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AN OPTICAL TIME DOMAIN REFLECTOMETER
OPERATING AT 1.3 μ m WAVELENGTH

by

Lukas Hadellis, Dipl. Eng.

A thesis submitted to the Faculty of Graduate Studies and Research in
partial fulfillment of the requirements for the degree of

Master of Engineering

Department of Electronics

Faculty of Engineering

Carleton University

Ottawa, Canada

December 1983

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LIST OF SYMBOLS

P_f	Transmitted, forward travelling light power
P_b	Backwards travelling light power (backscatter + discrete reflections)
P_0	Light pulse peak power
T_{rep}	Repetition rate of light pulses
a (or a_t)	Fiber total attenuation
x (or z)	Distance
c	Group velocity of light in air
λ	Wavelength of light
n_1	Refraction index of the fiber core
n_2	Refraction index of the fiber cladding
n_0	Refraction index of the air
R	Reflection coefficient
P_a	Absorbed light power
P_s	Scattered light power
R_b	Backscatter coefficient
S	Capture fraction of the fiber
v	Group velocity of light in the fiber core
NA	Numerical apperture of the fiber
a_s	Rayleigh scattering coefficient
W	Width of the light pulse
a_{fs}	Forward scattering coefficient
a_{fa}	Forward total absorption coefficient
a_{ft}	Forward total attenuation coefficient
E_b	Backscattered light energy

I	Current
I_{th}	Threshold current of a laser diode
V_{ce}	The collector-emitter voltage of a transistor
M	Avalanche gain
η	Quantum efficiency of a photodiode
R	Responsivity of a photodiode
I_D	Dark current
I_{DM}	Multiplied dark current by the avalanche process
I_{DN}	Non-multiplied dark current
B	Bandwidth
q	Electron charge
R_L	Load resistance
C_d	Photodiode depletion layer capacitance
C_s	Stray capacitance
C_a	Preamplifier input capacitance
C_T	Total input capacitance in an optical preamplifier
I_s	Photocurrent
R_F	Feedback resistor in a transimpedance front-end
C_F	Feedback capacitor in a transimpedance front-end
Z_F	Effective feedback impedance
R_{in}	Total input resistance in an optical preamplifier
C_{gs}	Gate-to-source capacitance in a FET
A_v	Voltage gain
g_m	FET transconductance
R_D	Drain resistance
r_d	FET drain resistance
σ_{p0}	Transmitter noise standard deviation
σ_n	Receiver noise standard deviation

i_d	Dark current noise source
i_l	Thermal noise source
i_a	Preamplifier noise current source
e_a	Preamplifier noise voltage source
Y_a	Preamplifier input admittance
Z_{in}	Total input impedance in an optical preamplifier
Y_{in}	Total input admittance in an optical preamplifier
f	Frequency
I_g	FET gate leakage current
F	FET numerical factor equal to 4.1 for GaAs FETs
i_T	Total preamplifier noise current (with quantum noise)
i_A'	Preamplifier noise current density
i_Q	Quantum noise current
N	Averaging times
M	Number of samples per trace
Δ_q^2	Quantization noise variance

CHAPTER 1

INTRODUCTION1.1 Objective Of This Study

In this study the design and implementation of an Optical Time Domain Reflectometer (OTDR) at a wavelength of $1.3 \mu\text{m}$ for multimode fiber testing is presented. The OTDR is an instrument which is especially designed to detect the backscattered light from a fiber, when a light pulse is propagating down the fiber. The OTDR is used to locate the position of faults, discontinuities and breaks in a fiber-optic link, to measure the distributed link loss and to measure the link length. It is also possible to measure the loss these faults introduce.

Because the backscatter is very weak, all OTDR instruments available on the market introduce various signal enhancement techniques, such as filtering or boxcar averaging in order to improve their performance. The digital averaging technique seems to outperform the already existing signal enhancement techniques, since the obtainable gain from digital averaging may be more than 50 dB while the other techniques cannot have that gain.

At the beginning of this work there were no long wavelength ($\lambda=1.3 \mu\text{m}$) OTDR field instruments on the market. The objective of this work was the design and development of an OTDR system, for field use, operating at $1.3 \mu\text{m}$ wavelength. A long wavelength semiconductor

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laser is used for the OTDR light pulse source since all other available lasers, such as gas lasers, are bulky and require water cooling thus they are not suitable for field use. A new digital averaging technique is introduced which makes the OTDR system capable of unlimited averaging (the averaging is only limited by the analog to digital conversion noise which is a drawback common to all digital averaging techniques).

.. Basic operation of an OTDR

The OTDