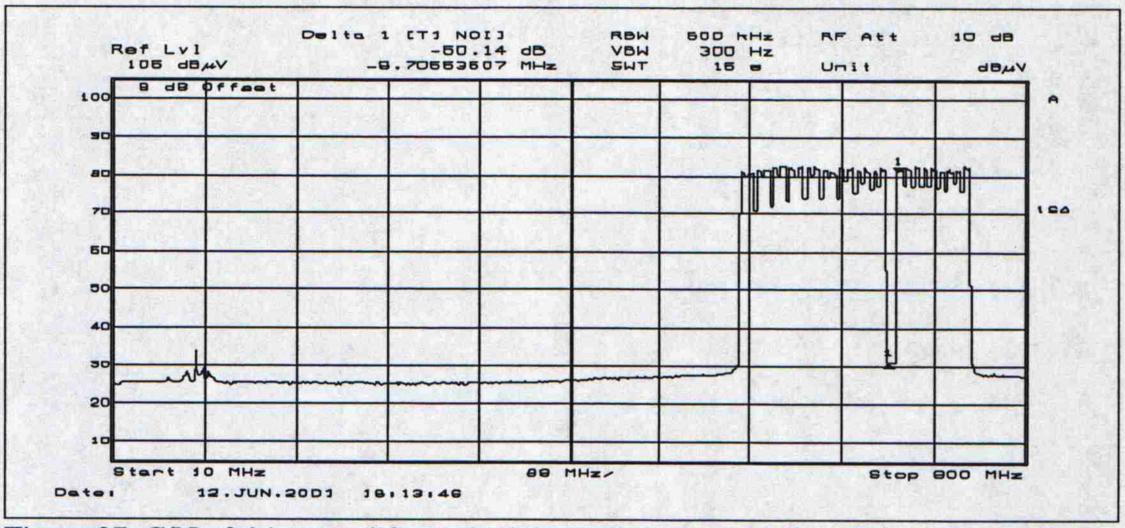


Figure 37. C/N of drive amplifier, which is much better than that of the laser.

5.3 Chirp

Chirp measurements of the laser transmitters were done with a chirp analyser /22/. The principle of the chirp measurement is clarified in Fig. 38. Modulated light of the laser is fed to a fast detector through a Fabry-Perot (F-P) etalon and then measured with sampling oscilloscope. F-P etalon has a wavelength dependent periodic transmission loss, which can be tuned with temperature. A measurement result of the transmission loss is shown in Fig. 39. By scanning the temperature a minima and a maxima of the transmission are located. Based on this information two temperature points where transmission is equal but the slope is opposite can be found. As the measurement of the same input signal is done both on negative and positive slope of the response, the chirp can be determined. The average of the two measurements is similar to the original transmitted signal, and the difference of the two measurements divided by the sum is proportional to the chirp /22/. Therefore, this system limits measurements to non-clipping signals as clipping causes division by zero.

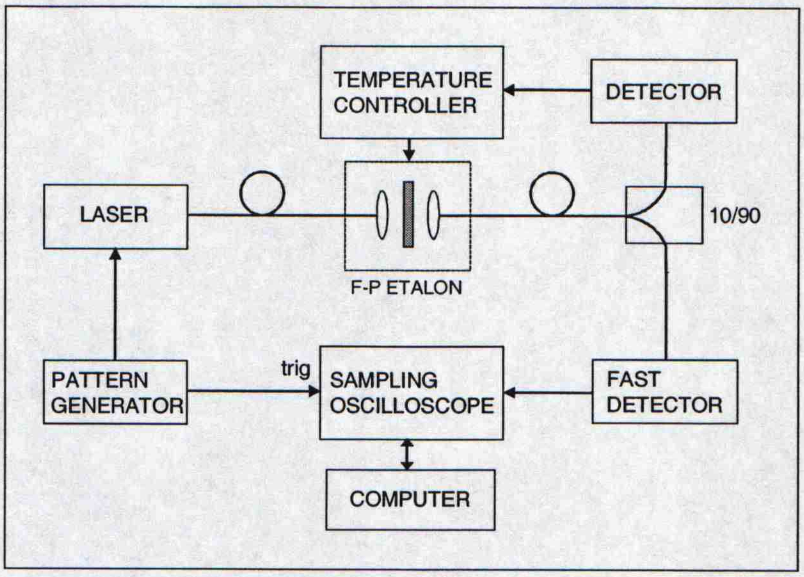


Figure 38. Principle of the chirp measurement /22/.

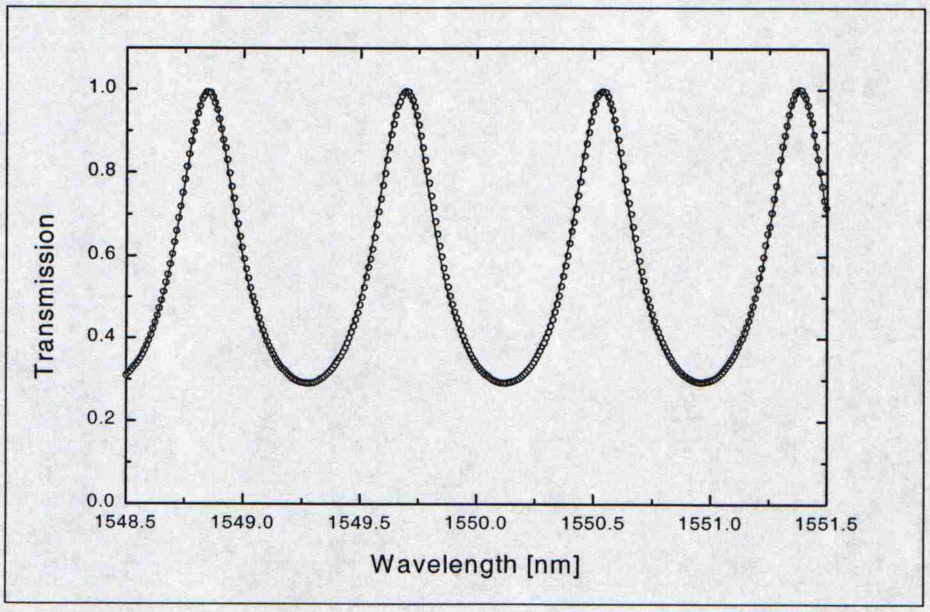


Figure 39. Transmission of the F-P etalon as the function of wavelength /22/.

Figures 40 - 44 represent the measured frequency chirps of several DWDM laser chips manufactured by Mitsubishi, Lucent, JDS and Sumitomo. All of the lasers are biased at same output power and have a similar load. No significant difference in chirp between the chips can be observed. Shape of the chirp curve in Lucent laser is different from the others but amplitude of the chirp is no larger. JDS CFQ933 is the only one especially meant for QAM transmission and one of the features it is advertised by is the low chirp. However, chirp of the laser is not smaller than in the rest of the diodes.

All the other chips are meant for 2.5Gbit/s digital transmission where low chirp is also important in achieving long transmission distances and therefore the chirp is as low as possible. Chirp is quite commonly expressed in terms of MHz/mA or as a dispersion penalty. The latter means the excess power needed to achieve the same bit error rate compared to measurement without dispersion. This specification is only meaningful in baseband transmission, but smaller value is of course better for any case where dispersion is present. Chirp given in MHz/mA is quite informative way of specifying the phenomena but this is not a constant value and changes as function of frequency and modulation index.

Typically the threshold current of a laser is 15 mA at 25 °C and typical bias current at 5 dBm is 35 mA, the difference 20 mA corresponds to peak value of modulating current for 100% OMI. Peak-to-peak value of the chirp (adiabatic) in Figs. 40 - 44 is very close to 3GHz in each, which is the same as peak value of 100% OMI would be. From this we can calculate that the adiabatic chirp is ~ 150 MHz/mA.

As only one chip per type was typically available it would be unfair to judge which of the lasers is the best. Differences between two chips of the same type might be larger than in results measured here.

The effect of modulation frequency can be seen in Figs 44 - 46. At 100 MHz the chirp compared to modulation is smallest and the phase shift between the signal and the chirp is smallest as well. Both of these factors get larger as modulation frequency increases. Increase is not large but it can be seen. Chirp can be expected to be slightly larger at higher frequencies.

In Fig. 47 the effect of the large OMI can be seen. Laser is not clipping but at the minimas of output power the operation of laser is no longer stable and a oscillation transient in the chirp can be seen. In Fig. 48 the OMI is larger yet and this oscillation can be seen in output power as well. Oscillating frequency is about 5GHz and this is the relaxation oscillation frequency of the laser.

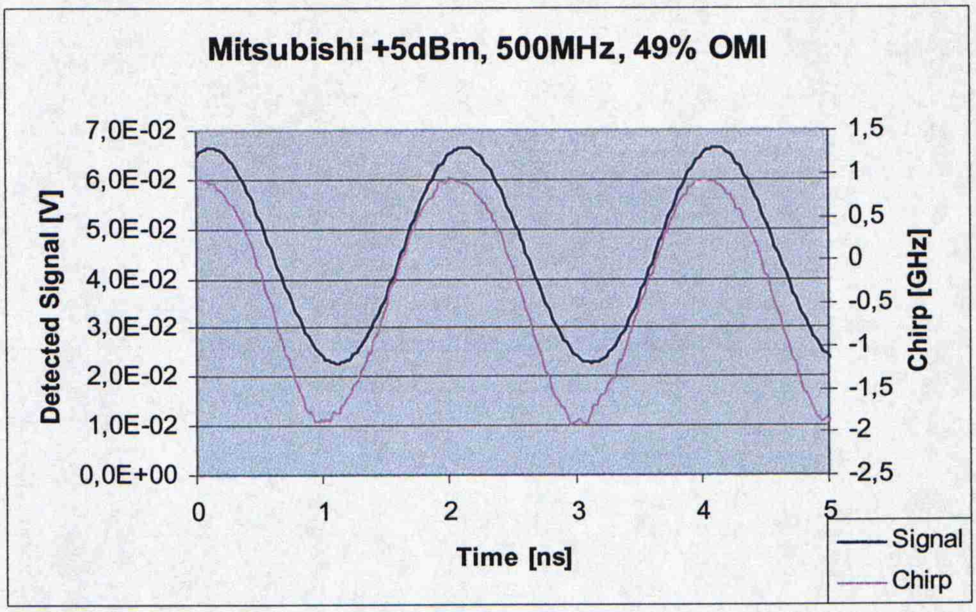


Figure 40. Chirp measurement result for Mitsubishi laser.

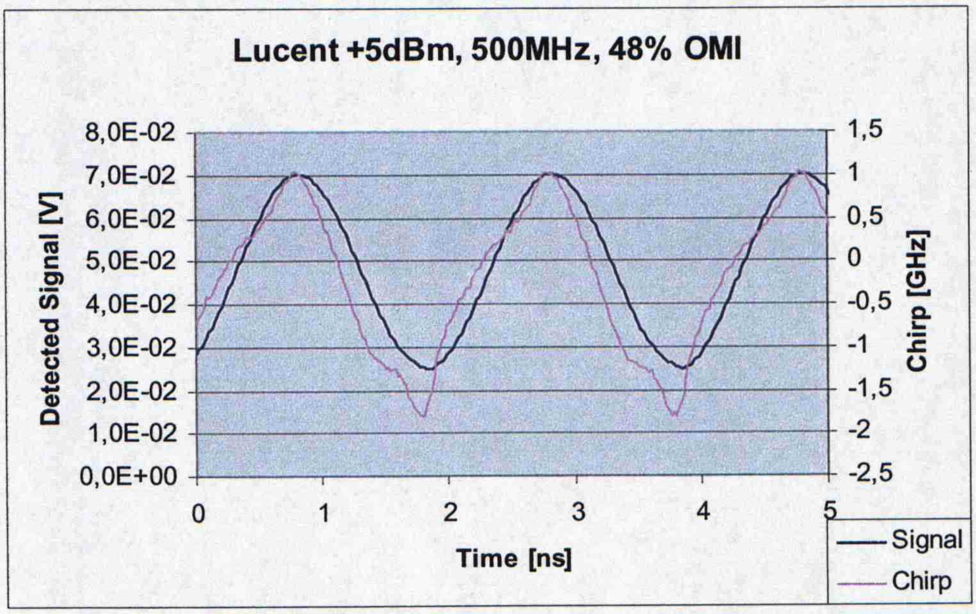


Figure 41. Chirp measurement result for Lucent laser.

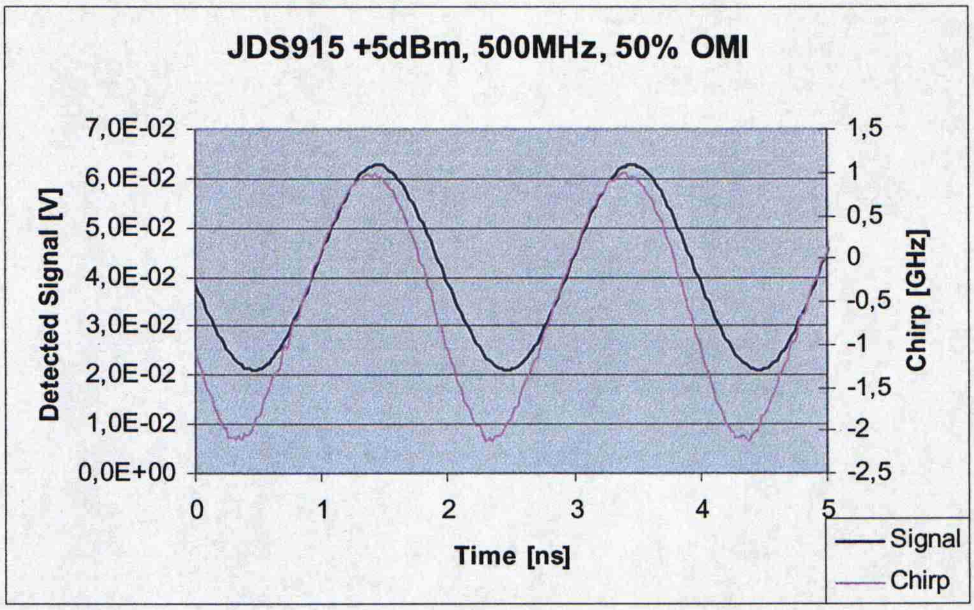


Figure 42. Chirp measurement result for JDS 2.5 Gbit/s laser.

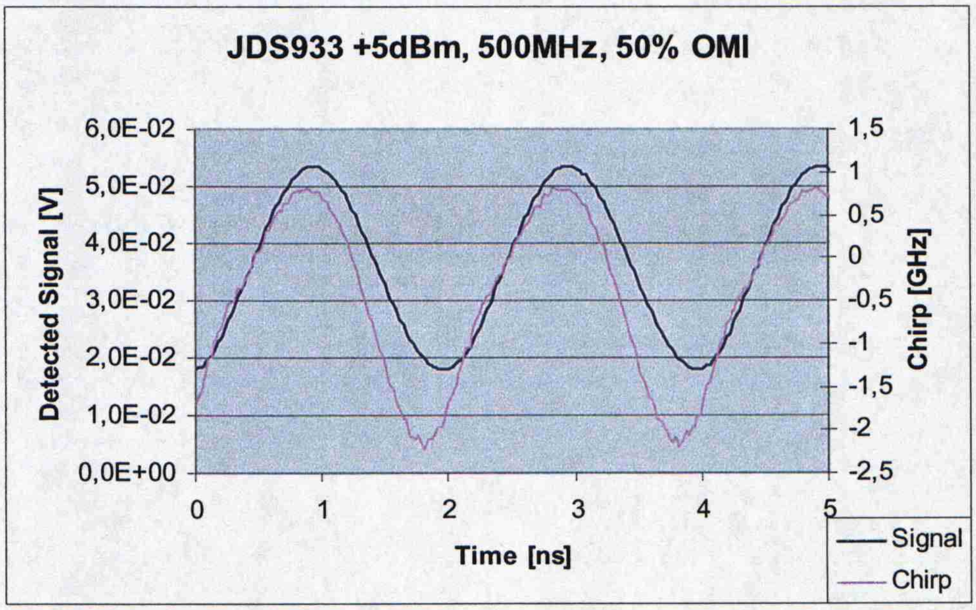


Figure 43. Chirp measurement result for JDS QAM laser.

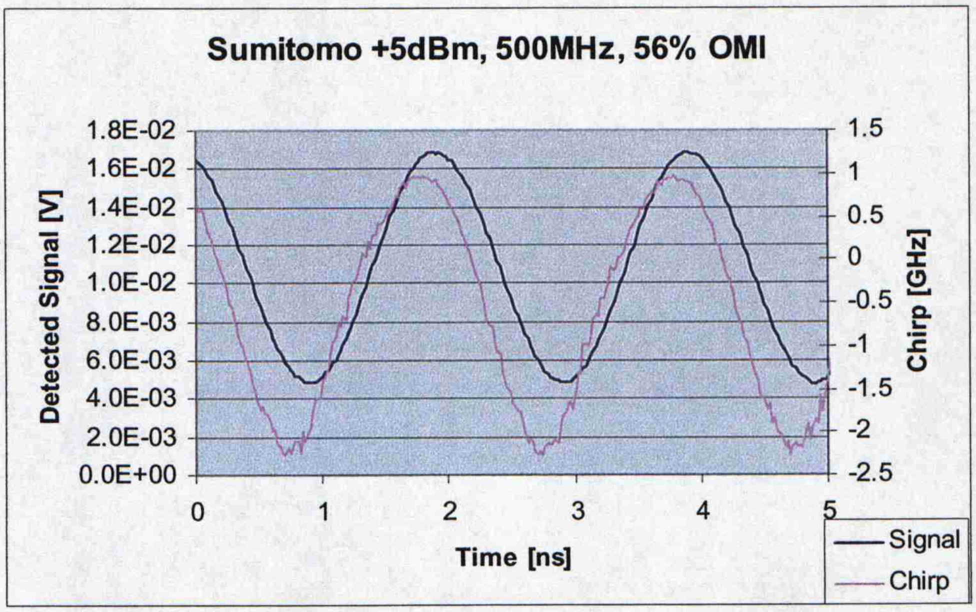


Figure 44. Chirp measurement result for Sumitomo laser.

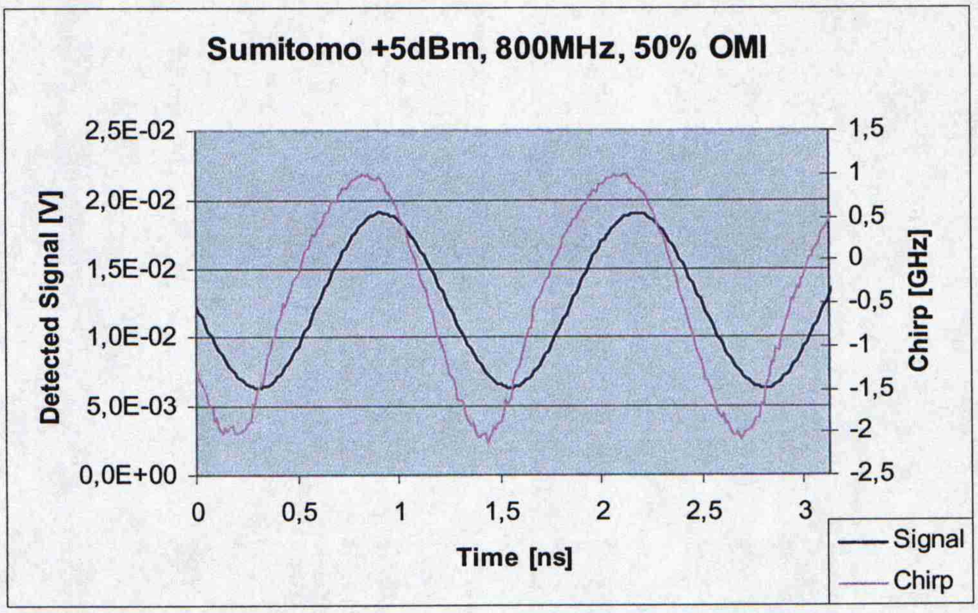


Figure 45. Chirp measurement result for Sumitomo laser at higher frequency.

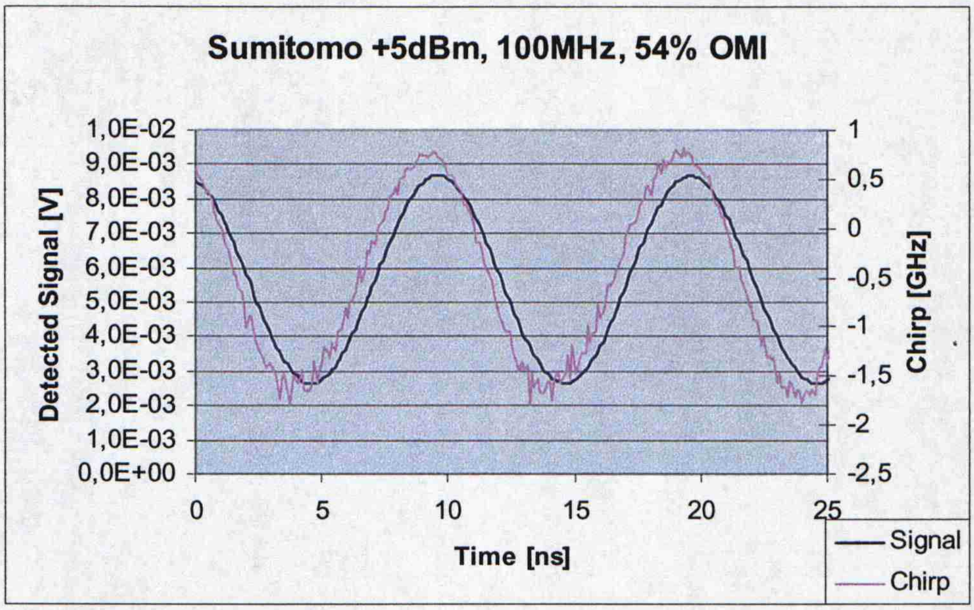


Figure 46. Chirp measurement result for Sumitomo laser at lower frequency

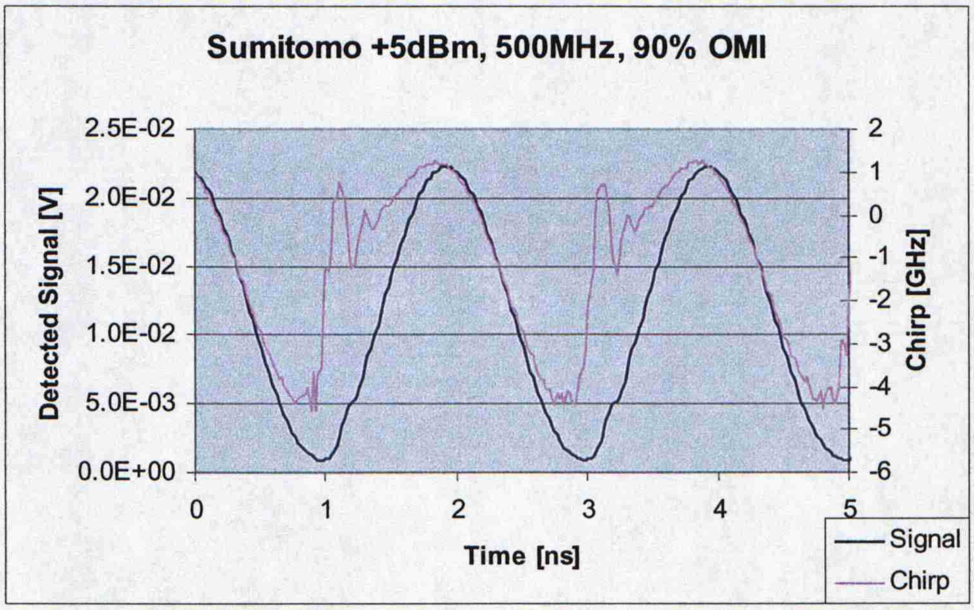


Figure 47. Chirp measurement result for Sumitomo laser at high OMI.

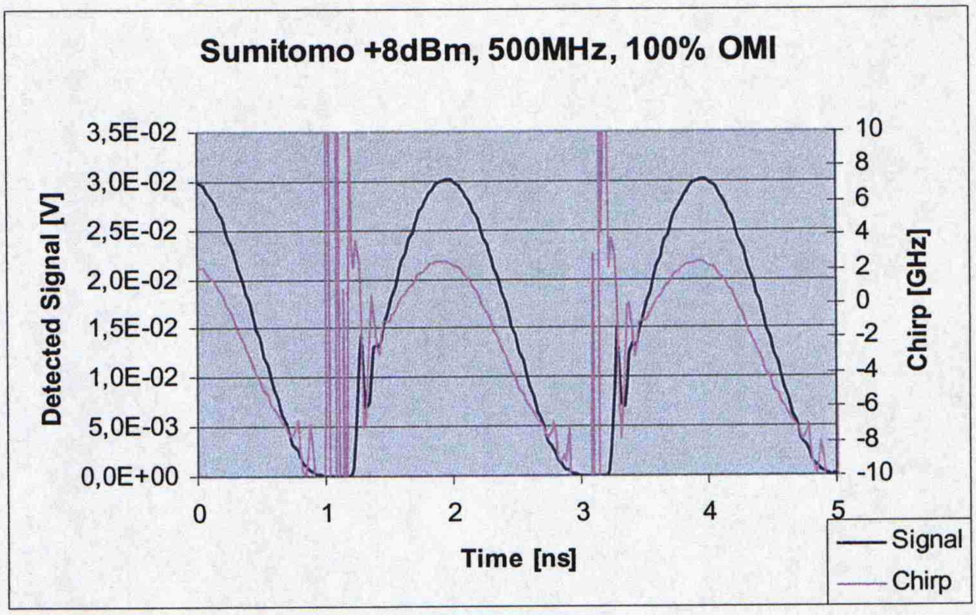


Figure 48. Chirp measurement result for Sumitomo laser at very high OMI demonstrating the limitation of the measurement system to non-clipping signals.

5.4 Cooling current of the laser over temperature

Wavelength of the DWDM laser is fine tuned to the exact channel wavelength with temperature. Temperature of the laser chip is tuned and held constant with thermo electric cooler (TEC). The TEC can cool the chip about 45 degrees. Warming is much easier, and larger differences can be achieved. Cooling current of a laser was measured as function of the difference in temperature between the chip and the case. Case of the laser was held in constant temperature by mounting it to a large finned aluminium block inside a thermal cabinet. Temperature of the block was measured with a bimetal sensor and it followed the forced temperature of the cabinet regardless of cooling or warming of the laser chip. Temperature of the chip was monitored using the thermistor inside the laser package. Resistance of the thermistor in 25°C was measured and found to be the same as in the inspection report of the component (9900Ω). Resistance of a NTC thermistor can be written as

$$R = R_0 \cdot e^{B \cdot \left[\frac{1}{T} - \frac{1}{T_0}\right]}, \tag{41}$$

where R_0 is the nominal value of the resistance in room temperature, B is the thermistor constant and T_0 is the room temperature 298K.

Schematic of circuit used in temperature measurement is shown in Fig. 49. Thermistor constant was assumed to be the nominal 4000. It is specified to be between 3800 and 4100. Fig. 50 shows the dependency between voltages measured and the temperature and possible error due uncertainty in thermistor constant. Possible error is small and is not worth further treatment.

The measured differece between the chip and the case as function of the cooling current is shown in Figs. 51 and 52.

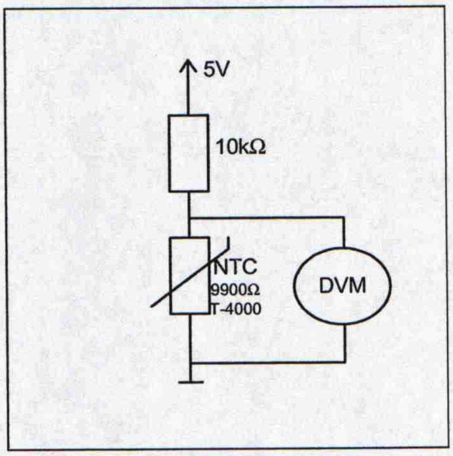


Figure 49. Circuit used in temperature measurements.

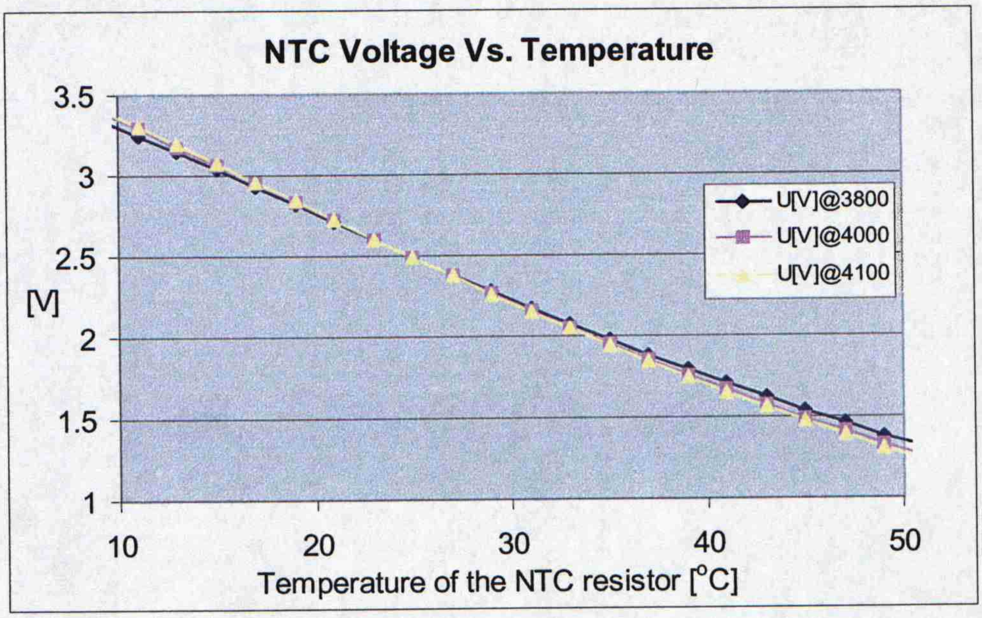


Figure 50. NTC voltages calculated with three different thermistor constants.

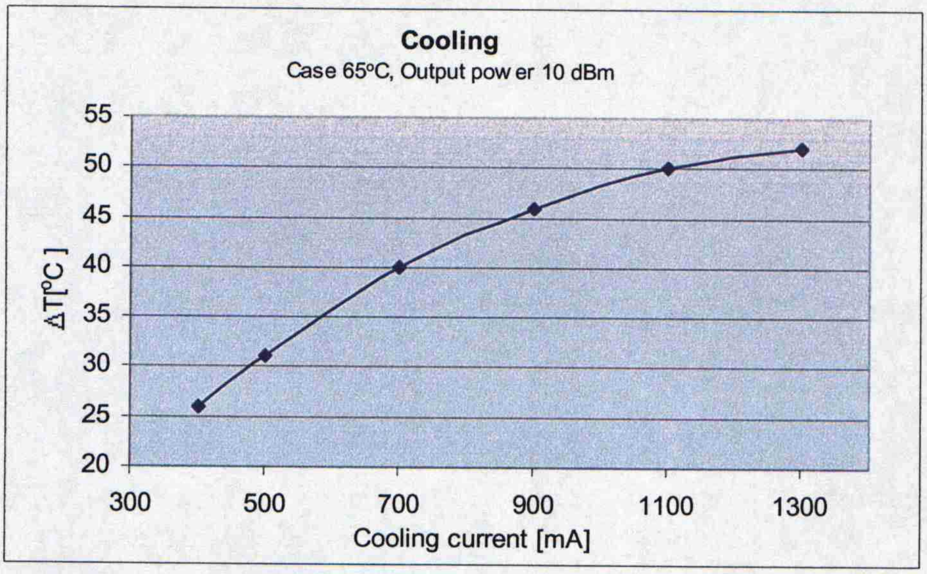


Figure 51. Measured temperature difference between the chip and the case as function of the cooling current.

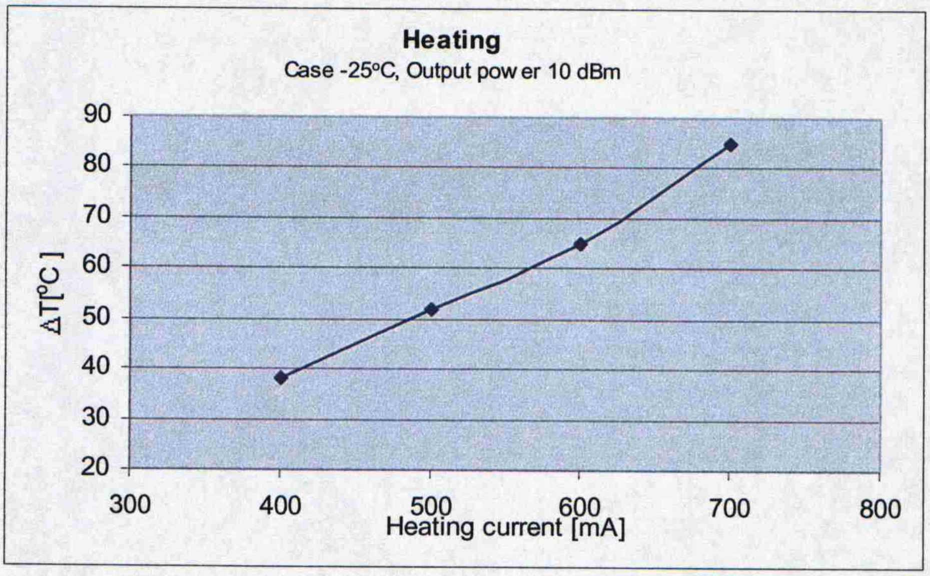


Figure 52. Measured temperature difference between the chip and the case as function of the heating current.

5.5 Wavelength stability of the laser

Environmental temperature changes might change the temperature of the laser chip itself too. This will cause uncertainty in the wavelength. The change of the output power of the chip will also have an effect in the wavelength. This is a problem since the output power needs to be adjustable and the wavelength still has to be very accurate.

Environmental temperature has very little effect on the wavelength of the laser as long as the TEC can handle the cooling or warming. At the temperature in which wavelength controlling circuitry has to change the polarity of the current a small discontinuity will exist. The amount of this effect was measured with optical spectrum analyser using the maxhold function. With a rapid change in the temperature, the wavelength drift was less than 0.15 nm as can be seen from Fig. 53.

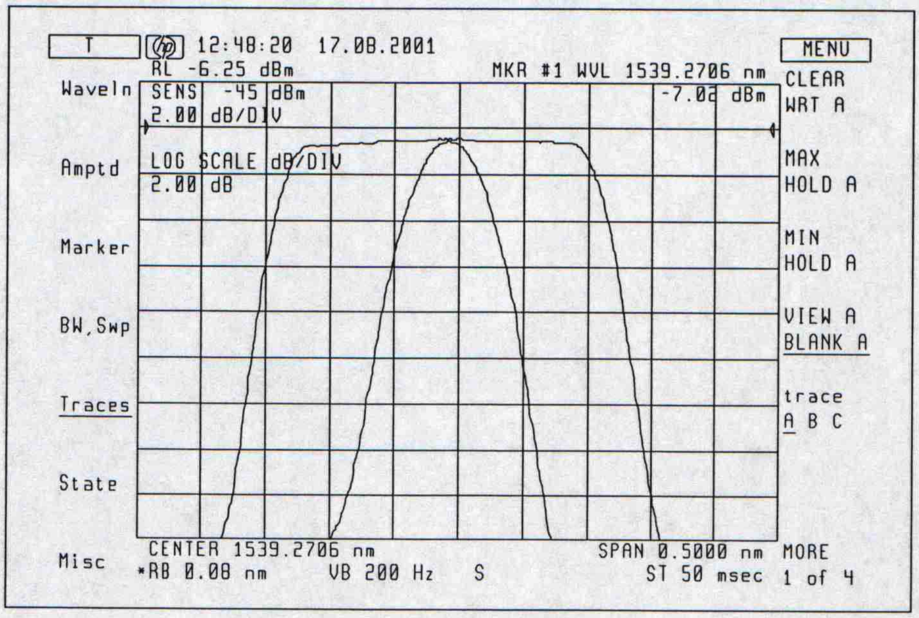


Figure 53. Measured wavelength drift due to change in the environmental temperature.

The rapid change in the temperature was simulated with a small temperature cabinet by driving the temperature several times up and down as fast as possible. This rapid change in the temperature is not likely to occur but with a thermostat driven forced ventilation the temperature stabilising circuit may be forced to change polarity quite fast.

In a more typical case when the temperature changes slowly the circuitry is able to keep the error in the wavelength smaller than 0.05 nm. All measured drifts are acceptable as the minimum channel bandwidth of the DWDM filters is specified to be greater than 0.5 nm.

At different output powers density of charged particles in the chip is different and this causes wavelength to change slightly. In other words, the change in the bias current changes the wavelength. The phenomenon is similar to the adiabatic chirp.

The change in wavelength due to change of the output power was measured using Teleste DVO771 transmitter prototype. Results of the measurements are shown in Fig. 54. The change in the wavelength was very small and could be tolerated. As the change is constant over the whole practical temperature range (15 … 35 °C) and the tuning of the wavelength is a linear function of the controlling DAC value the compensation of this effect with software is quite easy. With a larger output power a smaller DAC value needs to be used for the same wavelength. This compensation is made in the DVO771 transmitter.

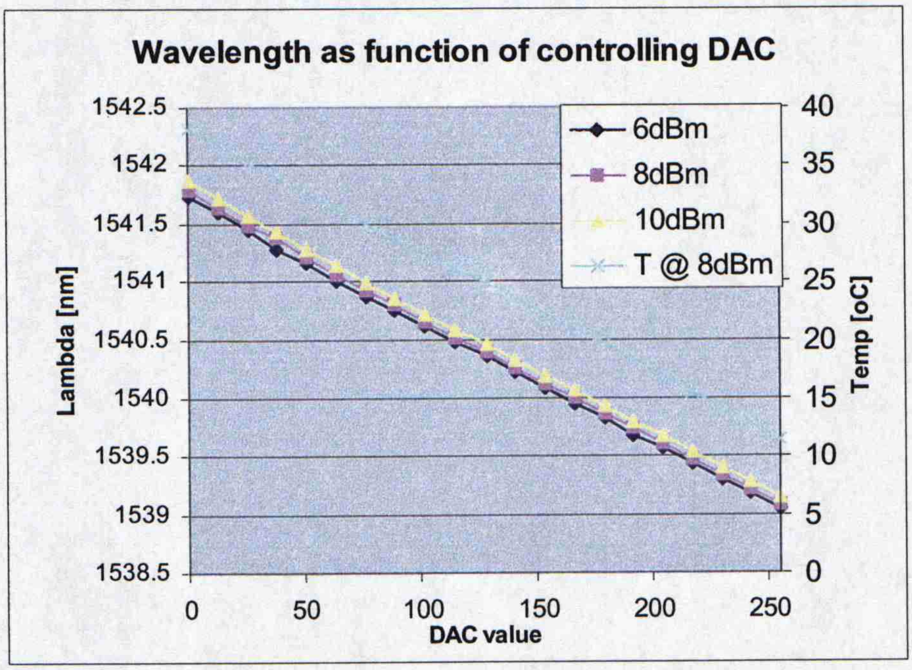


Figure 54. Laser wavelength as function of controlling DAC value for three different output powers and the chip temperature at 8 dBm output power.

5.6 Wavelength response of DWDM filters

The wavelength response of the DWDM filters was measured by connecting an 8-channel mux and a demux in series with a patch fibre. White light was inserted into mux and output of demux connected to optical spectrum analyzer (OSA) (HP 70950A). For reference, white light was inserted directly to OSA. Each channel was measured separately, so that the pass band can be seen clearly. Example of the measurement result is presented in Fig. 55.

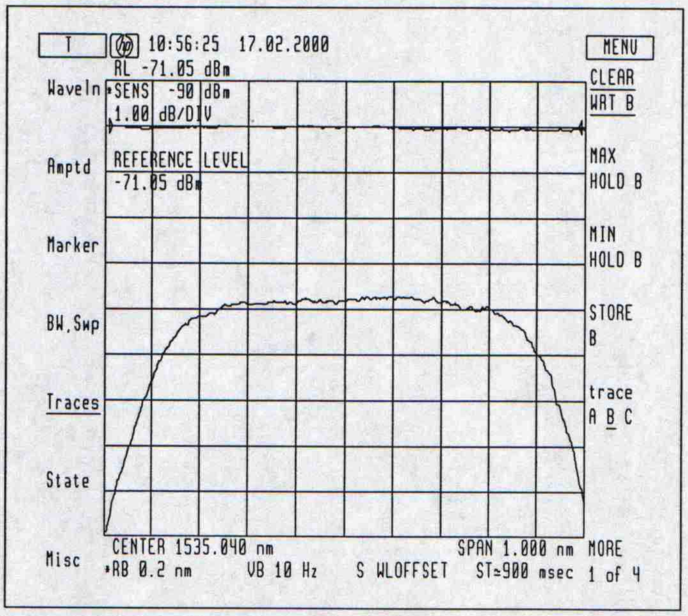


Figure 55. Attenuation of channel 53.

Table 1 shows the attenuation values for the mux and the demux measured by the manufacturer and and sum of these. Last column shows attenuation values for the pair (my measurement). All measurements seem to match, although 0.5 to 1 dB extra attenuation can be seen in values measured at Teleste. Manufacturers values might be measured without connectors. Channel 55 does not fit into this series very well, as it is the only value smaller than measured by manufacturer, but there must be some uncertainty in manufacturers measurements as well.

Table 1. Attenuation of the multiplexers.

Channel	Att. Mux	Att. DeMux	Att. Pair	Att. Meas.
45	1,08 dB	2,26 dB	3,34 dB	4,00 dB
47	1,32 dB	1,59 dB	2,91 dB	3,40 dB
49	1,31 dB	1,26 dB	2,57 dB	3,50 dB
51	1,37 dB	1,68 dB	3,05 dB	3,80 dB
53	1,93 dB	1,29 dB	3,22 dB	3,80 dB
55	2,01 dB	1,24 dB	3,25 dB	2,90 dB
57	2,15 dB	0,66 dB	2,81 dB	3,30 dB
59	2,77 dB	0,66 dB	3,43 dB	3,90 dB

Adjacent channel attenuation of the demux was measured as well. In this measurement, white light was inserted directly into the demux and measured with OSA. Result of the measurement is shown in Fig. 56. Dynamics of the measurement is limited by the sensitivity of the OSA and the small power density of the white light source. More than 30 dB of attenuation of adjacent channel light can be seen. This means better than 60 dB crosstalk attenuation, which is a very good value for the QAM signals. Dynamics of the measurement could be improved by over 20 dB by using a good quality tuneable laser source. A tuneable source was not available for the measurements.

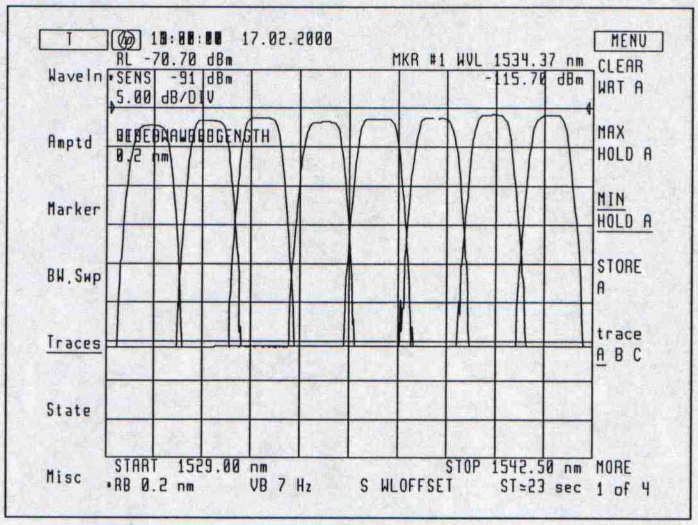


Figure 56. Adjacent channel attenuation of DeMux.

5.7 Linear crosstalk in DWDM filters

Linear crosstalk can appear in DWDM components. Optical isolation of adjacent channels in DWDM filters is typically better than 30 dB. For electrical signals this indicates crosstalk values better than 60 dB. For the case of two adjacent channels having only 30 dB optical isolation the crosstalk would increase to 57 dB. In the filters used in the measurements, the adjacent channel attenuation was always better than 30 dB and crosstalk values were always better than 60 dB. This isolation is acceptable in QAM transmission, for analogue signals typically 70 dB isolation is required and this in some cases is not achieved. However, for analogue transmission the linear crosstalk is not the most limiting factor.

5.8 Temperature stability of the DWDM mux and demux

Typically both mux and demux are located inside buildings and their environment is therefore stable. The drift of the components is also specified to be very small, less than 0,001 nm/°C (Appendix C). Even with a temperature variation of 60°C the drift would still be small and the measurement would be pointless.

5.9 Wavelength response of EDFA

Wavelength response of Teleste +13dBm EDFA (DVO713) was measured using 8 wavelengths (channels 45,47, ..., 59). The input power was varied from –8 to +10 dBm using variable optical attenuator. DVO713 EDFA keeps the output power constant by automatically changing the pump power if needed.

The output spectrum of the EDFA was adjusted to be flat (see Fig. 60), at 0 dBm input power by tuning the input powers of the individual transmitters. The input spectrum was quite flat as well as can be seen in Fig. 57. Thus the wavelength response is flat with input powers around 0dBm with the 8 chosen channels. Smaller input power gives rise to the shorter wavelengths around the ASE-peak at 1530 nm. Higher input power has the opposite effect. The change of 2 dB in input power to either direction causes just under 1 dB tilt to the response (see Figs. 59 and 61). The same trend seems to continue to +10 dBm (~4dB tilt Fig. 62) and to -8 dBm (~3dB tilt Fig. 58).

±2 dB output power tuning range of DVO771 transmitter is wide enough to compensate the EDFA gain shape in links with one EDFA. Links with two or more EDFAs can in most cases be realized as well with no problems. To achieve a good noise performance and still some gain, the input power should be around 0 … +3 dBm. In this region the gain shape is quite flat, and no problems should occur.

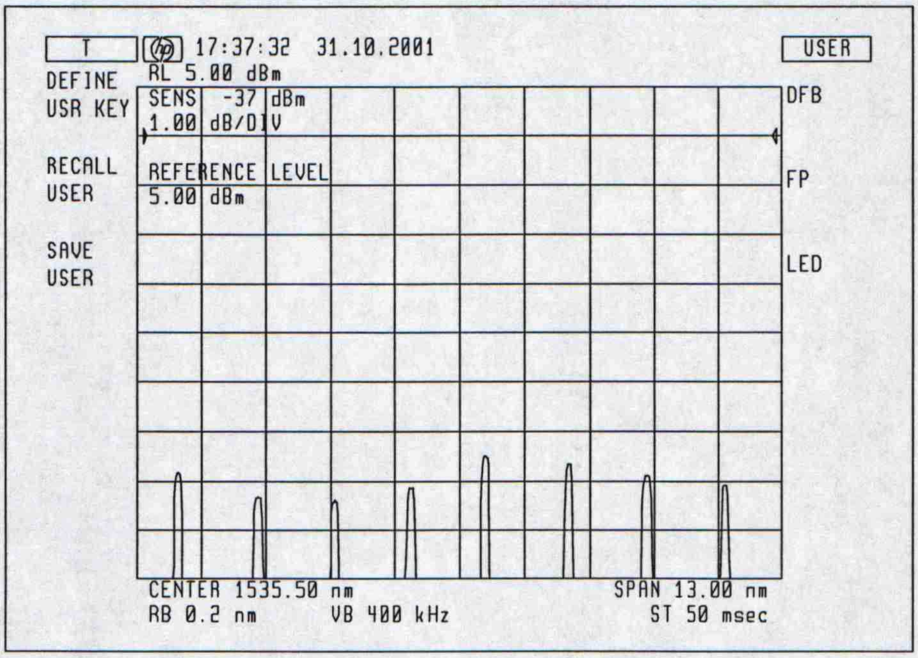


Figure 57. Input spectrum to the EDFA

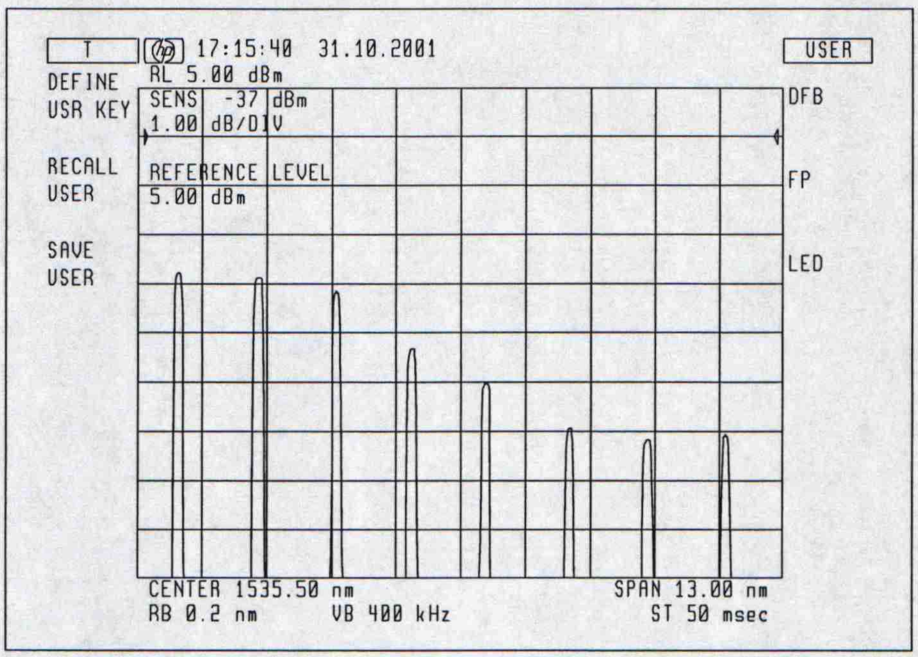


Figure 58. Input power -8dBm, -17dBm per channel.

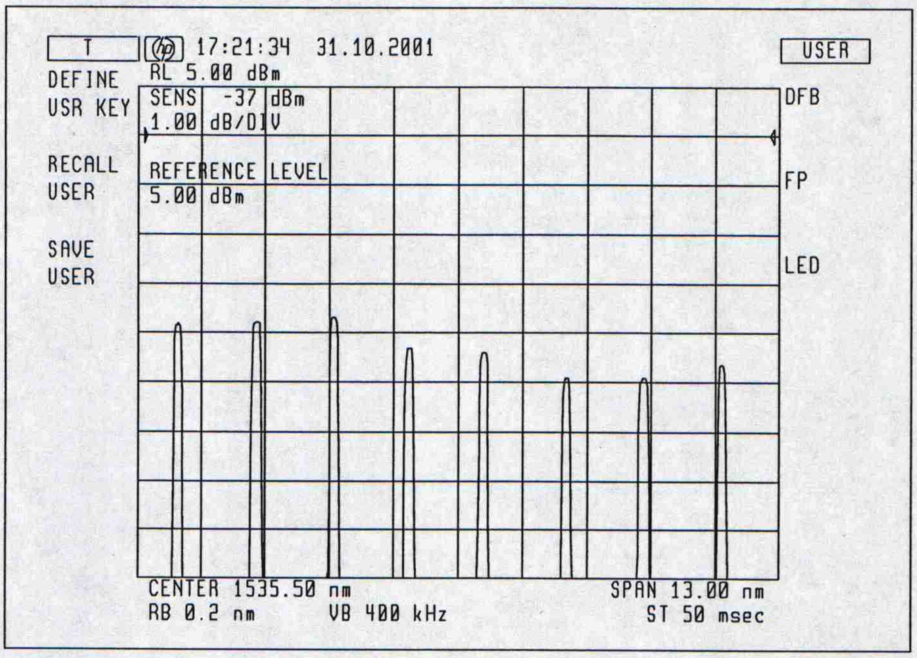


Figure 59. Input power -2dBm, -11dBm per channel.

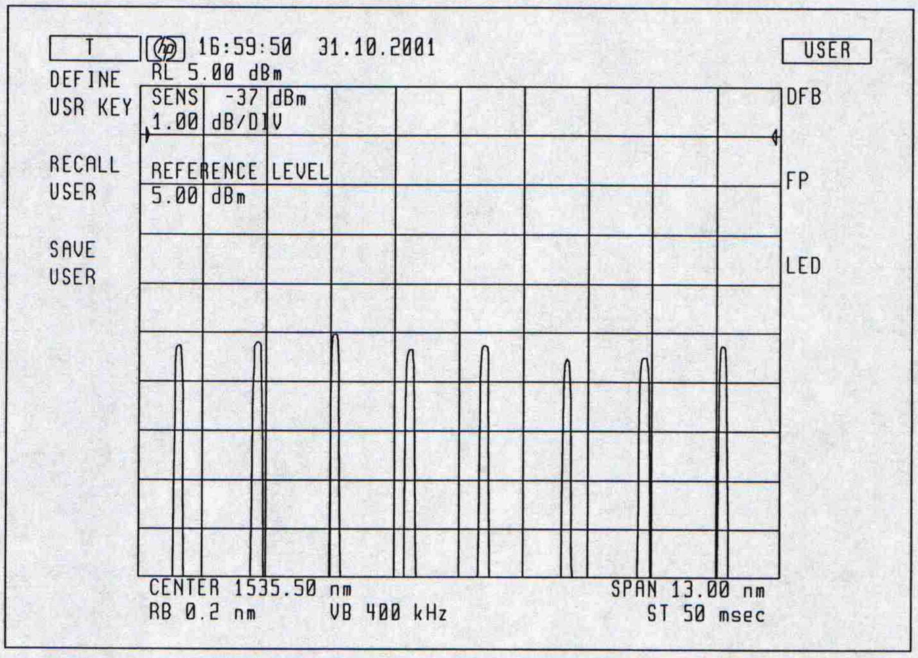


Figure 60. Input power 0dBm, -9dBm per channel.

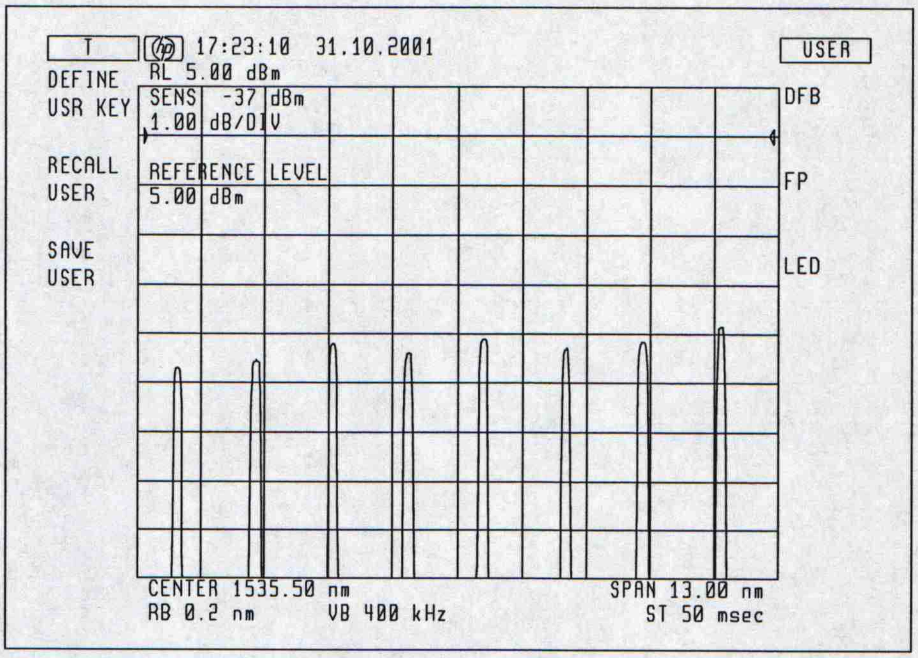


Figure 61. Input power 2dBm, -7dBm per channel.

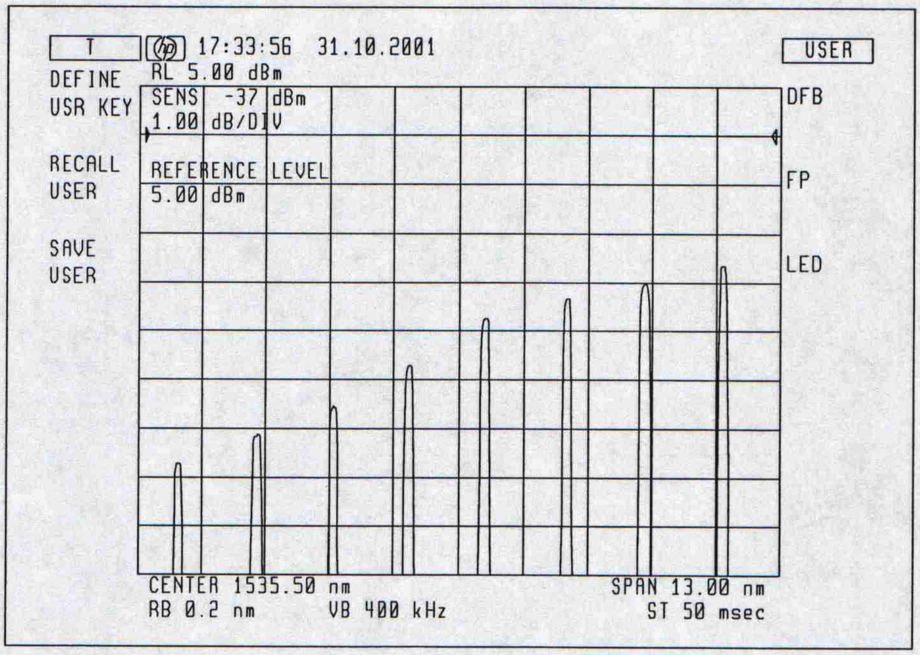


Figure 62. Input power 10dBm, 1dBm per channel.

6 MEASUREMENTS OF FIBRE INDUCED EFFECTS

Fibre causes some unwanted effects to the transmitted signals. In DWDM transmission of digital TV channels with directly modulated transmitters, the chirp and dispersion induced noise is the most important to understand. This phenomenon sets the biggest limitations to the transmission, but with proper design, it causes no problems. Chirp and dispersion induced distortion and couple of other nonlinear fibre effects are investigated in this chapter.

6.1 Chirp and dispersion induced distortion

Measurements with different fibre lengths show the effect of chirp and dispersion together. Predistorted TX was measured with no fibre, with 25 km and with 50 km. Output power of the transmitter was +8 dBm. Link attenuation was approximately 10 dB + 1 dB connector losses in each case (~ -3 dBm at RX). Different fibre loss was compensated with an attenuator in 25 km and 0 km cases.

The load consisted of 18 QAM channels, each at 9.6% OMI (total OMI 40.6%).

Centre frequencies of the QAM channels were:

338, 346, 354, 362, 370, 378,578, 586, 594, 602, 610, 622, 818, 826, 834, 842, 850 and 858 MHz.

C/N without the fibre was measured to be > 50 dB as can be seen in Fig.63. With 25 km a deterioration of ~10 dB can be seen in Fig.64. With 50 km an additional 5 dB will be lost as can be seen from Fig. 65.

Chirp and dispersion induces mostly second order distortion is which gets worse on higher frequencies. For these reasons, only the upper half of the spectrum should be used. These effects can be seen in measurements of 125 km link with 8 wavelengths. With two EDFAs in the link C/N of 43 dB could be achieved. Fig. 66 shows the measurement setup of the 125 km link. Fig. 67 shows the measured spectrum of the link. The effect of chancing one of the channels to lower frequencies can be seen in Fig. 68. Some of the channels in the Figs. 67 and 68 have lower level signals as they are QPSK cable modem channels and survive with smaller C/N. The distortion products seen in the Fig. 68 make most of the band useless.

The setup shown in Fig. 66 demonstrates that DWDM transmission of 64-QAM channels in long CATV links is possible with directly modulated transmitter. However, understanding the chirp and dispersion induced distortion and applying careful frequency planning is essential.

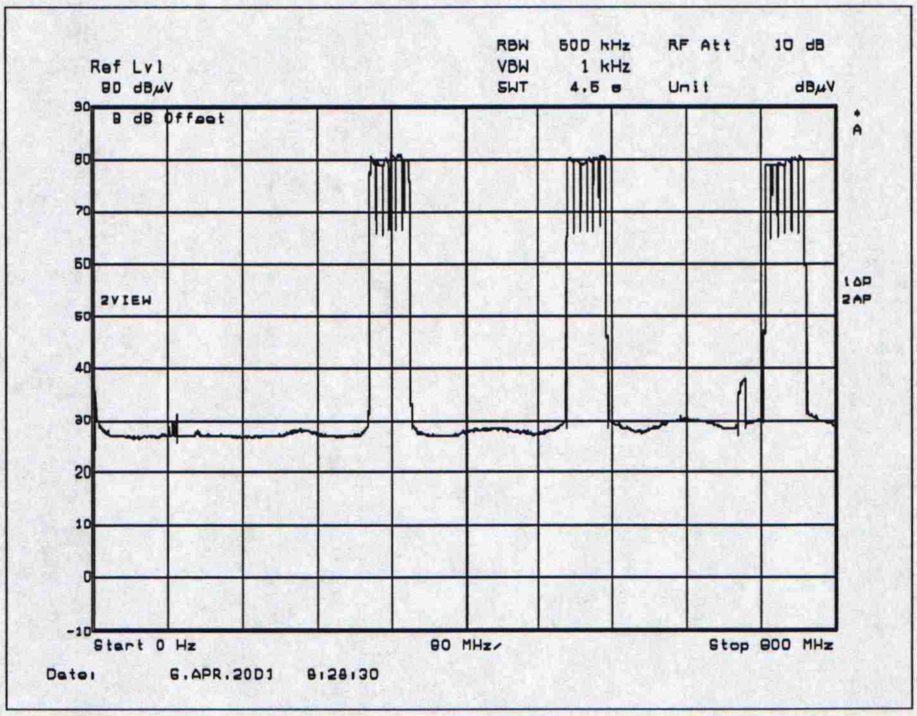


Figure 63. Measured spectrum without the fibre.

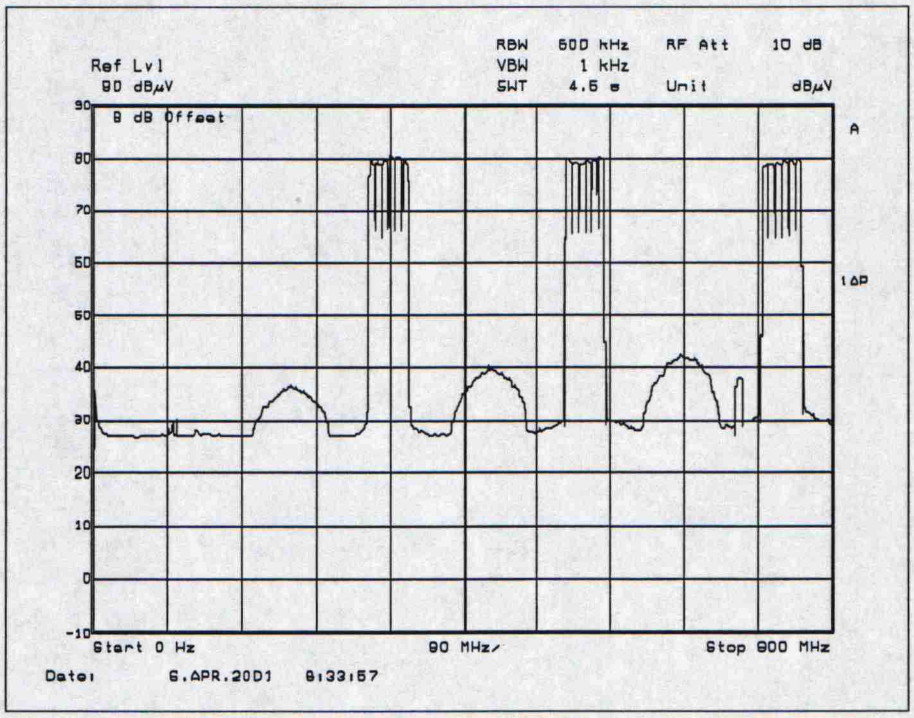


Figure 64. Measured spectrum after 25 km of fibre.

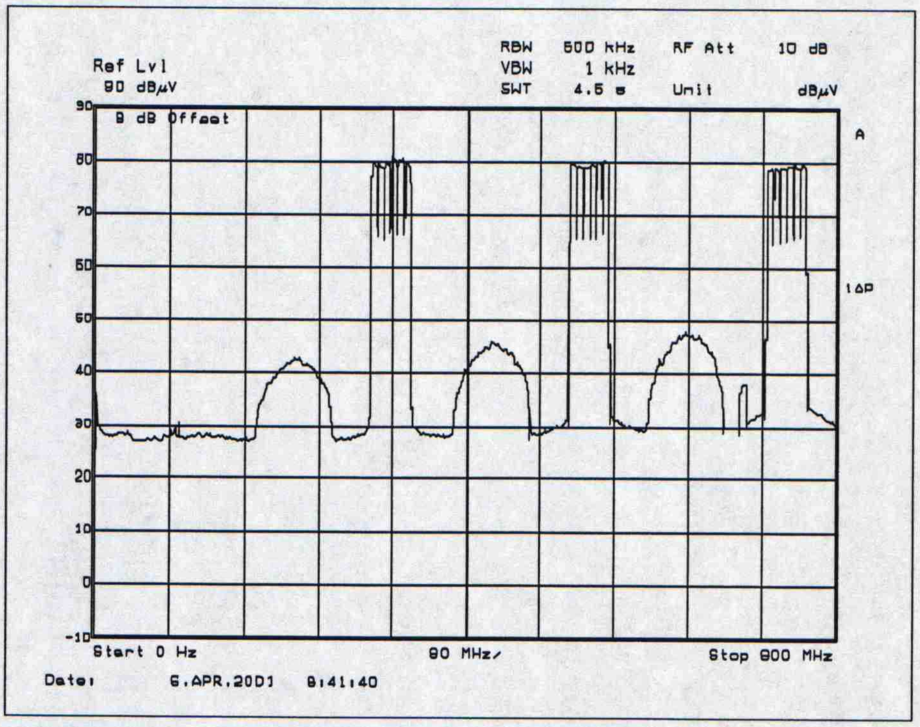


Figure 65. Measured spectrum after 50 km of fibre.

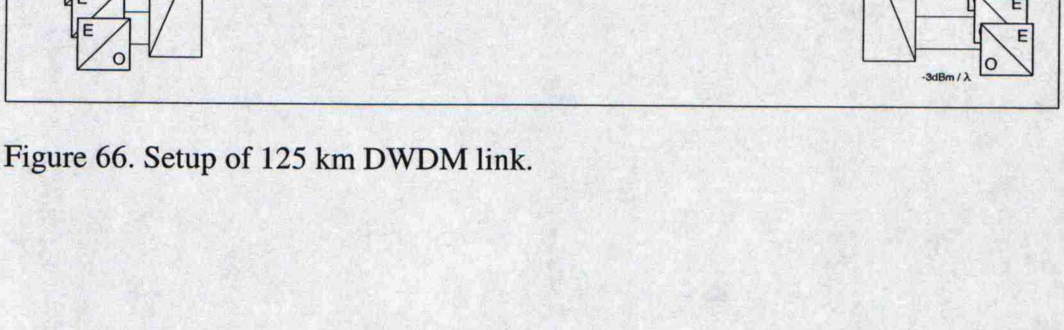


Figure 66. Setup of 125 km DWDM link.

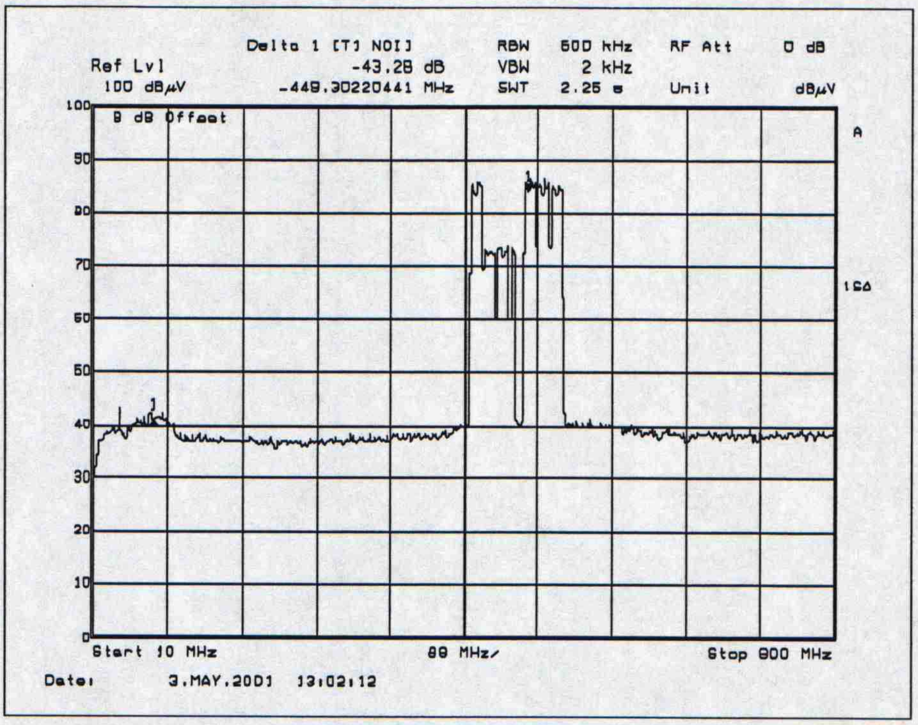


Figure 67. Measured spectrum after 125 km. OMI ~8%.

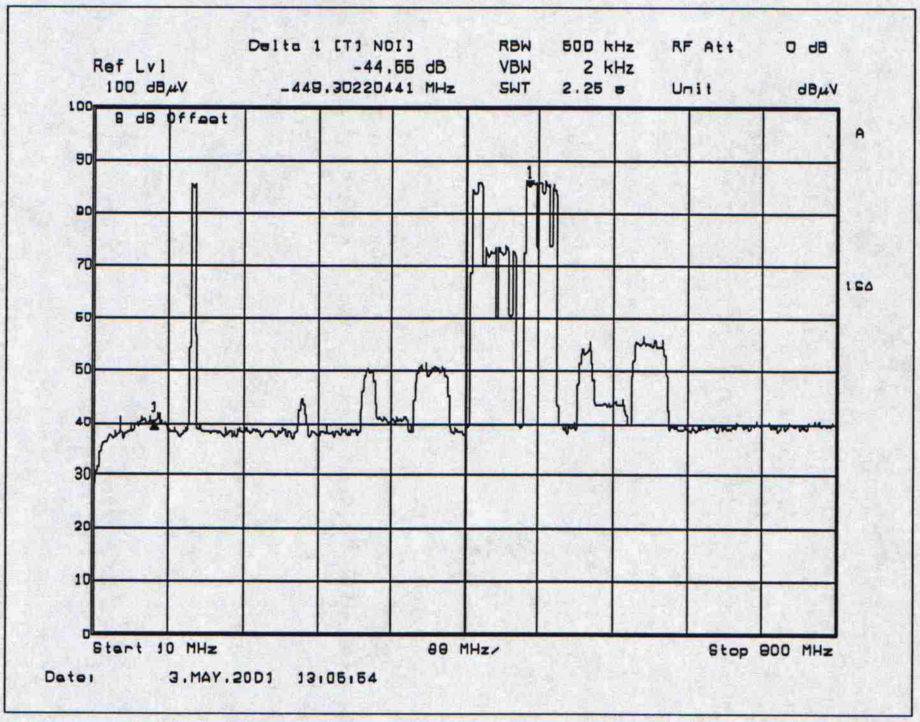


Figure 68. Measured spectrum after 125 km. OMI ~8%. Highest channel moved from 570 MHz to 130 MHz.

6.2 Stimulated Brillouin scattering

Measurements show that the SBS is not a problem in directly modulated 1550 nm links with typical launch powers and practical loads. With high powers into fibre problems may arise if only few modulating channels are used. Unmodulated linewidth of the DVO771 transmitter is much narrower than 20 MHz (SBS gain linewidth), which means that SBS threshold is low. The linewidth of the transmitter is specified to be less than 3 MHz. Chirp broadens the linewidth of the directly modulated laser and in typical cases no problems with the SBS occur.

All of the SBS measurements were made with a DVO771 TX and a 16 dBm EDFA was used to achieve the high powers needed. Link length was 25 km and two couplers in front of the fibre were used to simultaneously measure the power launched into the fibre and the power reflected from the fibre. One to ten 64-QAM signals were used as a load.

Measured power reflected from the fibre as the function of the power into the fibre with the un-modulated transmitter is shown in Fig. 69.

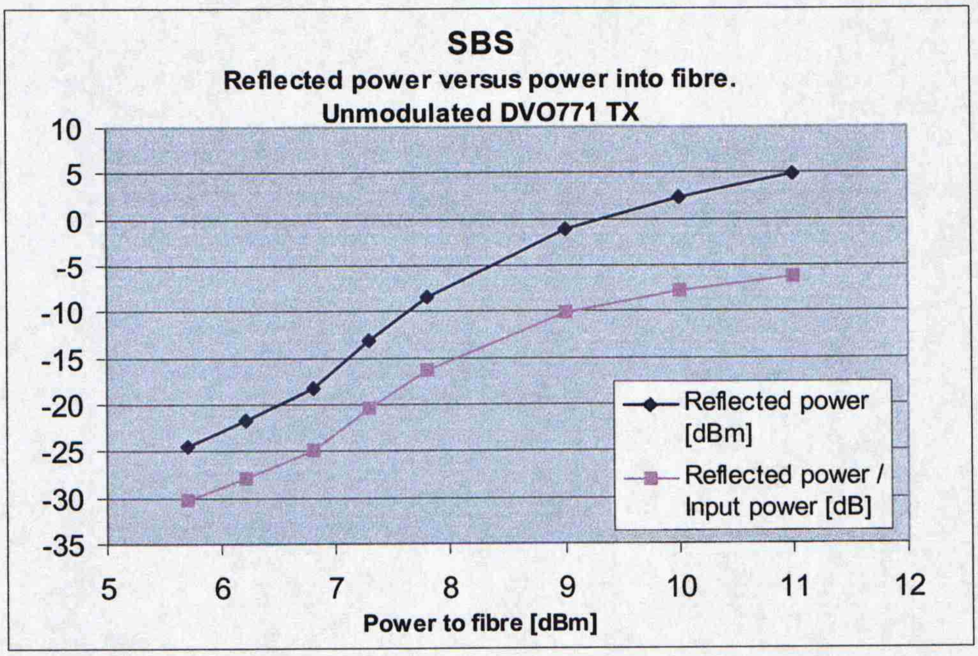


Figure 69. Measured reflected power as a function of power into the fibre.

In Fig. 70 a measurement result with 10 QAM channels is shown. Even with very low modulation index, SBS cannot be seen. Added noise in smaller frequencies is assumed to be due to the double Rayleigh scattering. The noise floor started to rise long before any increase in reflected light could be measured. Highest possible launch power into fibre in the measurement setup was 13.4 dBm. With higher OMI (6%) signal is heavily distorted due to chirp and dispersion. 10 channels at 6% OMI is a typical load but measured channel allocation is not practical.

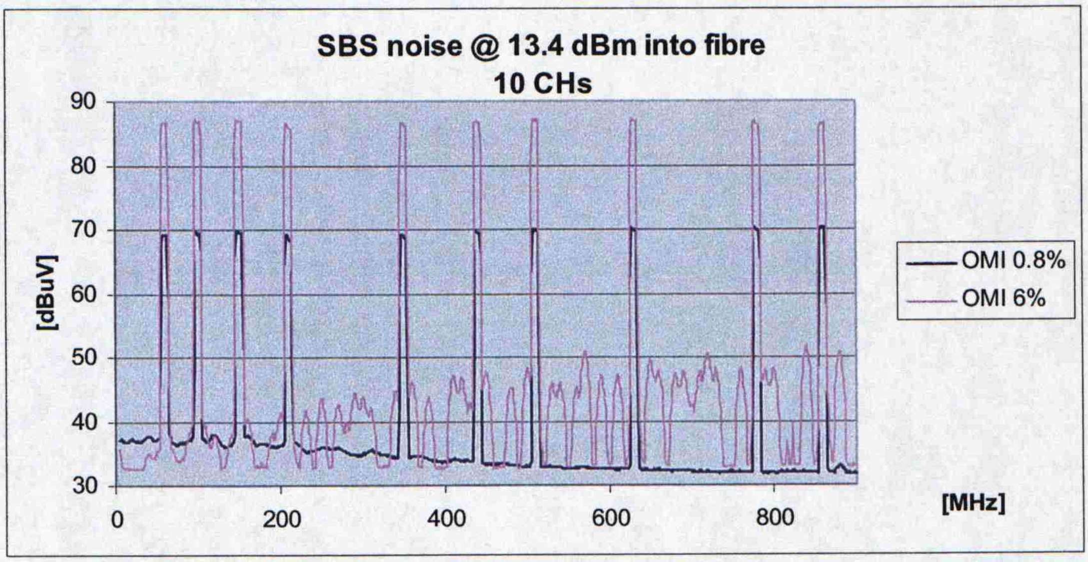


Figure 70. Measured spectrum with 10chs.

With a load of 10 channels no SBS occurred. Reducing load to five channels SBS starts to appear. In Fig. 71 dramatic change in received spectrum can be seen in transition from 4.2 to 3.8% OMI. SBS appears as 2nd and 3rd order distortion in this case. Total OMI is only about 9%, and in practical cases 13.4 dBm into the fibre should be safe. Fig. 72 shows the increase of the reflected SBS power. Amount of the reflected power is increased by 6 dB in changing OMI from 4.2% to 3.8% (Less than a dB). Heavy distortion products seen with 3.8% OMI are not due to chirp and dispersion induced distortion, because they don't increase with higher OMI.

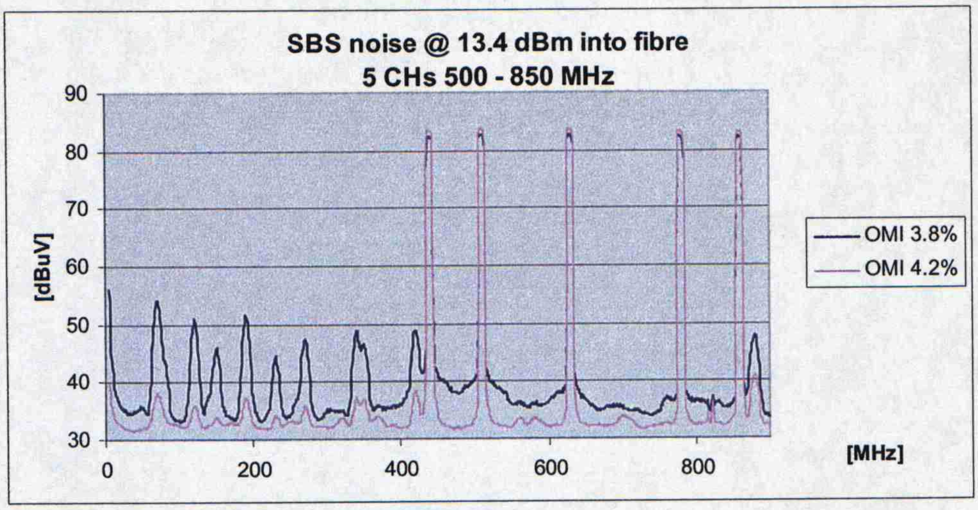


Figure 71. Measured SBS behaviour with 5 channels at 500 – 850 MHz.

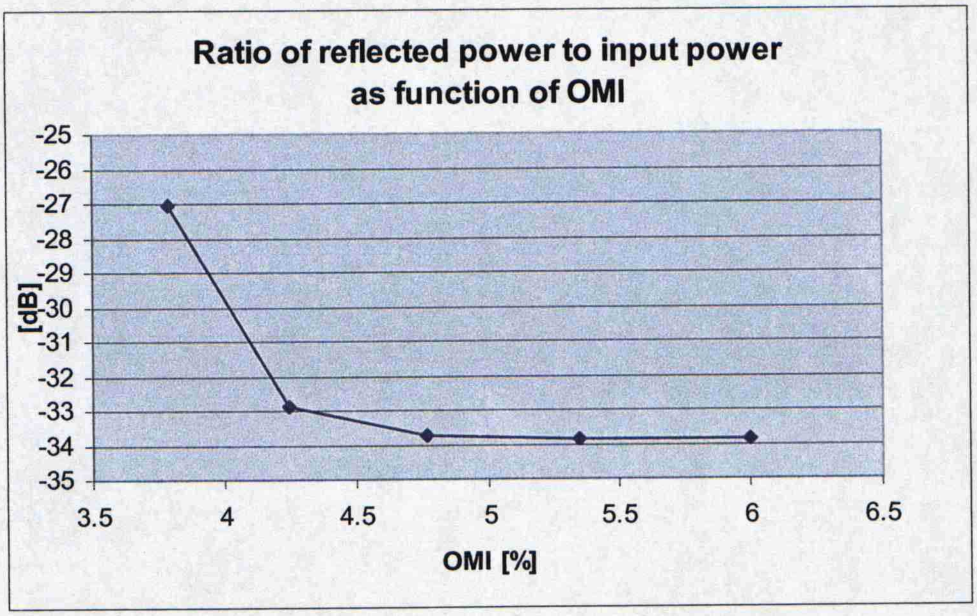


Figure 72. Ratio between reflected light and power launched into fibre. (5CHs 500-850 MHz.)

Changing the load of five QAM channels to lower frequencies helps to attenuate the SBS. The load in Fig. 73 is equal in power compared to load in Fig. 71, but frequencies are lower in this case. Again almost no SBS can be seen. Light reflected back increased only by 0.8 dB, and added noise is mainly due double Rayleigh scattering.

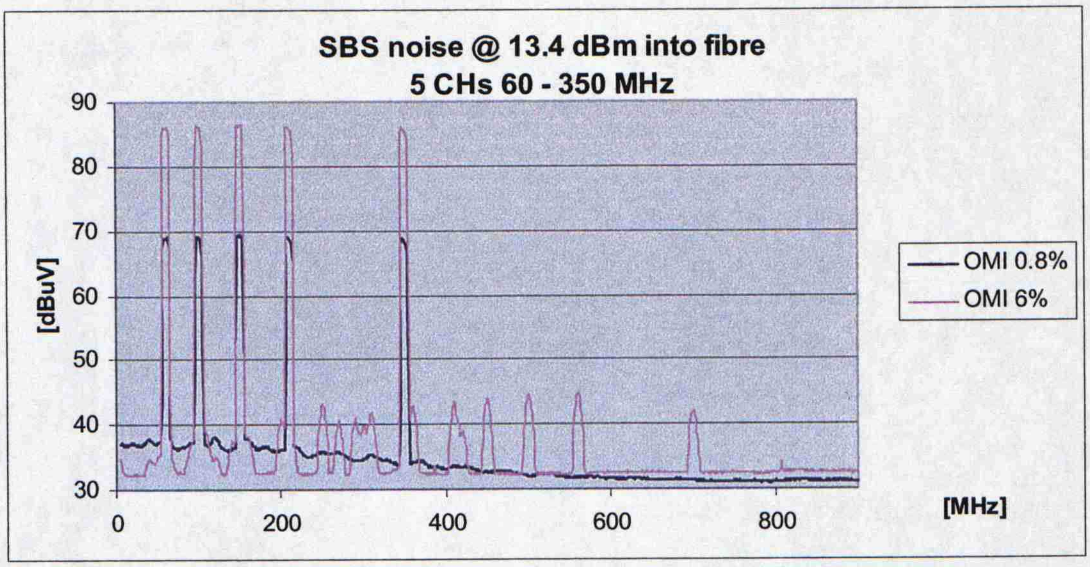


Figure 73. Measured SBS behaviour with 5 channels at 60 – 350 MHz.

SBS frequency dependency is studied by measuring with only one channel. The results can be seen in Figs 74-76. Relatively large change in OMI required to attenuate SBS can be observed in relatively small change in frequency.

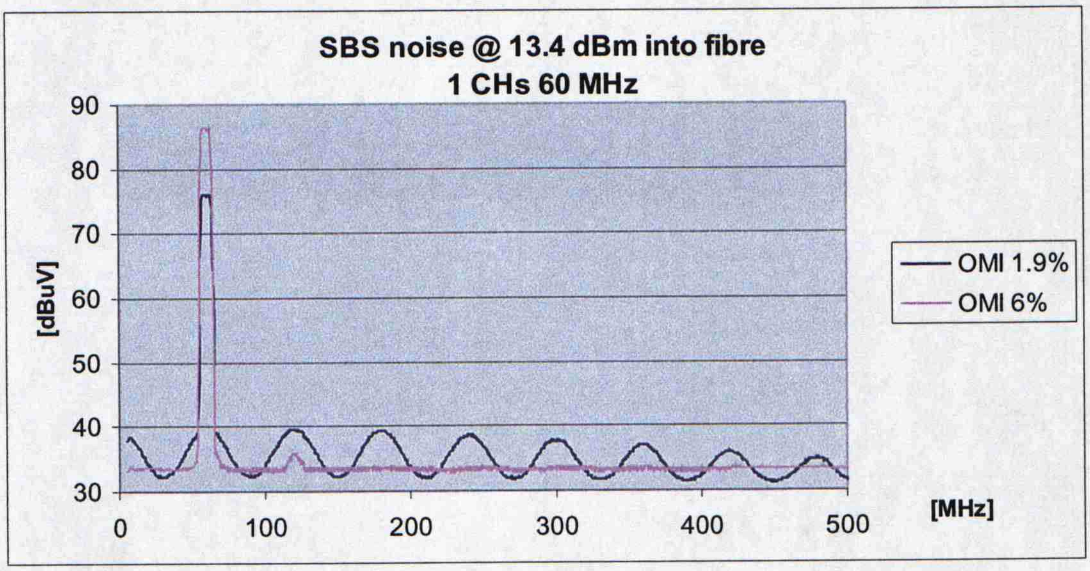


Figure 74. Measured SBS behaviour with a single channel at 60 MHz as load.

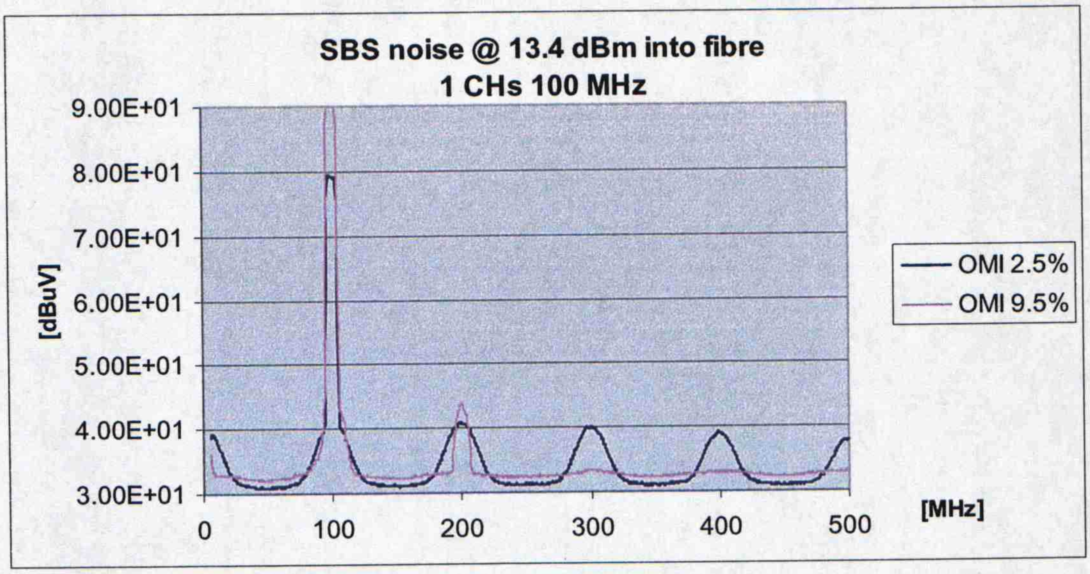


Figure 75. Measured SBS behaviour with a single channel at 100 MHz as load.

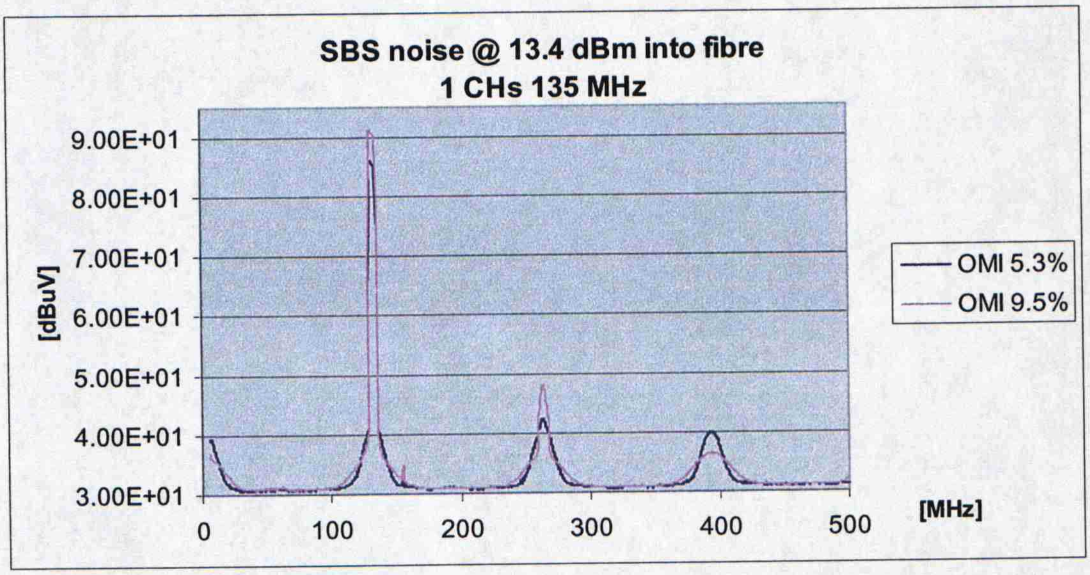


Figure 76. Measured SBS behaviour with a single channel at 135 MHz as load.

One channel with OMI of 6 % is enough to attenuate SBS at 60 MHz. At 100MHz 9.5 % OMI is needed. With modulation frequency increased to 135 MHz, the same 9.5 % OMI failed to attenuate the SBS.

6.3 XPM and SRS

Crosstalk over a 25 km link was measured with 8 wavelengths (8 DVO771 transmitters). Setup of the measurement is presented in Figure 77. All channels but the one being measured were modulated with six 64-QAM channels (60, 100, 210, 440, 630 and 780 MHz) each at 6% OMI. Frequencies were chosen so that no second order distortions fall on these frequencies. This allows BER measurements to be performed. Or in the other words, to be able to distinguish the MER limited by XPM or SRS from distortion limited MER. As crosstalk values measured were all well below -35 dBc, no BER measurements were performed. Higher OMI would have helped to achieve good measurement dynamics, but this is a typical value used in real cases having more channels.

Both XPM and SRS are sensitive to polarisation state. No polarisation controllers were available but patch fibres were twisted and taped to lab table in order to find the worst case result. Maxhold function in the spectrum analyser was used to catch the worst crosstalk values as polarisation states in the setup were fluctuating.

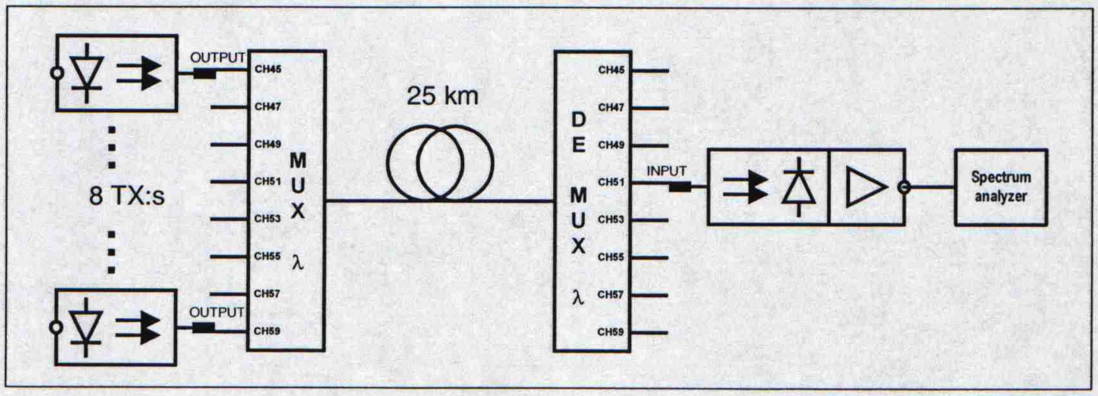


Figure 77. Measurement setup for the nonlinear crosstalk.

Output power of the transmitter was varied between 8 and 10 dBm. As power to fibre was still quite low with 10 dBm one extra patch fibre was removed between the Mux and the fibre. This patch fibre was bad, and had an attenuation over 2 dB. Figures showing measured crosstalk have been marked with TX output power. The "*" in 10 dBm* stands for reduced attenuation between the Mux and the fibre.

Actual measured power to fibre for each wavelength is shown in Table 2. Accuracy of measurement is around ±1 dB, biggest uncertainty is caused by optical connectors.

Table 2. Power into fibre for each channel.

Power to fibre for each channel [dBm]			
CH # \ TX Power	8dBm	10dBm	10dBm*
45	5.3	6.9	9.6
47	4.6	6	8.5
49	4.8	5.7	8.3
51	4.5	5.7	8.1
53	4.9	6.3	8.7
55	3.3	4.7	7.8
57	4.2	5.9	8.3
59	3	5	7.2
AVG	4.3	5.8	8.3
TOTAL	13.4	14.9	17.4

First measurements were done with 8 dBm TX power. Worst result was measured on channel 59 at 100 MHz. Crosstalk value of 45 dB is still very good. Crosstalk is smaller at 60 MHz than at 100 MHz, which is not in line with the theory. Uncertainty in polarisation states may explain this disagreement.

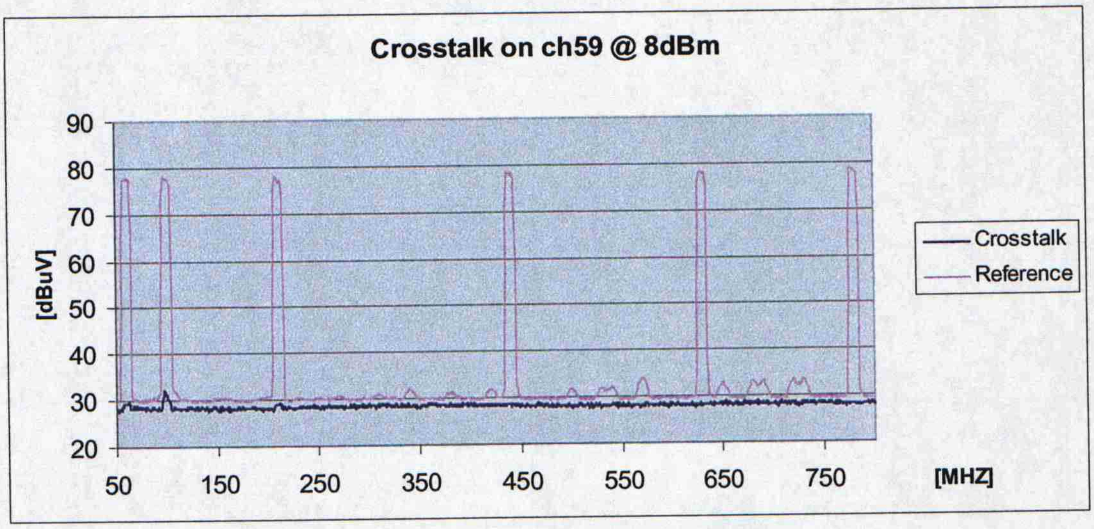


Figure 78. Modulated reference and crosstalk over the whole band.

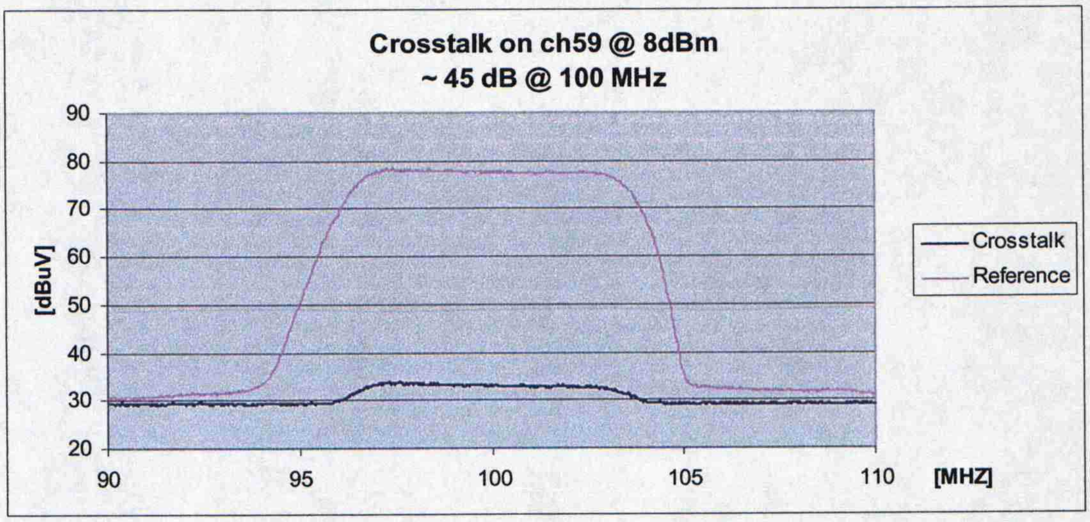


Figure 79. Modulated reference and crosstalk at 100 MHz, the worst case at ch 59.

On the channel 45 the crosstalk was slightly smaller than at channel 59. Here crosstalk is present also at 60 MHz. Figure 80 shows the biggest crosstalk at 60 MHz but the closer look shows that amount of crosstalk is equal at 60 and at 100 MHz.

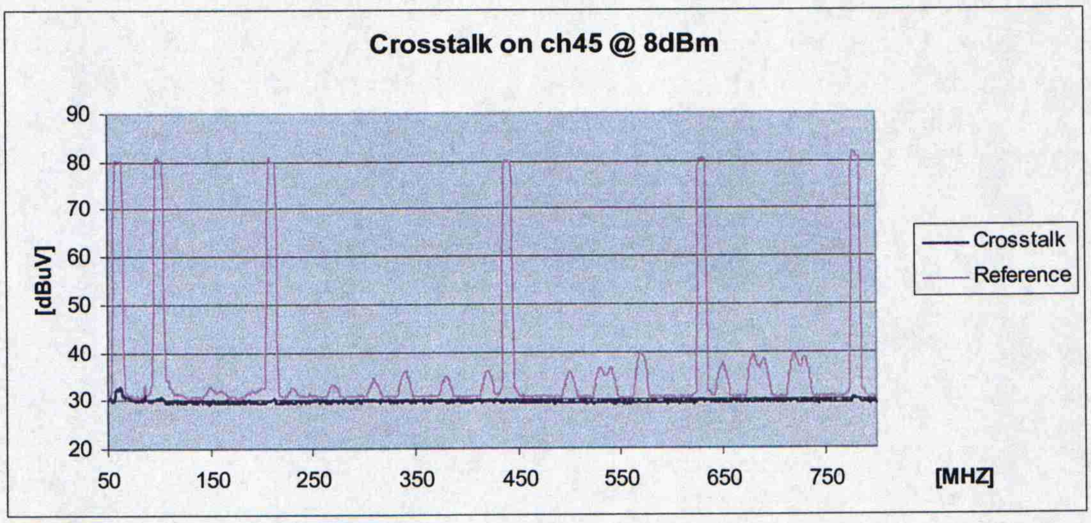


Figure 80. Modulated reference and crosstalk over the whole band.

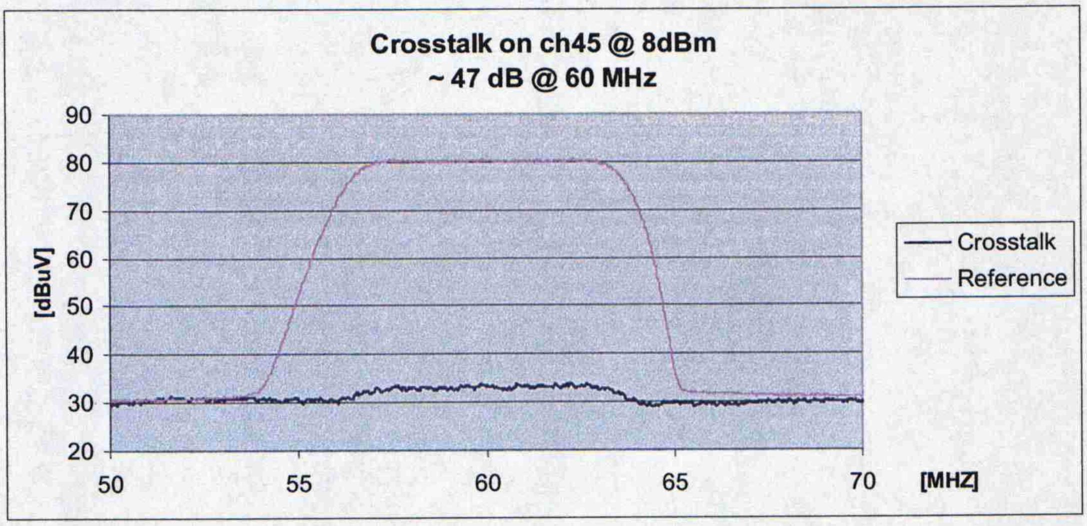


Figure 81. Modulated reference and crosstalk at 60 MHz.

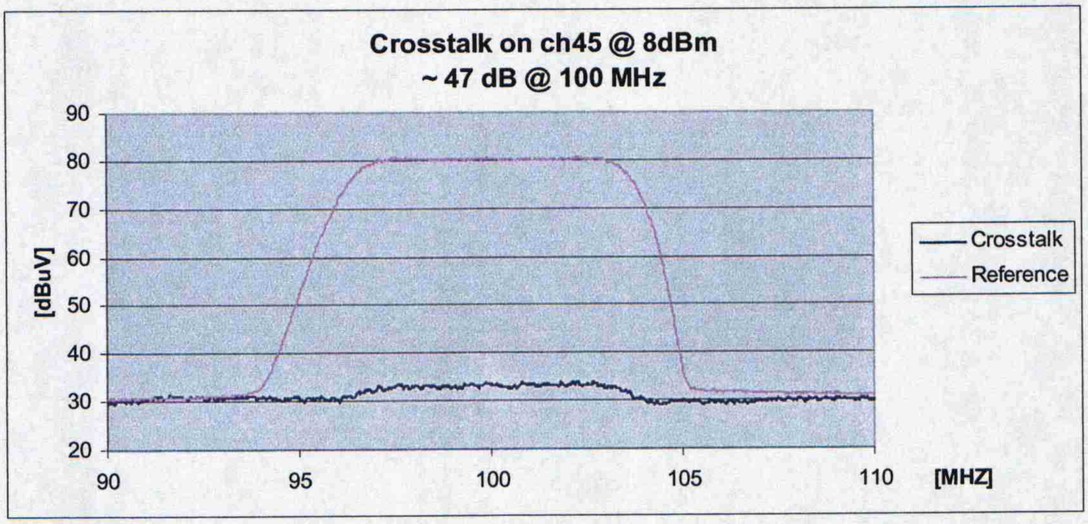


Figure 82. Modulated reference and crosstalk at 100 MHz.

On channel 49 like on the other middle channels the crosstalk was small. Other channels were measured as well but they are not documented. The results were in line with assumptions. Worst crosstalk values were found on channels 45 and 59.

TX output power increased to 10 dBm. Power to fibre was increased by 2 dB. Theory suggests 2 dB increase in crostalk per 1 dB power to fibre increased. Crosstalk ratio was 4 dB smaller than at 8 dBm TX output power which matches with the theory. With the higher power to fibre a pilot tone at 80MHz and a QAM channel at 150MHz had to be used to modulate the crosstalk channel in order to avoid SBS problems (Fig. 83).

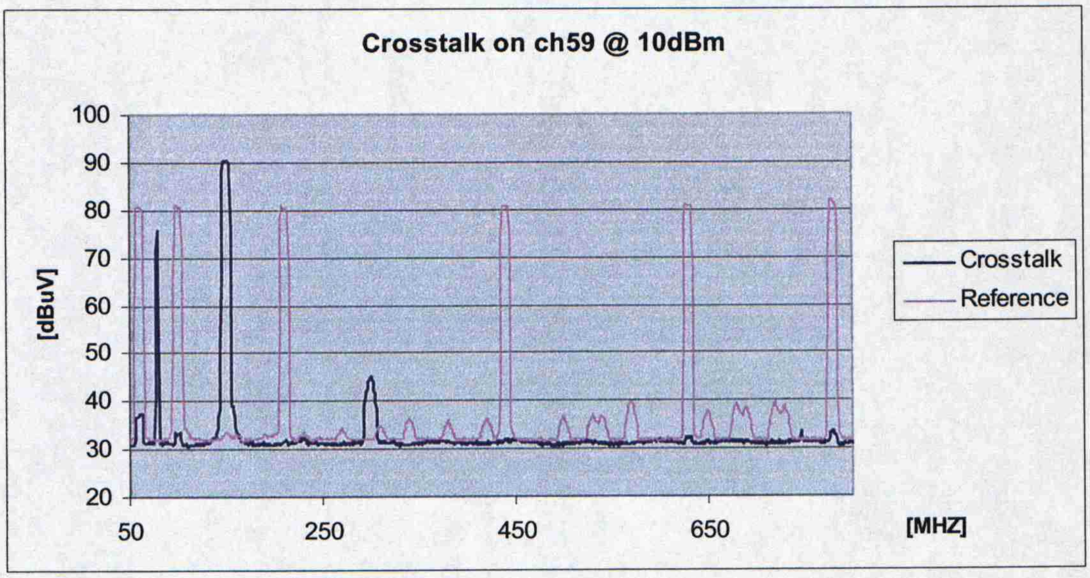


Figure 83. Modulated reference and crosstalk over the whole band.

Further increase in power results in more crosstalk.

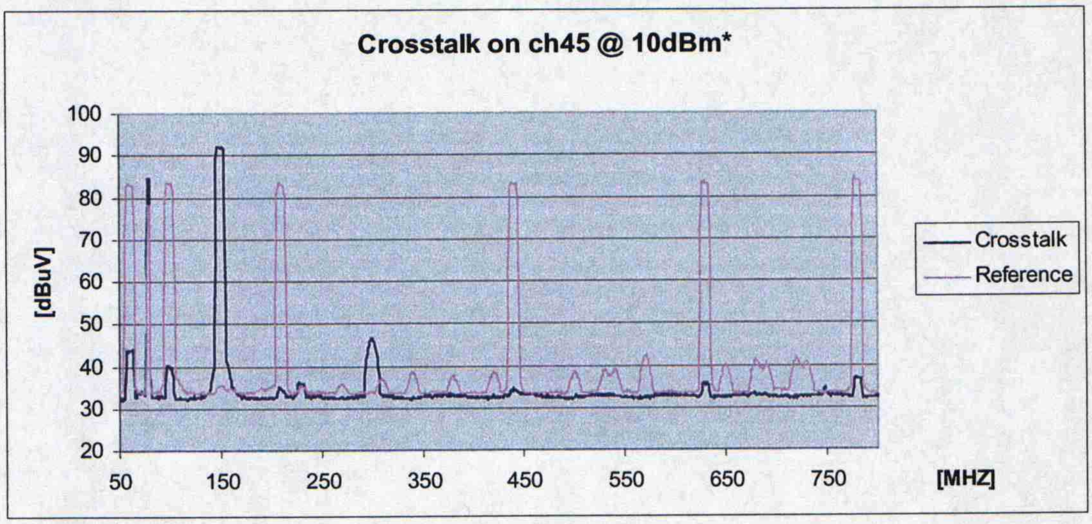


Figure 84, Modulated reference and crosstalk over the whole band.

Table 3 shows some measured crosstalk ratios for different powers into fibre. There is some disagreement between the theory and the results, but powers to fibre were quite low and test arrangement not optimal as fibres had to be disconnected to measure powers. Method for controlling polarisation was also poor.

Table 3. Measured crosstalk values

Crosstalk [dBc]	CH45 60MHz	CH45 100MHz	CH45 800MHz	CH59 100MHz
8dBm	47	47	55	45
10dBm	42	46	48	41
10dBm*	38	41	46	

Power measurement uncertainty could be improwed with a variable output EDFA in front of the fibre. This would enable power variation without touching the connectors. In this measurement, power adjustment was meant to be done by chancing the TX power but power into fibre was not as high as expected and some connector exercise had to be done. With a variable output power EDFA the total output power would be known quite accurately and relative powers could be measured at receiving end. Measurement setup would be more stable and powers known with better accuracy. Also higher power to the fibre could be achieved with at least 18 dBm EDFA. OMI doesn't affect the measured crosstalk values as long as it is the same for all channels, therefore since only 6 channels were used a higher OMI could be applied to increase measurement dynamics.

7 NETWORK PERFORMANCE

Three different measurement setups were measured; these setups are shown in Figs. 88, 93 and 99. In the first setup, both the broadcast and the narrowcast links have separate fibres to hub. Optical summing of the BC and NC signals is done in the hub. In the second setup, the possibility of using two receivers is investigated. Finally, in the third setup BC and NC is inserted into the same fibre right in the headend. In the last setup, the effect of a coaxial amplifier cascade is measured as well.

Signal quality in all examples is considered good. Loading is always chosen to be the optimum. In all cases, channels can be left out, and analogue carriers can be replaced with digital ones. Analogue carriers in the tests are un-modulated, and actual distortion performance with real signals would be better.

7.1 Target specification

The quality of an analogue TV picture is a subjective matter. However, standards exist, and in Europe minimum C/N at wall outlet is specified to be 44 dB. Any unwanted disturbances in the channel frequency shall be more than 57 dB below the carrier /23/.

Typically an analogue optical link has C/N over 50 dB and the composite distortion products are better than 60 dBc for CSO and over 63 dBc for CTB. These distortion products are measured with 42 un-modulated TV carriers. Amplitude modulation decreases the average power of the modulated signals approximately by 3 dB depending of the picture content. Therefore, distortion ratios with modulated signals are better CSO typically by 6 dB and CTB typically by 8dB, and no distortion can be seen in the picture.

C/N of a 64-QAM at the customer wall outlet should be at least 31 dB /24/. In theory 25 dB is enough for quasi error free reception. Margin is quite big, but as the noise figure of the set top box may be quite bad (~15 dB) and the minimum signal level at the customer outlet is 47 dBμV, this might be needed. No C/N specification for the 256-QAM exists yet, but a sophisticated guess is that it will be 6dB higher (37 dB).

In the example setups the C/N target for analogue channels has been ~50 dB after the link and over 48 dB after the coaxial amplifier cascade. The target distortion level was better than 63 dB after the link and better than 60 dB after the coaxial amplifier cascade for both CSO and CTB.

7.2 Effect of the future channel plans on the network

Typical load on the CATV network today is ~50 analogue TV channels and a few QAM channels. Common assumption is that the number of analogue channels will be smaller in the future. The analogue channels will be replaced by the QAM channels. This will most likely be done gradually, a few channels at the time. Some of the customers will not replace their receivers for a long time, and some services will have to be maintained for this group as well.

The effect of the different loading on the network performance was tested with a few different loads. The loads were:

1) 50 analogue channels and 30 QAM channels

2) 10 analogue channels and 80 QAM channels

3) 93 QAM channels.

Examples of the channel plans used in the tests are presented in Figs. 85 - 87.

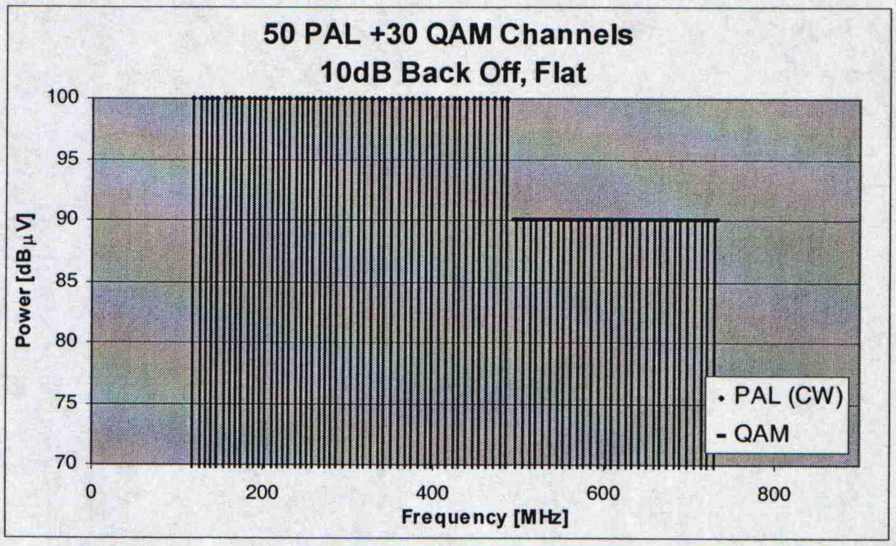


Figure 85. Example load 1. 50 analogue and 30 64-QAM channels. Flat frequency response.

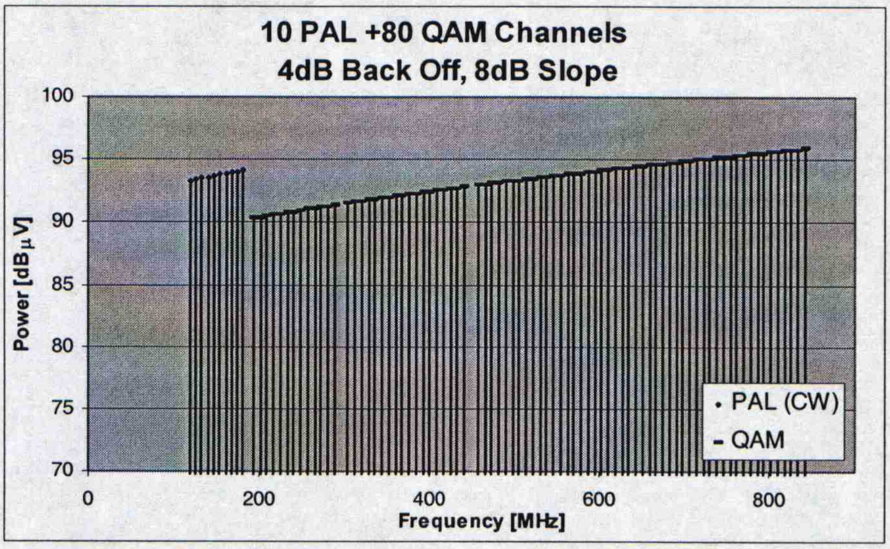


Figure 86. Example load 2. 10 analogue and 80 256-QAM channels. 8 dB sloped frequency response.

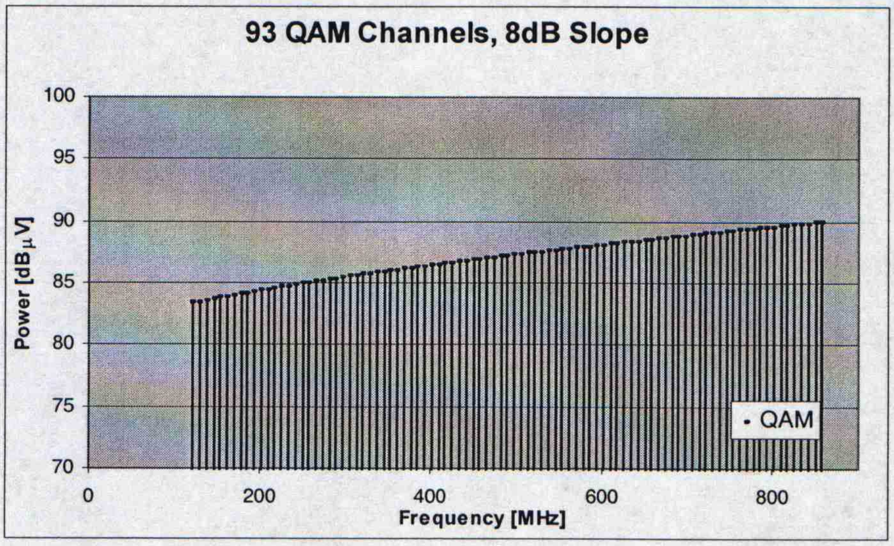


Figure 87. Example load 3. 93 64-QAM channels. 8 dB sloped frequency response.

The mixed analogue / digital loads were analysed with both 4 dB and 10 dB back-offs. In the tests, a 1550 nm externally modulated transmitter were used as broadcast transmitters. Coaxial cable amplifier cascades with push-pull, power doubler and GaAs technologies were tested. Replacing analogue channels with digital ones makes the loading easier. Distortion products of the digital channels may behave differently and they appear as noise. For this reason the tests were performed. No big surprises were found. All the tests with DWDM narrowcast were performed with 50 analogue channels, as this is the most demanding application. All the tests performed confirm the assumption that replacing analogue channels with QAM channels has either positive or no effect to performance.

Table 4 shows the total power of different loads used in the tests. Flat frequency response is used with the optical transmitters and the sloped with the coaxial cable amplifiers. The total powers are calculated with the analogue reference channel at 862MHz at 100dBμV. The channel plans don't have an analogue channel at 862MHz, but if there would be a channel, it would be at 100dBμV. Total power of commonly used test raster 42 channel CENELEC is also shown in the table.

Table 4. Total power of the loads used in the tests.

Total Power [dBμV]		
Load	Flat	8 dB Sloped
50/30 4dB Back Off	117.90	114.10
50/30 10dB Back Off	117.20	112.90
10/80 4dB Back Off	115.80	112.40
10/80 10dB Back Off	112.30	107.40
93 QAM "4dB Back Off"	115.00	112.00
93 QAM "10dB Back Off"	109.00	106.00
Cenelec 42	116.20	113.60

For the full digital loads (93 QAM channels), back off from the analogue channels cannot be measured, as there is not one left. It is however easier to compare the loads if a virtual analogue channel is assumed to exist. The total power of 93 QAM channels is calculated for the case where all the channels are 4 or 10 dB below the analogue channel that would exist at the same frequency. Replacing all analogue channels with 64-QAM channels leads to 7-9 dB lower total power than the typical loads used today. C/N requirements for the digital channels are lower and it is fare to assume that with no analogue channels left the level of all channels can be raised by 6 dB and 256-QAM used.

7.3 Test setup 1

First narrowcast setup is formed up of two separate links from the headend to the hub. In the hub, the broadcast light is divided to many nodes with an optical splitter. Narrowcast light is divided with a DWDM demux, thus there is a wavelength for each node. In the hub the divided BC signal and the de-multiplexed NC signal are added into a single fibre for each node with a red-blue filter.

The Optical back off in the node is chosen so that not much harm is done to C/N of the BC signal because of the RIN and intensity noise of the NC light. Higher optical back off means better BC noise performance, but less NC channels.

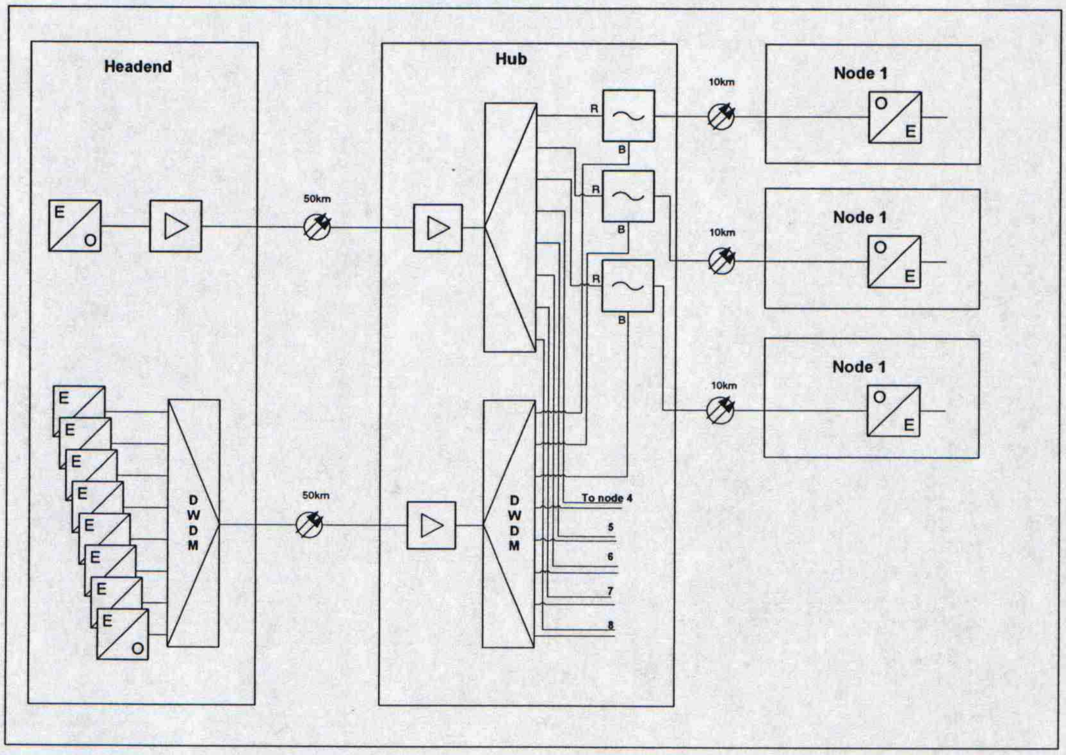


Figure 88.Outline of the test setup 1 for narrowcast.

Fig. 89 shows the measured RF spectrum of the broadcast signals only and Fig. 90 shows the spectrum of both the broadcast and the narrowcast signals.

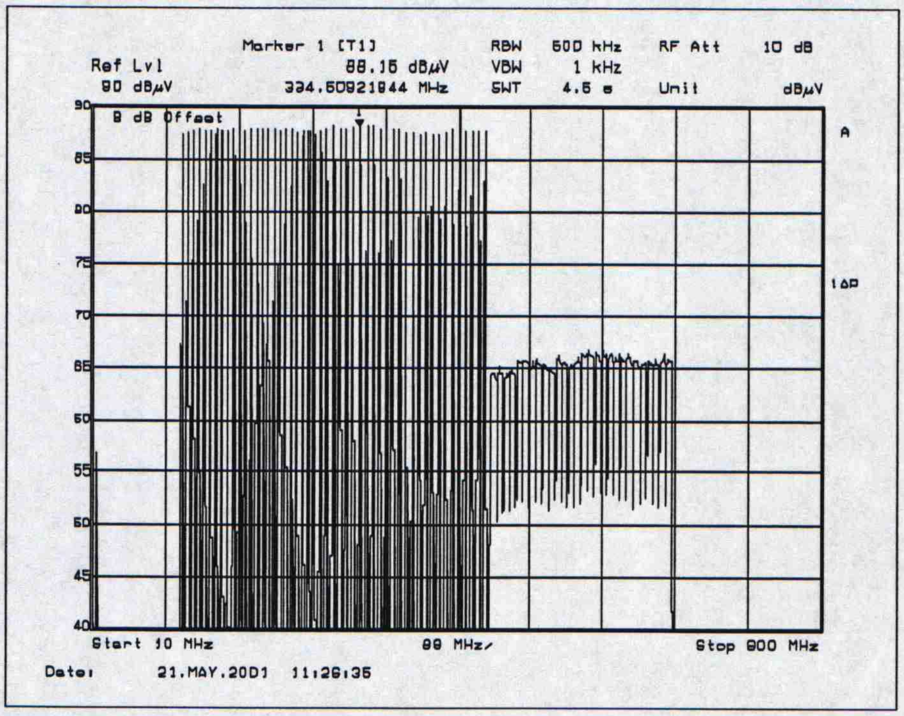


Figure 89. Frequency spectrum of BC signals only.

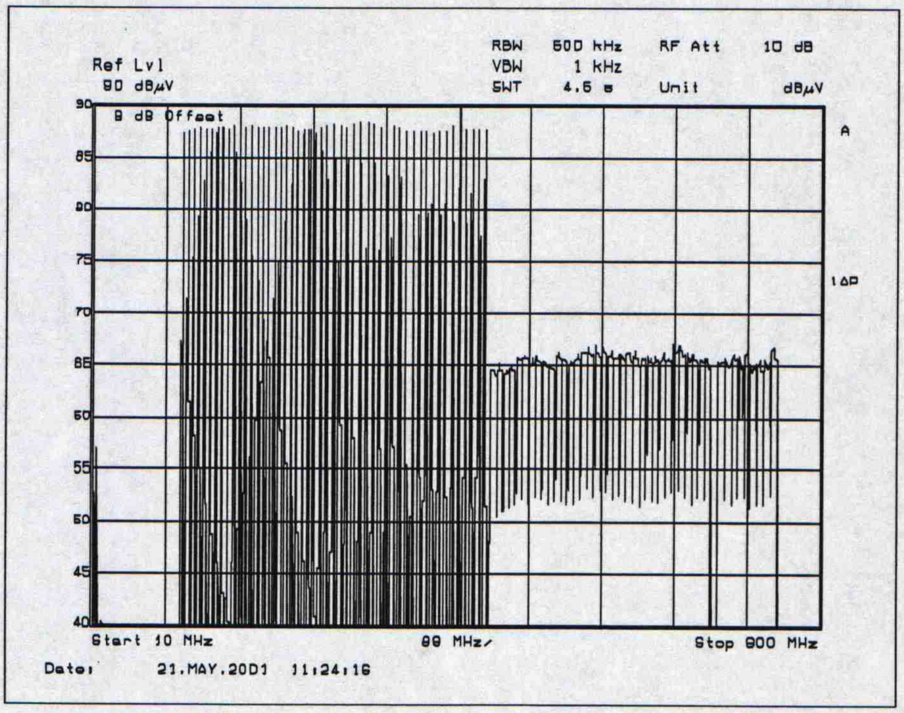


Figure 90. Frequency spectrum of BC and NC signals.

Figs. 91 and 92 show the measured noise and distortion performance for the setup 1 with and without narrowcast light. C/N decreases by 1 dB due to the added shot noise in the receiver and a bit more in the low frequencies due to the noise like CSO of the narrowcast transmitter. Distortion performance is almost untouched. Narrowcast OMI was 6.3% in the measurement and the optical back off about 7 dB.

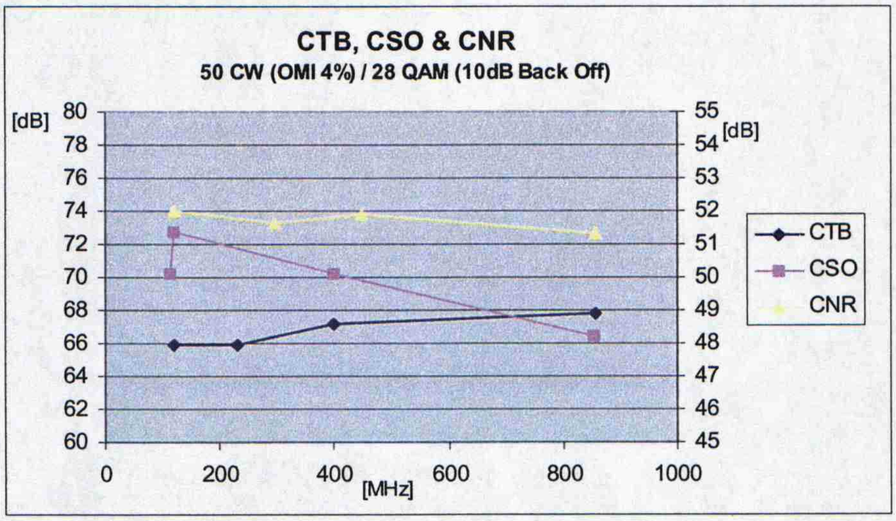


Figure 91. Noise and distortion performance of the broadcast link.

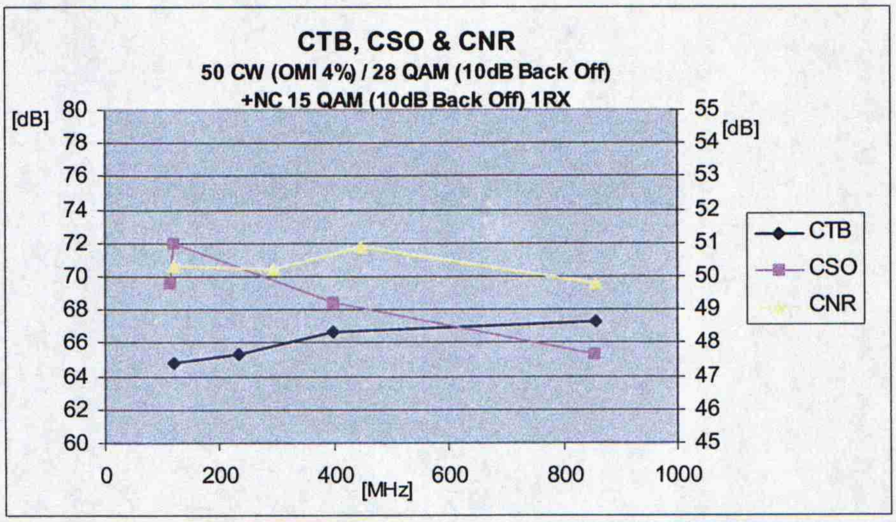


Figure 92. Noise and distortion performance of the link with narrowcast channels.

7.4 Test setup 2.

In setup 1, the OMI and the C/N of the narrowcast transmitter was limited by the chirp and dispersion induced CSO distortion. This problem can be overcome with separate receivers for broadcast and narrowcast light and a high pass filter in the narrowcast receiver. A red / blue filter is used to separate the wavelengths of the NC and BC transmitters. Test setup 2 was built to investigate this possibility.

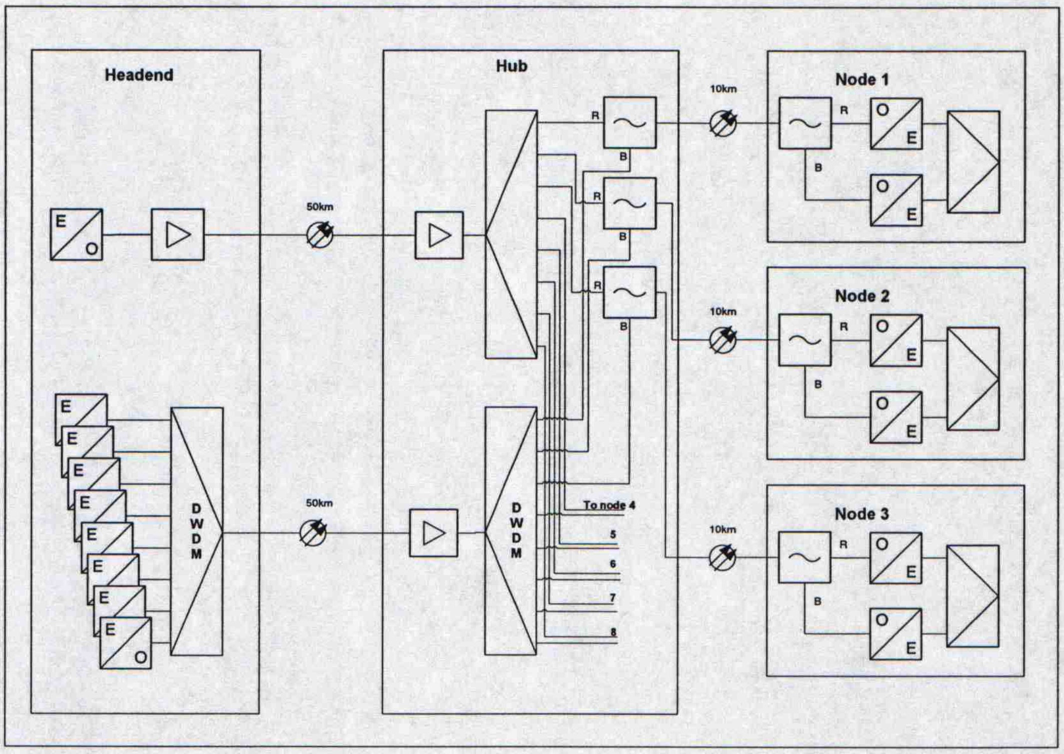


Figure 93. Outline of the setup 2 introducing the two receiver configuration.

Fig. 94 shows the RF spectrum of the narrowcast link with 33 QAM channels. In Fig. 95 the effect of filtering can be seen. Instead of the low frequency CSO the C/N is limited with CTB distortion approximately in the modulation frequencies.

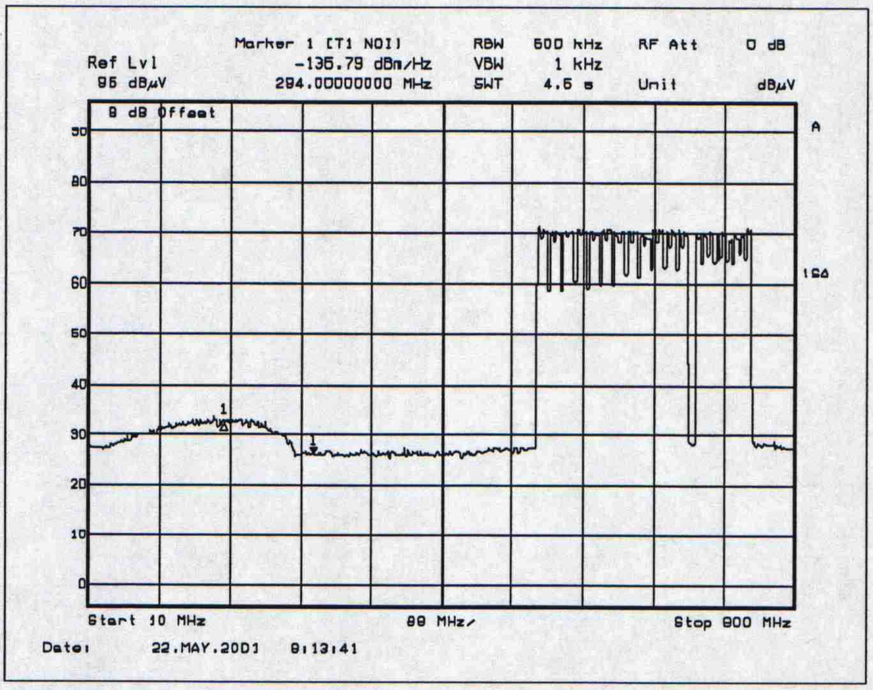


Figure 94. RF spectrum of the narrowcast link (33 QAM channels).

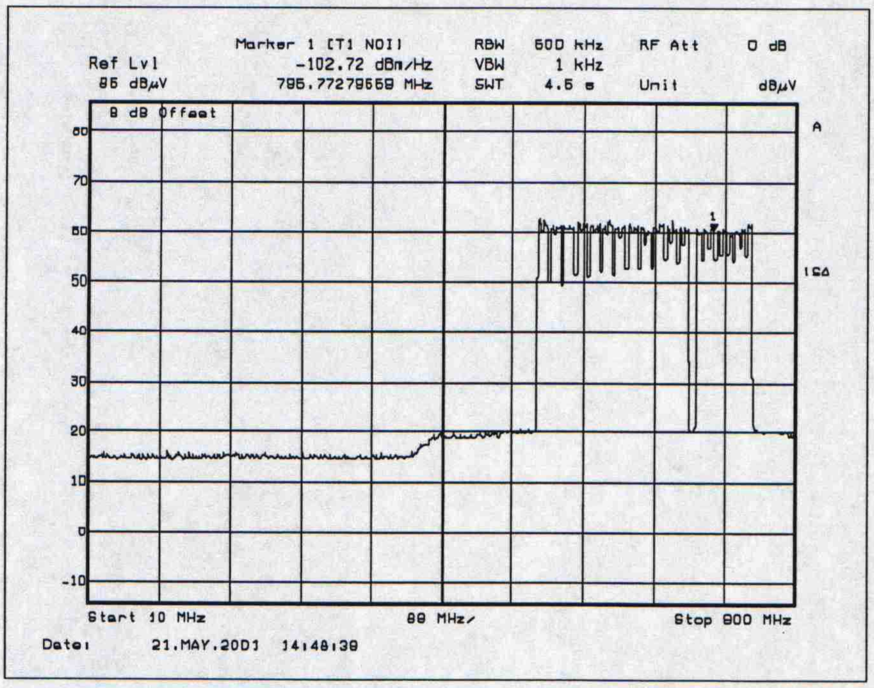


Figure 95. Filtered RF spectrum of the narrowcast link.

Fig. 96 presents the measured performance of the link with 2 receivers for the same amount of channels than in test setup 1 (Fig. 92). Filtering of the CSO allows higher OMI and about 1.5 dB larger optical back off can be used. As a result ~0.5 dB better C/N can be achieved, but in low frequencies the filtering of CSO gives close to 2 dB better C/N.

With 2 receivers the performance level can be maintained if the number of narrowcast channels is doubled (Fig. 97). Low frequency C/N is considerably smaller if no high-pass filtering is used (Fig. 98).

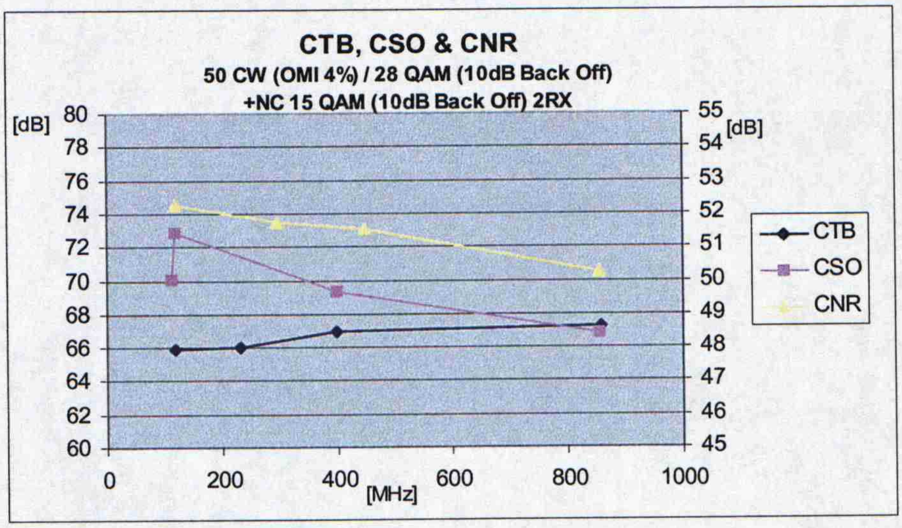


Figure 96. Noise and distortion performance of the link with 15 NC channels.

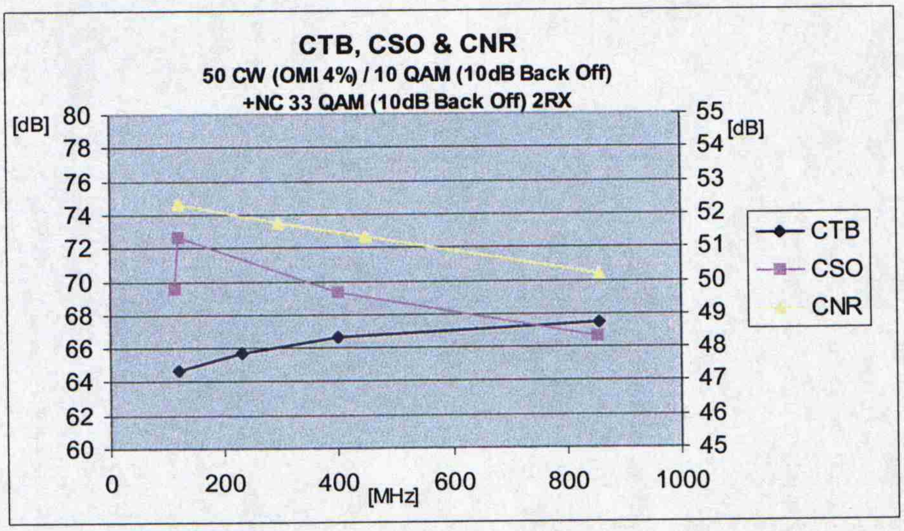


Figure 97. Noise and distortion performance of the link with 33 NC channels.

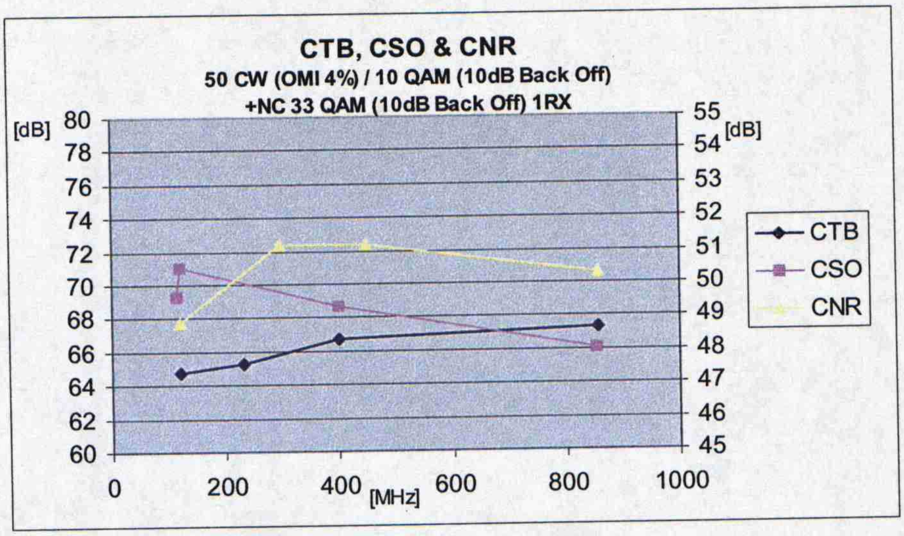


Figure 98. Noise and distortion performance of the link with 33 NC channels. The high pass filter is removed.

7.5 Test setup 3.

In the 3rd setup the effect of using only one fibre from headend to hub was investigated. The outline of the setup is shown in the Fig 99. Some nonlinear effects in the long fibre with relatively high light powers can be measured, but no harm to transmitted signals can be seen. No reason for avoiding this configuration was found.

With this setup also the effect of coaxial amplifier cascade after the link was measured. The outline of the cascade is shown in Fig. 100.

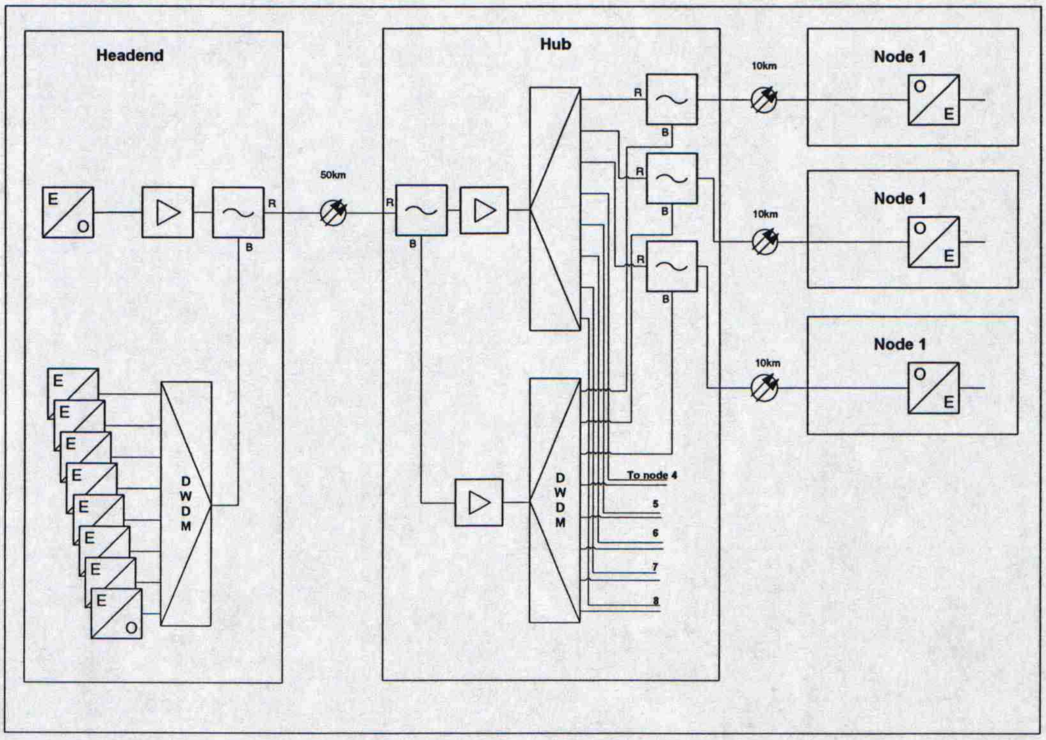


Figure 99. Outline of the test setup 3. BC and NC in the same fibre.

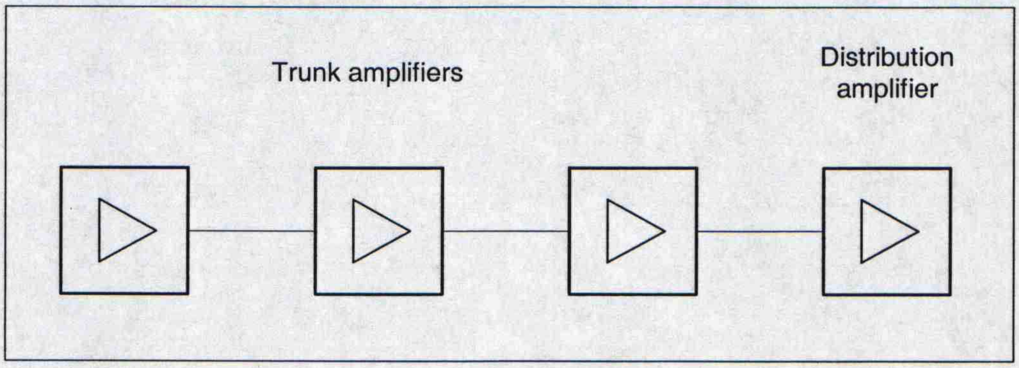


Figure 100. Coaxial cable amplifier cascade of 3 trunk amplifiers and a distribution amplifier.

Fig. 101 shows crosstalk from the BC channels to the NC signal. As it is located in the low end of the spectrum, it is most likely due to SRS. The crosstalk seen in the picture is not dangerous as it will be ~60 dB lower than the broadcast channels after summing of the BC and NC signals. 60 dB below the carrier is unacceptable for any disturbing signal in analogue TV, but here the disturbing signal has the same modulation, and the delay compared to the main signal is small. Therefore, the crosstalk will not be visible in the picture, and can be tolerated. Fig. 102 shows the acceptable level for the double reflected signal as the function of the delay. Propagation with two different velocities in different wavelengths is analogous to double reflection as some part of the signal will be received with a delay. This delay is very small if it is due to dispersion only. If the other wavelength has physically longer distance by 50 meters, this delay would be 250 ns and still -25 dB of delayed signal would be acceptable. The crosstalk caused by SRS and XPM is harmless, but it may look alarming.

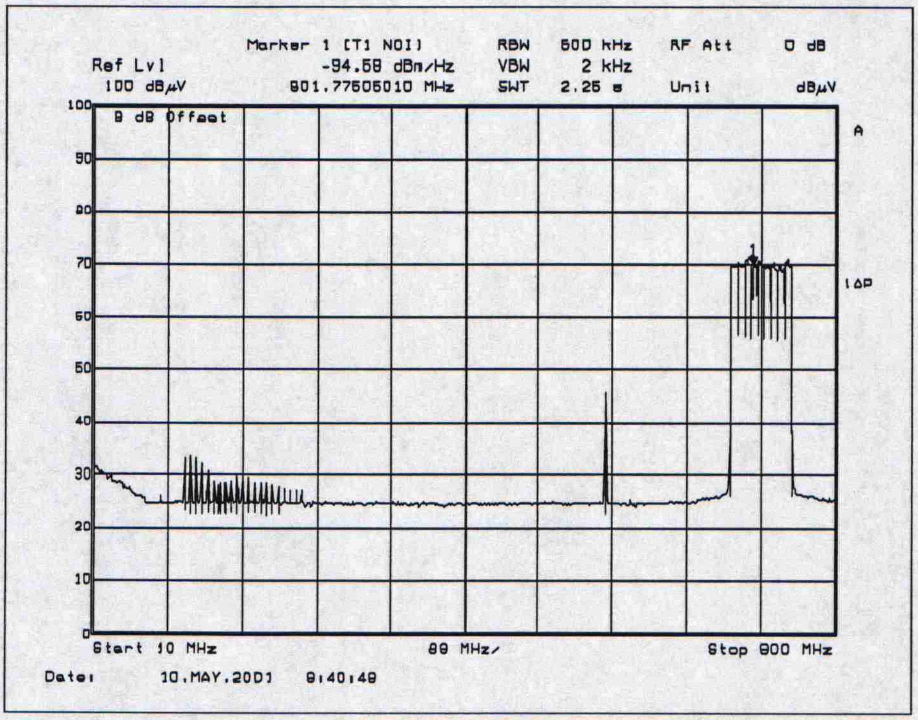


Figure 101. Some analogue carriers can be seen on the narrowcast wavelength.

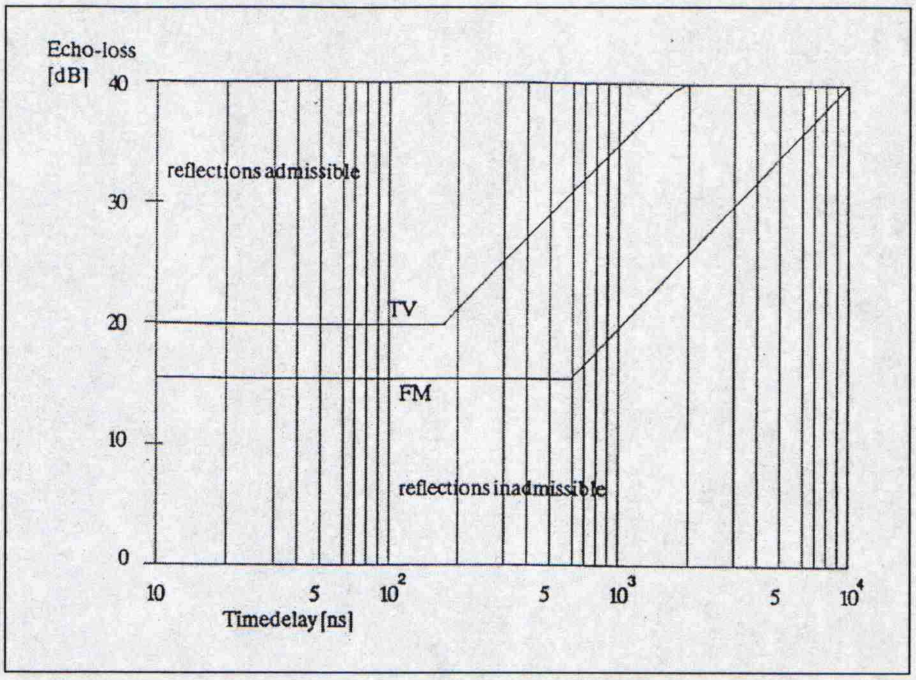


Figure 102. Acceptable level of the double reflected signal as function of time delay /23/.

Fig. 103 shows the measured RF spectrum after the link and the coaxial amplifier cascade. The spectrum now has a slope, since this is advantageous in coaxial cable transmission.

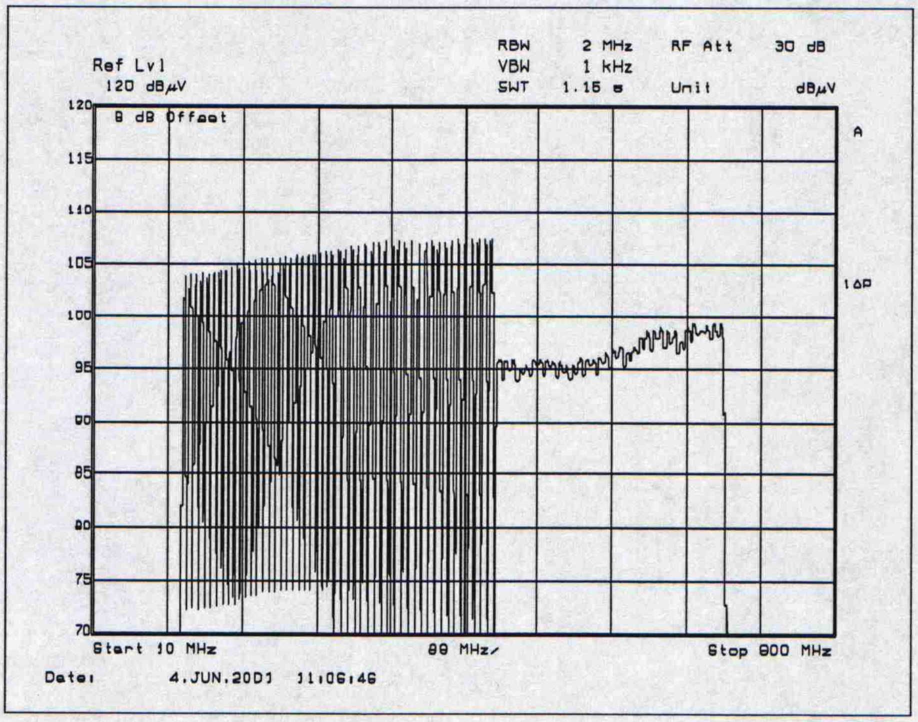


Figure 103. Measured RF-spectrum of the link and the cascade with load of 50 analogue channels and 34 QAM channels.

In this test setup 256-QAM was used. A 256-QAM channel requires four times more power than a 64-QAM channel, but gives just 33% better bit rate. Greater power per channel means that fewer channels can be used to load NC transmitter. 12 channels on the NC transmitter still gave satisfactory results. If 64-QAM would have been used, the whole digital load of 34 QAM channels could have been on the NC transmitter. The total bit rate on all 8 wavelengths with 64-QAM could be over 10 Gbit/s, but is in this case less than 7 Gbit/s. Thus it is often better to use 64-QAM instead of 256-QAM.

Measured noise and distortion results are presented in Figs. 104-106. The three trunk amplifiers at relatively low signal level add practically no distortion, but 1dB of C/N is lost(see Fig. 105). Distribution amplifier operated at a high signal level adds some distortion, but almost no noise (see Fig. 106).

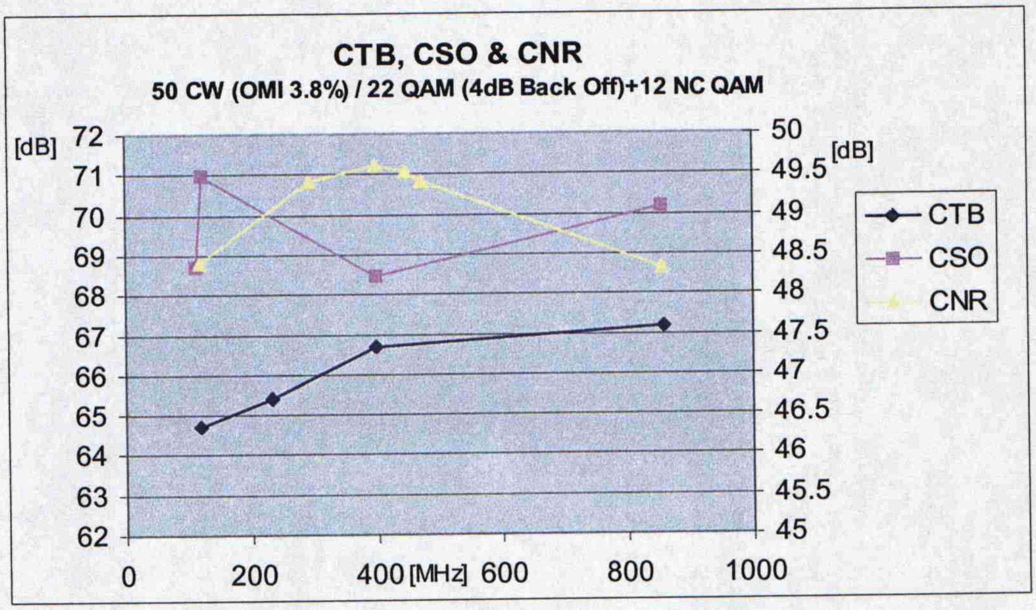


Figure 104. Measured noise and distortion performance of the test setup 3.

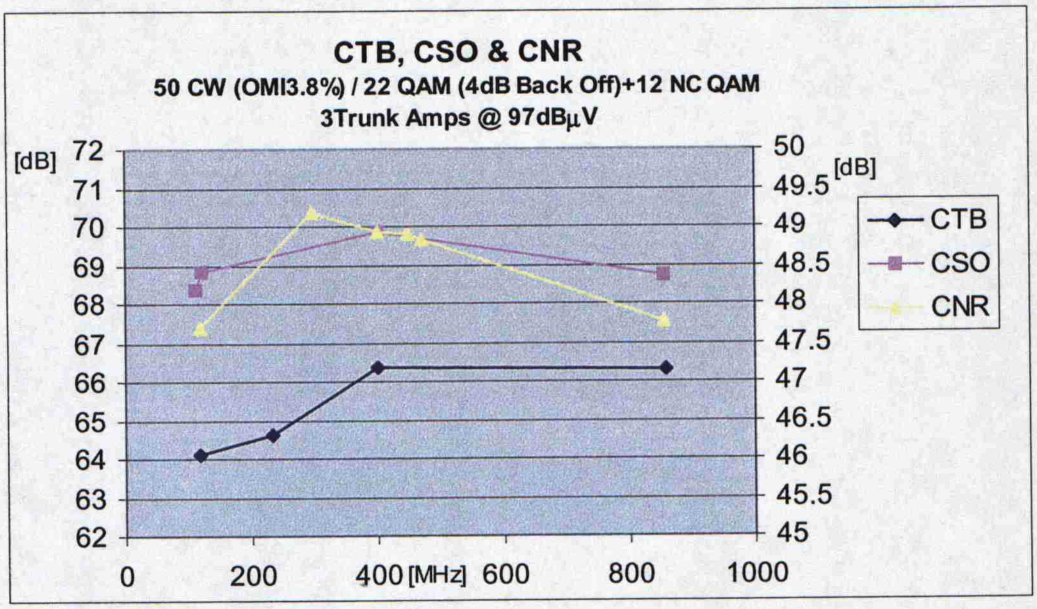


Figure 105. Measured noise and distortion performance of the test setup 3 with the effect of 3 trunk amplifiers included.

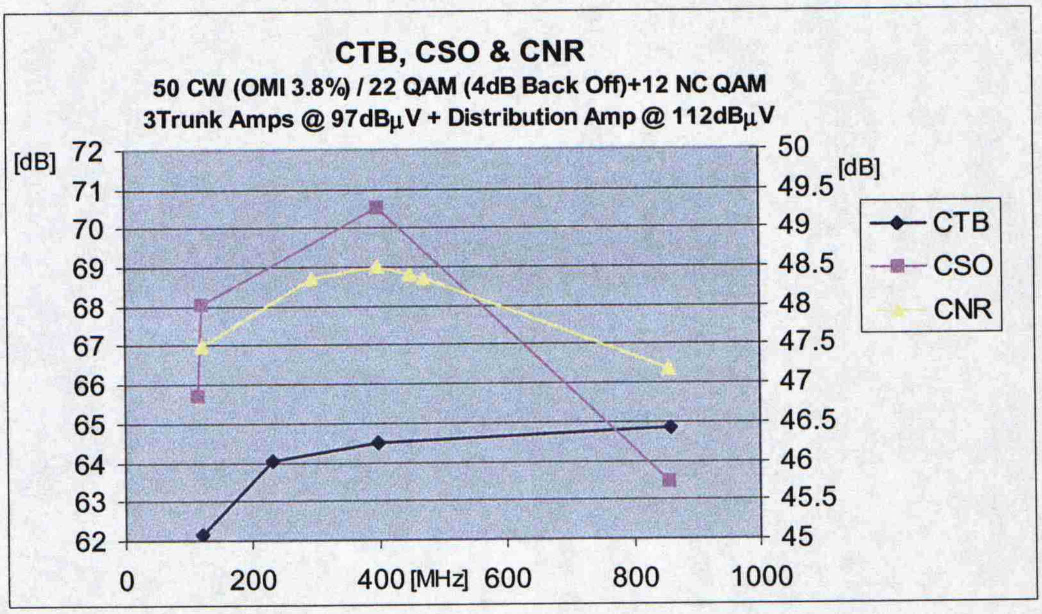


Figure 106. Measured noise and distortion performance of the test setup 3 with the effect of 3 trunk amplifiers and a distribution amplifier included.

QAM signal quality was checked by measuring the C/N. For this test using the 256-QAM measured C/N values were between 38 and 42 dB. Example result for the C/N measurement is shown in Fig. 107. Measurement of MER would not give any significant information as it is still mostly limited by the quality of the modulator used and not by the network.

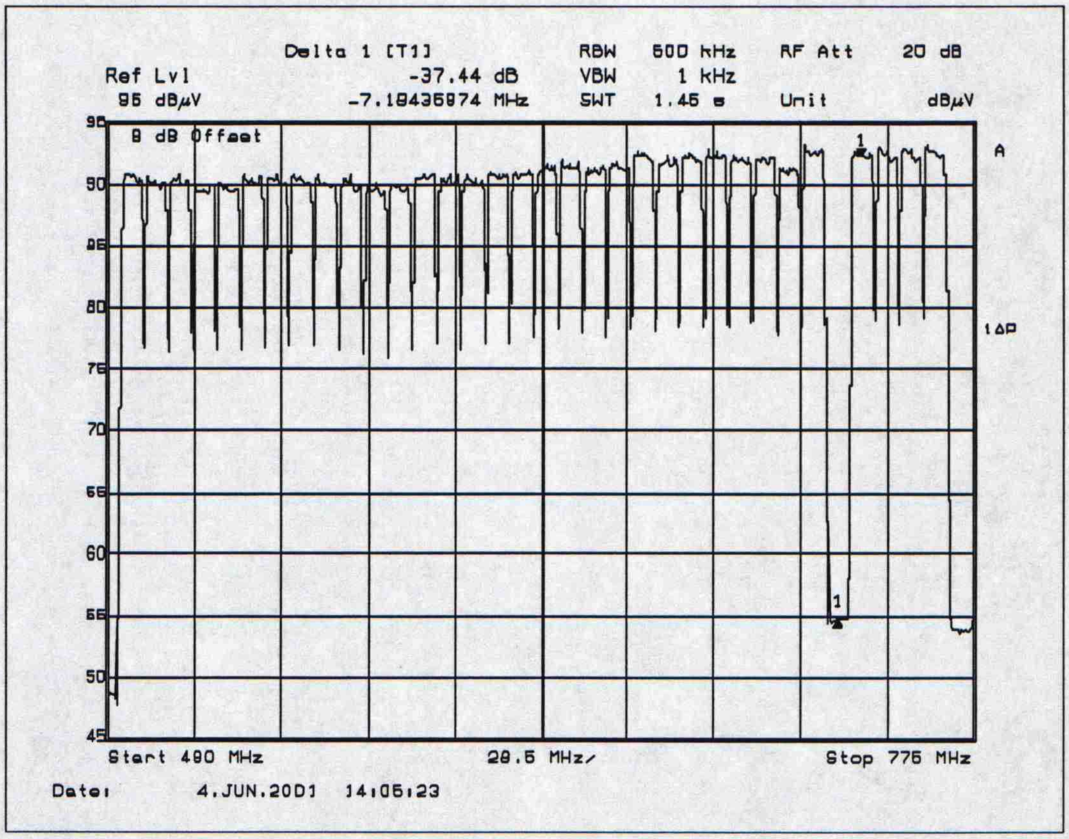


Figure 107.Example measurement of the C/N @ 730 MHz.

Similar tests with 64-QAM were performed as well. The amount of channels in the narrowcast was increased to 33. Performance of the network was good again. The only problem is the C/N at low frequency analogue channels, which is ruined by the NC CSO that could be filtered out if two receivers were used.

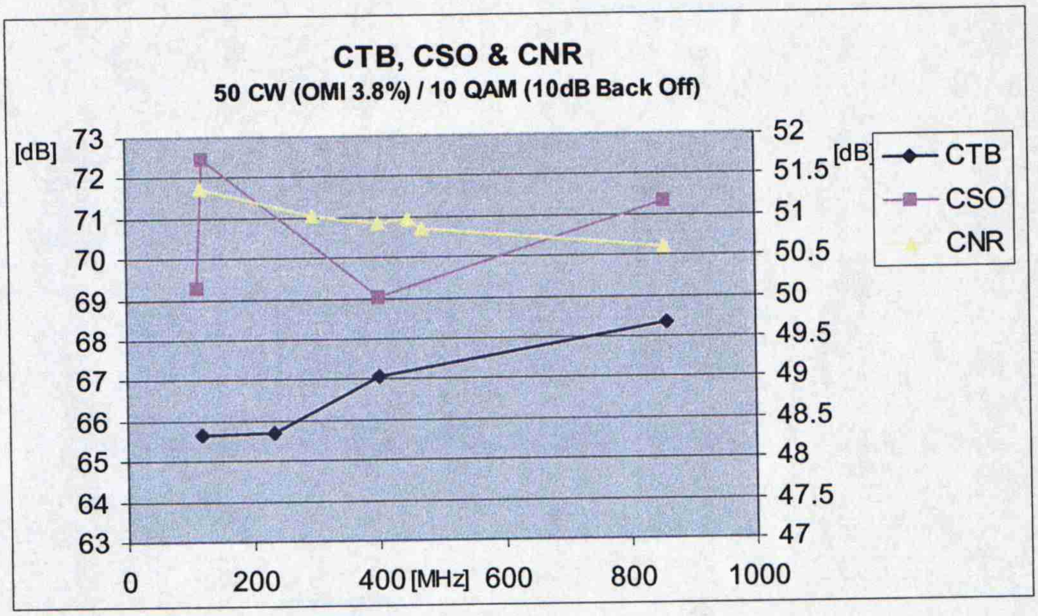


Figure 108. Measured noise and distortion performance of the test setup 3. Only broadcast signal.

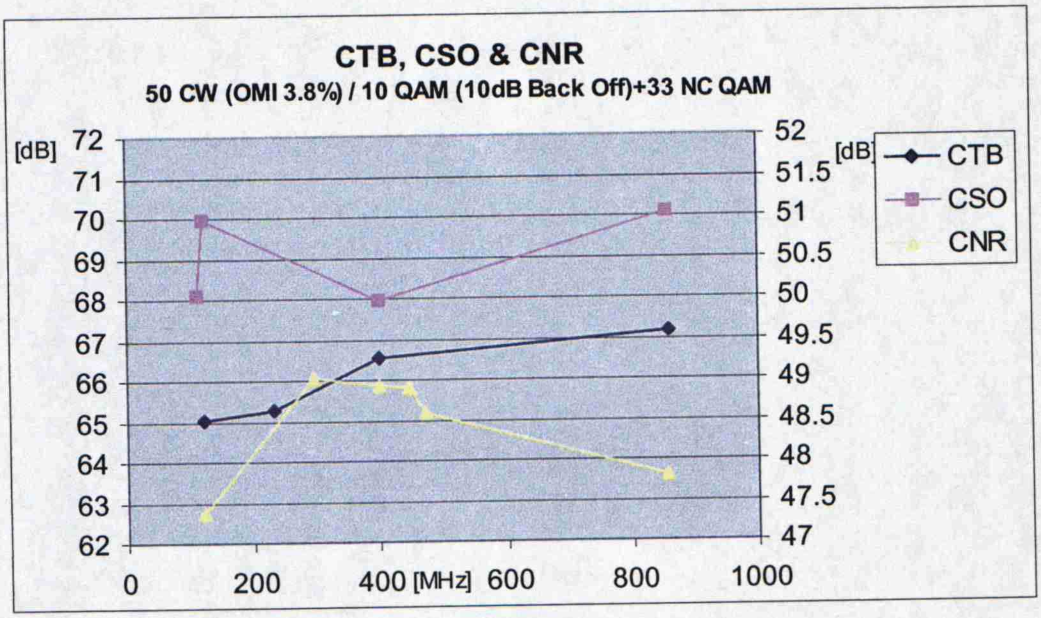


Figure 109. Measured noise and distortion performance of the test setup 3. Both BC and NC signals.

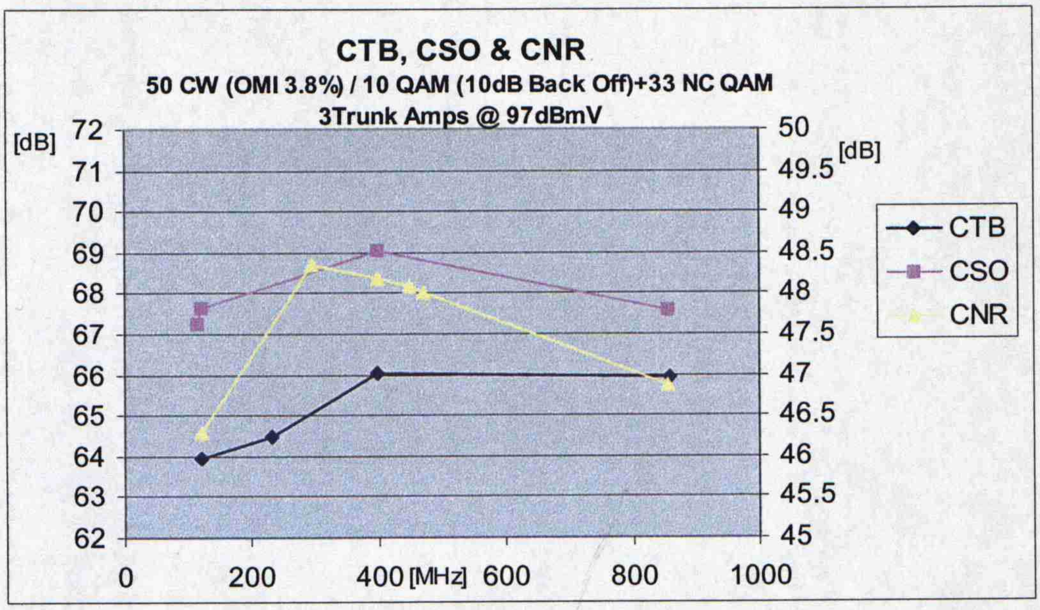


Figure 110. Measured noise and distortion performance of the test setup 3. The effect of 3 trunk amplifiers.

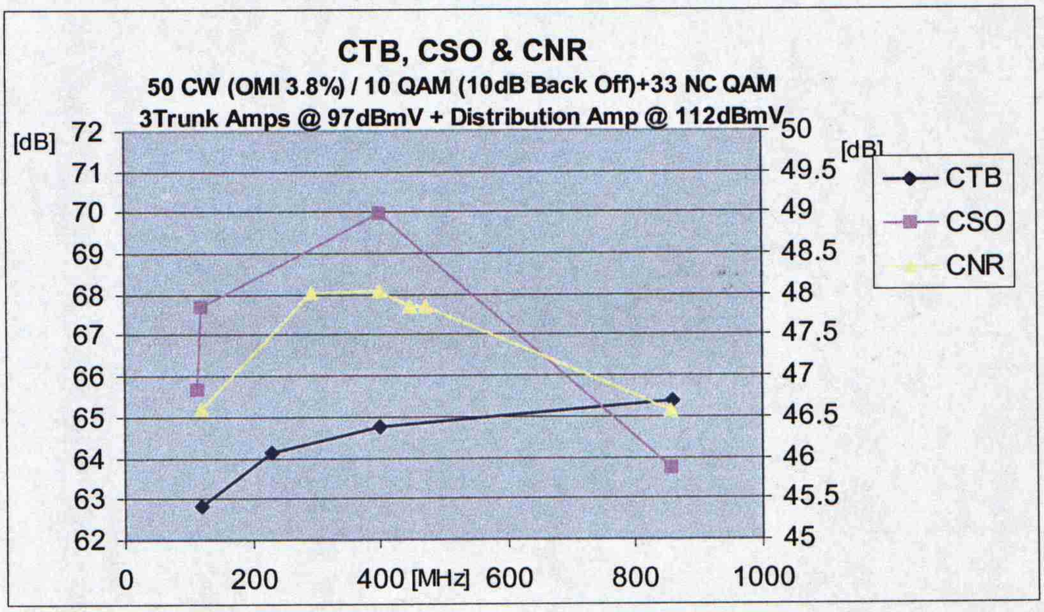


Figure 111. Measured noise and distortion performance of the test setup 3.The effect of 3 trunk amplifiers and a distribution amplifier included.

Fig 112 presents the measured C/N performance for the QAM channels. C/N is ~35 dB for all frequencies. This performance gives good signal quality and deterioration margin.

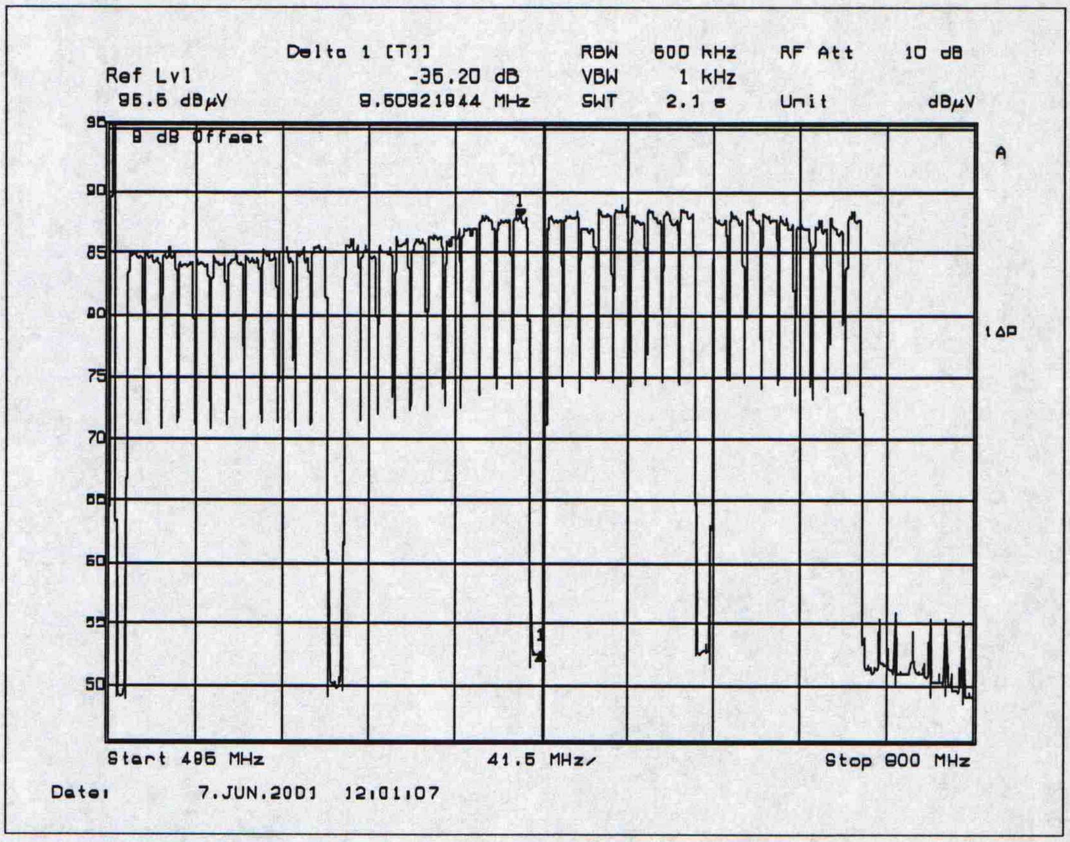


Figure 112. Measurement of the C/N performance. The C/N is 35 dB for all QAM channels with 10 dB Back Off

8 CONCLUSION

Transmission of DVB-C signals with the DWDM with the fibre optic network was investigated in this work. It was shown that this modulation format can be transmitted using directly modulated 1550 nm lasers. The typical existing service with ~50 analogue channels can be upgraded by adding narrow cast services with directly modulated DWDM transmitters. Penalty in the C/N of the existing service by 0.5 – 1.5 dB is acceptable in most cases. Lowering the number of analogue carriers in the system can be used to compensate the deterioration. The bit rate per wavelength can be over 1Gbit/s, and equivalent to over 200 video streams.

Most severe technical limitation to DWDM transmission is the chirp and dispersion induced distortion. This sets the limit to the number of channels that can be used with the directly modulated transmitters. Typically, 10-30 QAM channels can be transmitted with a directly modulated laser. Using two receivers and a high-pass filter to remove the low frequency CSO of the NC, the number of QAM channels can be increased.

EDFA gain tilt may cause problems with long links, but with a proper selection of equipment and network topology the EDFA gain shape is not a problem. With high launch powers, the effects of SRS and XPM can be observed, but they should not limit the transmission quality.

64-QAM is a robust modulation method and adequate C/N performance is quite easy to accomplish in CATV networks. 256-QAM is much more demanding and capable of carrying only 33 % more information. Better and more reliable performance can be achieved by using more 64-QAM signals at lower level. In many practical cases the total capacity of the network will be larger with 64-QAM than with 256-QAM, since more channels can be allocated to narrowcast transmitters. For these reasons, 256-QAM should only be used when available bandwidth is limited.

In return-channel applications, the environmental temperature is a problem for the cooling of the laser component. Besides the heat, the physical size can be a limiting factor in upstream applications. The cost of DWDM is also higher than the industry is accustomed to pay for the return channel equipment.

Added complexity of the network makes the maintaining and installation of the network more difficult. More than technical limitations, the lack of skilled staff to operate the DWDM networks can be a factor in preventing the building of these networks.

With a typical node size of today, 500 homes passed per optical node, true video on demand services can be realized with QAM modulated DWDM narrowcast services. Downstream data capacity to each home in the node area is also large, comparable to that of any other media.

REFERENCES

/1/ CENELEC standard EN 300 429, "Digital Video Broadcasting (DVB); Framing structure, channel coding and modulation for cable systems", European Committee for Electrotechnical Standardisation, 1998.

/2/ Govind P. Agraval, *Fiber-Optic Communcation Systems*, Wiley Series in Microwave and Optical Engineering, John Wiley & Sons, Inc., 1997.

/3/ Walter Ciciora, James Farmer, David Large, *Modern Cable Television Technology*, The Morgan Kaufmann Series in Networking, Morgan Kaufmann Publishers, Inc. 1999

/4/ G. C. Wilson, J.M Delavaux, A. Srivastava, Cyril Hullin, C. McIntosh, C. G. Bethea, C. Wolf, "Long-haul DWDM/SCM transmission of 64-QAM using electroabsorption modulated laser transmitters", ECOC 2000, Post-deadline papers, 2000.

/5/ Datasheet, Fujitsu, FID3Z1KX/LX.

/6/ ITU-T Recommendation G.692,"Transmission media characteristics – Characteristics of optical components and sub-systems", International Telecommunication Union, 1998.

/7/ Dennis Dericson, *Fiber Optic Test and Measurement*, Hewlett-Pacard Professional Books, New Jersey, Prentice Hall, 1998.

/8/ Winston I. Way, *Broadband Hybrid Fiber/Coax Access System Technologies*, Academic Press, 1999.

/9/ CENELEC standard EN50083-6, "Cabled distribution systems for television and sound signals Part 6: Optical equipment", European Committee for Electrotechnical Standardisation, 1997.

/10/ K.Petterman, "FM-AM noise conversion in dispersive single-mode fibre transmission lines", *Electron. Lett.*, 1990, vol.26, no.25, pp. 2097-2098

/11/ Juha Nikkanen, ”Kaapelitelevisiovahvistimen epälineaarisuus ja verkon mitoitus suurilla kanavamäärillä” , Lisensiaatintyö, Teknillinen korkeakoulu, Sähkötekniikan osasto, 1991.

/12/ R. S. Tucker, “High-speed modulation of semiconductor lasers” , *J. Lightwave Technol.*, 1985, Vol. LT-3, No. 6, pp. 1180-1192.

/13/ Pi-Yang Chiang, Winston I. Way, “Ultimate Capacity of a Laser Diode in Transporting Multichannel M-QAM Signals”, *J. Lightwave Technol.*, 1997, Vol. 15, No. 10, pp. 1914-1924.

/14/ Frank S. Yang, Michel E. Marhic and Leonid G. Kazovsky, “Nonlinear Crosstalk and Two Countermeasures in SCM-WDM Optical Communication Systems,” *J. Lightwave Technol.*, 2000, Vol. 18, No.4.

/15/ Mary R. Phillips and Daniel M. Ott, “Crosstalk Due to Optical Fiber Nonlinearities in WDM CATV Lightwave Systems,” *J. Lightwave Technol.*, 1999, Vol. 17, No.10.

/16/ Rajiv Ramaswami and Kumar N. Sivarajan, *Optical Networks: A Practical Perspective*, Morgan Kaufmann Publishers, Inc., 1998.

/17/ H. Schmuck, “Comparison of optical millimetre-wave system concepts with regard to chromatic dispersion”, *Electron. Lett.*, 1995, Vol. 31, No. 21, pp. 1848-1849.

/18/ CENELEC Standard EN50083-3, “Cabled distribution systems for television and sound signals Part 3: Active coaxial wideband distribution equipment”, European Committee for Electrotechnical Standardisation, 1998.

/19/ ETSI Technical Report ETR 290, “Digital Video Broadcasting (DVB); Measurement guidelines for DVB systems,” European Telecommunications Standards Institute, 1997.

/20/John G. Proakis, *Digital communications*, McGraw-Hill Book Company, 1983.

/21/ Simulation program, Rohde & Schwarz WinIQSIM™, http://www.rohde-schwarz.de/www/dev_center.nsf/html/1117117down.

/22/ Tapio Niemi, "Measurements of time-resolved frequency chirp in modulated laser diodes," Masters thesis, Helsinki University of Technology, Metrology Research Institute,1997.

/23/ CENELEC Standard EN50083-7, "Cabled distribution systems for television and sound signals Part 7: System performance", European Committee for Electrotechnical Standardisation, 1996.

/24/ CENELEC Standard EN50083-7/A1, "Cabled distribution systems for television signals, sound signals and interactive services Part 7: System performance", European Committee for Electrotechnical Standardisation, 2000.

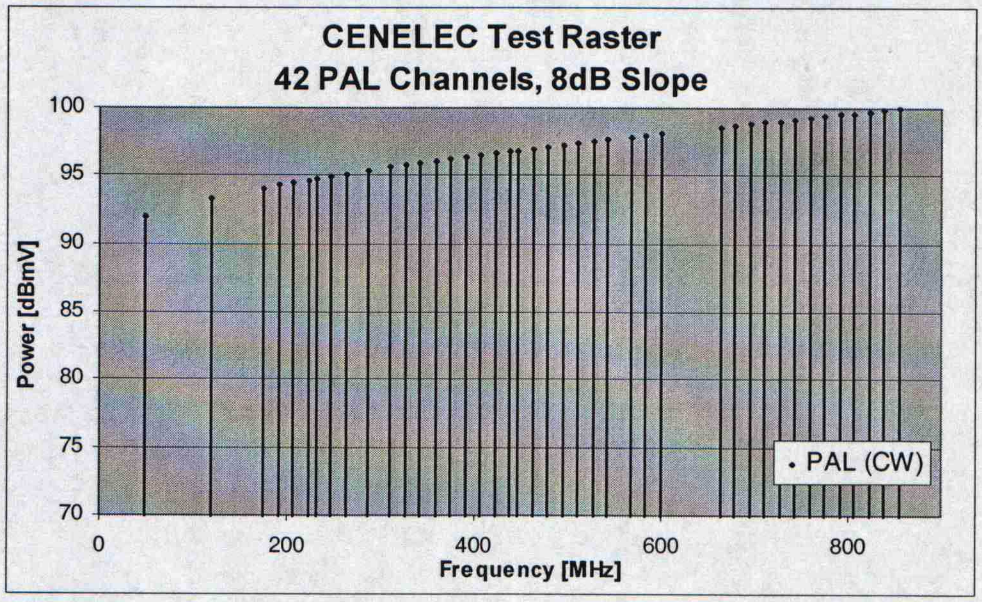
APPENDICES

Appendix A

CENELEC 42 CHANNEL TEST RASTER

The 42 frequencies of the CENELEC test raster are /18/:

48.25 MHz, 119.25 MHz, 175.25 MHz, 191.25 MHz, 207.25 MHz, 223.25 MHz, 231.25 MHz, 247.25 MHz, 263.25 MHz, 287.25 MHz, 311.25 MHz, 327.25 MHz, 343.25 MHz, 359.25 MHz, 375.25 MHz, 391.25 MHz, 407.25 MHz, 423.25 MHz, 439.25 MHz, 447.25 MHz, 463.25 MHz, 479.25 MHz, 495.25 MHz, 511.25 MHz, 527.25 MHz, 543.25 MHz, 567.25 MHz, 583.25 MHz, 599.25 MHz, 663.25 MHz, 679.25 MHz, 695.25 MHz, 711.25 MHz, 727.25 MHz, 743.25 MHz, 759.25 MHz, 775.25 MHz, 791.25 MHz, 807.25 MHz, 823.25 MHz, 839.25 MHz, 855.25 MHz.



Raster CENELEC42 8 dB sloped.

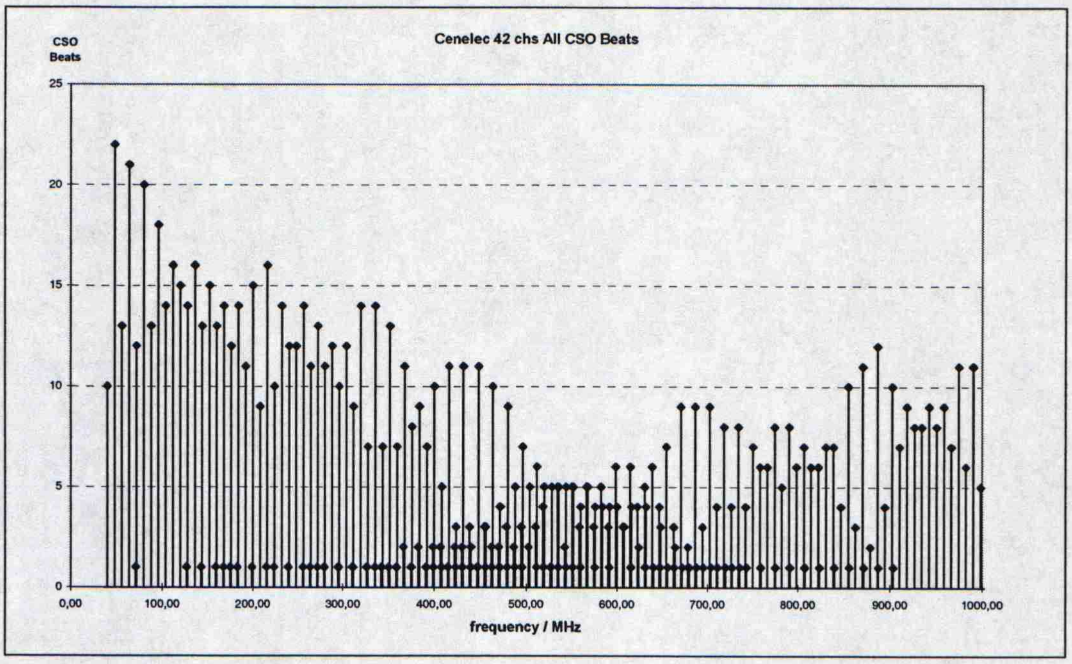


Fig.113, All CSO Beats up to 1000 MHz.

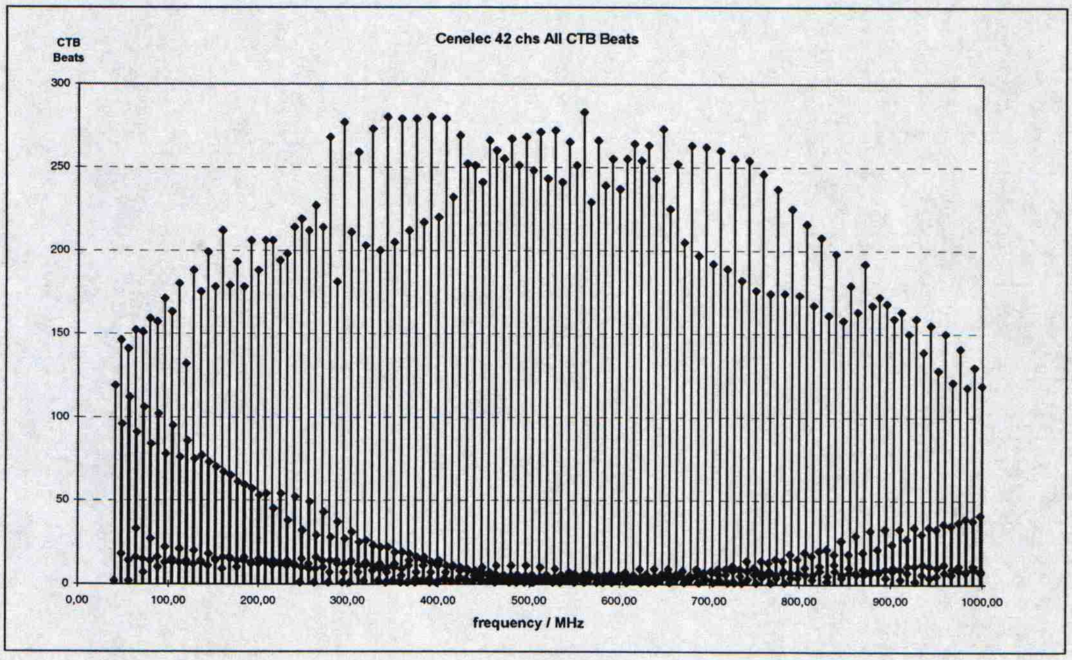


Fig.114, All CTB Beats up to 1000 MHz

Appendix B

ITU-GRID

CH	Freq (GHz)	Wavelength (nm)	CH	Freq (GHz)	Wavelength (nm)
15	191500	1565.50	44	194400	1542.14
16	191600	1564.68	**45**	**194500**	**1541.35**
17	191700	1563.86	46	194600	1540.56
18	191800	1563.05	**47**	**194700**	**1539.77**
19	191900	1562.23	48	194800	1538.98
20	192000	1561.42	**49**	**194900**	**1538.19**
21	192100	1560.61	50	195000	1537.40
22	192200	1559.79	**51**	**195100**	**1536.61**
23	192300	1558.98	52	195200	1535.82
24	192400	1558.17	**53**	**195300**	**1535.04**
25	192500	1557.36	54	195400	1534.25
26	192600	1556.55	**55**	**195500**	**1533.47**
27	192700	1555.75	56	195600	1532.68
28	192800	1554.94	**57**	**195700**	**1531.90**
29	192900	1554.13	58	195800	1531.12
30	193000	1553.33	**59**	**195900**	**1530.33**
31	193100	1552.52	60	196000	1529.55
32	193200	1551.72	61	196100	1528.77
33	193300	1550.92	62	196200	1527.99
34	193400	1550.12	63	196300	1527.22
35	193500	1549.32	64	196400	1526.44
36	193600	1548.51	65	196500	1525.66
37	193700	1547.72	66	196600	1524.89
38	193800	1546.92	67	196700	1524.11
39	193900	1546.12	68	196800	1523.34
40	194000	1545.32	69	196900	1522.56
41	194100	1544.53	70	197000	1521.79
42	194200	1543.73	71	197100	1521.02
43	194300	1542.94	72	197200	1520.25

Bold and underlined channels were used in this work.

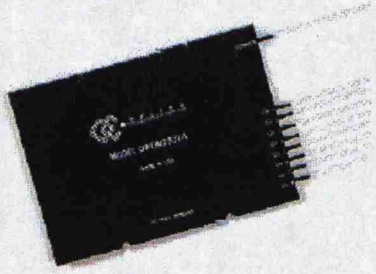
Appendix C

DWDM FILTER SPECIFICATION

200 GHz 8-Channel Dense Wavelength Division Multiplexer
DWDM-2 Series

Home
DWDM
Wavelength Locker
OADM
Micro-Optic WDM
Hybrid Components
Isolators
Fused WDM
Fused Couplers
Gain-Flattening Filter

Features
- 200 GHz Channel Spacing
- High Channel Isolation
- Low Insertion Loss
- Highly Stable & Reliable
- Epoxy-Free Optical Path
- Low Profile Packaging



Applications
- Dense WDM Systems
- CATV Fiberoptic Links
- Fiber Amplifier System
- Fiberoptic Instruments
- Sensor

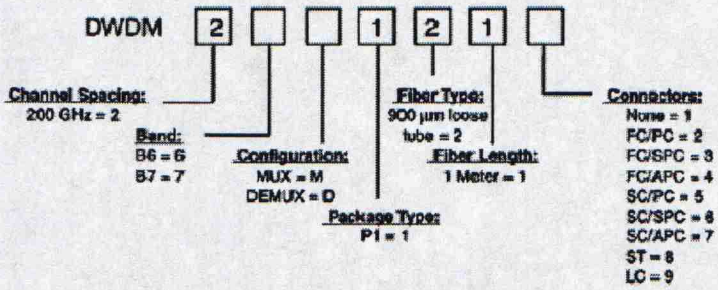
Performance Specifications

DWDM-2 Series	Multiplexer	De-Multiplexer
Center Wavelength (nm)	1550 nm Window, ITU 100 GHz Grid B6: 1531.90, 1533.47, 1535.04, 1536.61, 1538.19, 1539.77, 1541.35, 1542.94 B7: 1547.72, 1549.32, 1550.92, 1552.53, 1554.14, 1555.75, 1557.37, 1558.99	
Channel Spacing (GHz)	200 (1.6 nm)	
Channel Passband (-0.5 dB Bandwidth) (nm)	≥ 0.5	
Insertion Loss (dB)	≤ 5.0	≤ 6.0
In-band Loss Variation (dB)	≤ 1.0	≤ 1.0
Adjacent Channel Isolation (dB)	≥ 25	≥ 30
Non-adjacent Channel Isolation (dB)	≥ 40	
Insertion Loss Temperature Sensitivity (dB/°C)	≤ 0.004	
Wavelength Temperature Sensitivity(nm/°C)	≤ 0.001	
Return Loss (dB)	≥ 45	
Polarization Dependent Loss (PDL) (dB)	≤ 0.15	
Polarization Mode Dispersion (ps)	≤ 0.1	
Maximum Power Handling (mW)	300	
Maximum Tensile Load (N)	5	
Operating Temperature (°C)	0 to +65	
Storage Temperature (°C)	-40 to +85	
Package Dimension (mm)	P1: (L) 140 x (W) 100 x (H) 16	

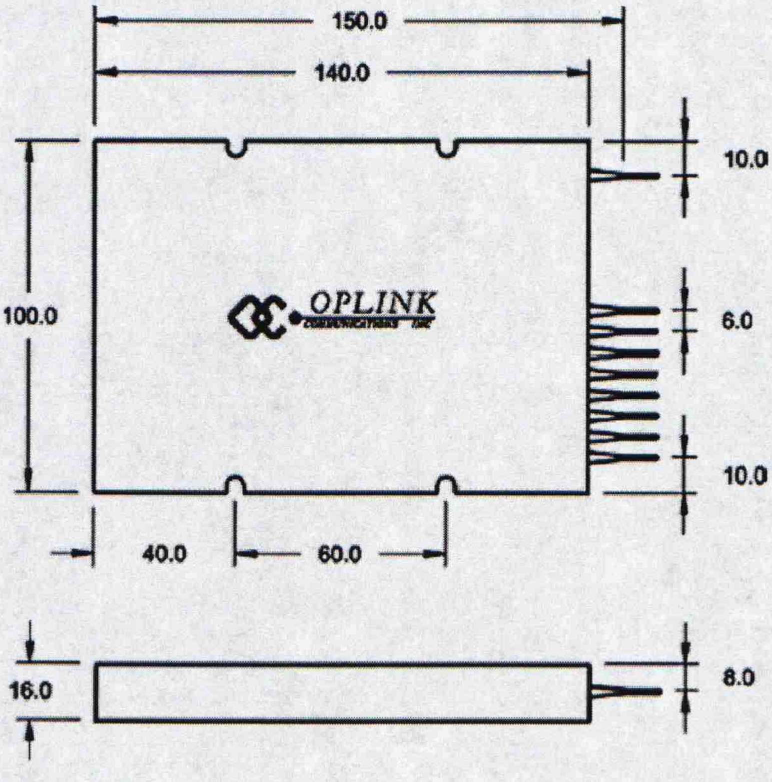
OPLINK dense wavelength division multiplexer (DWDM) is based on thin film filter technology and metal bonding micro optics packaging. This proven technology offers flat channel bandwidth, flexible channel configuration, low insertion loss and high isolation. The flexibility in channel configuration allows capacity upgrade as you grow. OPLINK DWDMs are modular units designed for easy upgrade. All OPLINK products are epoxy-free in the optical path. It is used in high-power applications in DWDM systems. OPLINK DWDM devices are Bellcore GR -1209 qualification tested.

Ordering Information

Standard Oplink Dense WDMs are specified below with appropriate part number options. If you require devices different than these standard devices, please contact our local Oplink Distributor, Sales Representative, or Factory Customer Service for ordering and price information

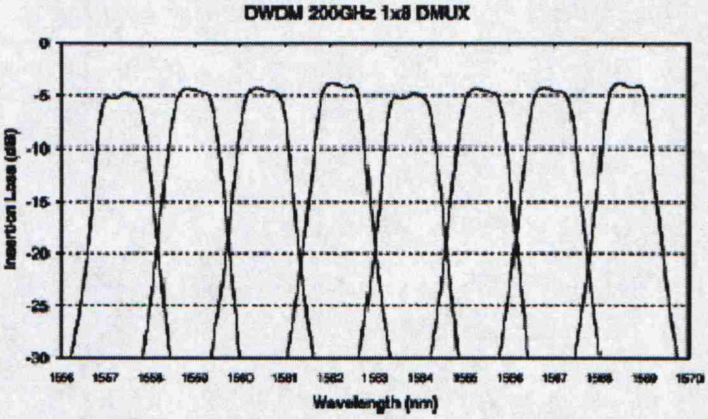


Package Dimensions:



Units: mm

Spectra:



DWDM2/990111.R08

Appendix D

DWDM LASER SPECIFICATION

Tentative Specification 9922 158 ###55 January 1999

CQF933/81##

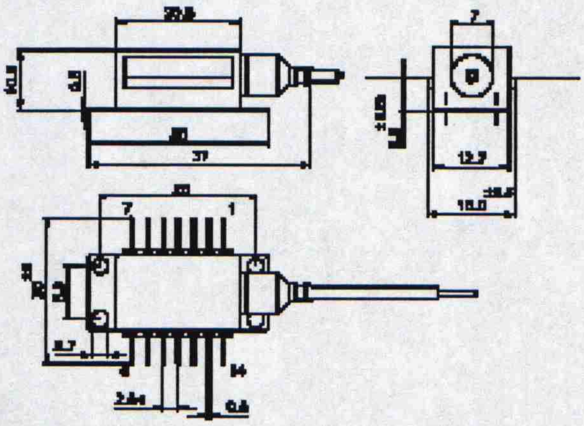
1550nm DFB laser with SMF fiber for Analog WDM

FEATURES

- 1550 nm DFB laser diode
- built-in thermo-electic cooler
- equipped with SC/APC connector
- cooled built-in optical isolator
- 1530 - 1564 nm wavelength range
- 1.6 nm (200 GHz) spacing

MECHANICAL DATA

Dimensions in mm



Pinning
1 thermistor
2 thermistor
3 LD cathode DC input via inductance
4 PD anode
5 PD cathode
6 cooler anode
7 cooler cathode
8 case GND
9 case GND
10 not connected
11 LD anode, case
12 LD cathode, AC input
13 LD anode, case
14 not connected

CHARACTERISTICS

($T_{chip} = T_{\lambda}$, T_{amb} = 25 °C, P_0 = 10 mW unless otherwise specified)

Symbol	Parameter	Conditions	Min	Typ	Max	Unit
I_{th}	threshold current		-	-	40	mA
λ_c	central wavelength, see table 1		1530	-	1561	nm
T_{λ}	laser set temperature for λ_c		15	-	35	°C
P_0	output power from pigtail	$T = T_{\lambda}$	10			mW
SMSR	side mode suppression ratio		30	-	-	dB
RIN	relative intensity noise		-	-	-157	dB/Hz

Uniphase Netherlands B.V.
Ordering information: 9922 158 ###55

Tentative Specification CQF933/81## 9922 158 ##55 January 1999

LIMITING VALUES

Symbol	Parameter	Conditions	Min	Max	Unit
Laser diode					
P_0	radiant output power from pigtail		-	15	mW
V_R	reverse voltage		-	2.0	V
I_F	forward current		-	200	mA
Monitor diode					
V_R	reverse voltage		-	20	V
I_F	forward current		-	10	mA
Module					
T_{op}	case operating temperature range	cooler active	-10	+65	°C
T_{stg}	storage temperature range		-40	+70	°C
Fibre pigtail					
R	bending radius		30	-	mm
F	tensile strength fibre to case		-	5	N

CHARACTERISTICS ($T_{chip} = T_L$, $T_{amb} \leq 25$ °C, $P_0 = 10$ mW unless otherwise specified)

Symbol	Parameter	Conditions	Min	Typ	Max	Unit
Laser diode						
I_{th}	threshold current	25 °C	-	-	40	mA
I_{op}	operating current	EOL	-	-	150	mA
P_0	radiant output power from pigtail		10			mW
V_F	forward voltage		-	-	3	V
η	slope efficiency		0.1	0.15		W/A
λ_c	central wavelength, see table 1		1530	-	1561	nm
T_L	laser set temperature for λ_c		+15	-	+35	°C
ISO	optical isolation		30	-	-	dB
ORL	optical return loss		40	-	-	dB
RIN	relative intensity noise	50 kHz - 2.5 GHz	-	-	-157	dB/Hz
$\Delta\lambda$	spectral linewidth	FWHM	-	1.0	3.0	MHz
$\Delta\lambda/\Delta T_c$	wavelength temperature tunability	$I_F = I_{op}$, $T = T_L$	0.07	-	–	nm/°C
SMSR	side mode suppression ratio	$I_F = I_{op}$	30	-	-	dB
S_{21}	bandwidth	5 - 870 MHz	-1	-	+1	dB
FM	laser chirp		-	125	150	MHz/mA
IM2	second order distortion	$I_F = I_{op}$, 35% OMI, 60 km fiber at 42 MHz, F1 = 595.25MHz, F2 = 553.25 MHz	-	-	34	dBc
IM3	third order distortion	$I_F = I_{op}$, 35% OMI, 60 km fiber at 637.25 MHz, F1 = 595.25MHz, F2 = 553.25 MHz	-	-	44	dBc
Monitor diode ($V_R = 10$ V)						
R	monitor diode responsivity		5	50	-	mA/W
I_{md}	dark current	$V_R = 10$V	-	-	0.2	µA
TE	temperature tracking error	T_{case}	-10	-	+10	%
Thermistor						
R_{th}	resistance	$T_{th} = 25$ °C	9.5	10	10.5	kΩ
B	Thermistor constant		3800	-	4100	K
Thermo-electric cooler ($\Delta T = 45$ °C)						
I_{cool}	cooler current		-	-	1.2	A
V_{cool}	cooler voltage		-	-	2.4	V
SMF 28 equivalent fibre pigtail						
$\varnothing_{mf}$	mode field diameter		9.5	-	11.5	µm
$\varnothing_{cl}$	cladding diameter		122	-	128	µm
$\varnothing_{sc}$	diameter of secondary coating		0.8	0.9	1.0	mm
l	length of pigtail		1	-	-	m
	Type of connector			SC/APC		
Reliability						
$MTTF_L$	long term wavelength drift, see note 1	EOL: $\Delta\lambda = 0.2$ nm	25	50	-	yr

Notes

1 Mechanical integrity/environmental endurance tested according to Bellcore TR-NWT-000468 and MIL-STD-883
MTTF = Mean Time To Failure EOL = End Of Life

Table 1

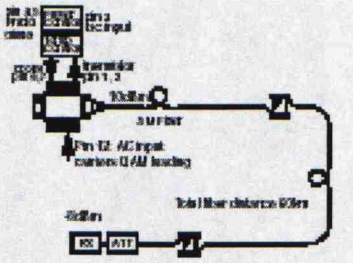
central wavelength (vacuum) λ_c (nm)	optical frequency f_c (THz)	channel #	TYPE (10 mW)	Ordering Code
1530.33	195.9	01	CQF933/8101	9022 158 57455
1531.90	195.7	03	CQF933/8103	9022 158 57355
1533.47	195.5	05	CQF933/8105	9022 158 57255
1535.04	195.3	07	CQF933/8107	9022 158 57155
1536.61	195.1	09	CQF933/8109	9022 158 57055
1538.19	194.9	11	CQF933/8111	9022 158 56955
1539.77	194.7	13	CQF933/8113	9022 158 56855
1541.35	194.5	15	CQF933/8115	9022 158 56755
1542.94	194.3	17	CQF933/8117	9022 158 56655
1544.53	194.1	19	CQF933/8119	9022 158 56555
1546.12	193.9	21	CQF933/8121	9022 158 56455
1547.72	193.7	23	CQF933/8123	9022 158 56355
1549.32	193.5	25	CQF933/8125	9022 158 56255
1550.92	193.3	27	CQF933/8127	9022 158 56155
1552.52	193.1	29	CQF933/8129	9022 158 56055
1554.13	192.9	31	CQF933/8131	9022 158 55955
1555.75	192.7	33	CQF933/8133	9022 158 55855
1557.36	192.5	35	CQF933/8135	9022 158 55755
1558.98	192.3	37	CQF933/8137	9022 158 55655
1560.61	192.1	39	CQF933/8139	9022 158 55555

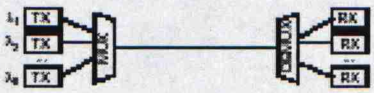
Tentative Specification **9922 158 ##55** **January 1999**

CQF933/81##

DWDM (dense wavelength division multiplexing) is rapidly gaining acceptance within HFC (hybrid fiber coax) network architectures as the solution for implementation of multi-format traffic and overlay technologies for additional interactive services like internet, video on demand, and telephony. DWDM also offers the opportunity for passive optical routing and insures that architectures are scalable for increased bandwidth needs in the near future. For these advanced HFC architectures Uniphase offers the CQF933, a DWDM laser with 10 mW output specifically designed for QAM (quadrature amplitude modulation) applications. The combination of low chirp, narrow linewidth and analogue properties offers excellent system performance with respect to CNR, CSO and CTB. The CQF933 is available with wavelengths covering the ITU grid from 1530 up to 1560 nm with 200 GHz spacing.

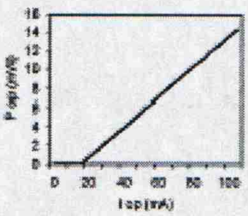
TYPICAL PERFORMANCE CHARACTERISTICS



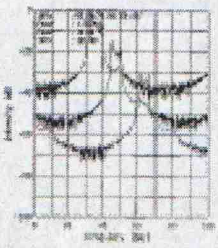


The CQF933 can be used to transmit channels with QAM loading over a distance of 60 km in a HFC network.

L - I curve

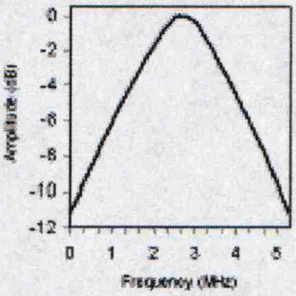


Chirp



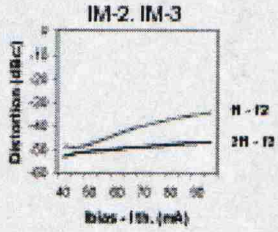
Typical chirp is better than 125 MHz/mA

Linewidth



Typical linewidth at -3dB is better than 2 MHz

IM-2, IM-3



F1 = 595 MHz, F2 = 553 MHz, modulation 35%. Second order and third order distorsion of the laser versus bias current. The laser operates at specified wavelength.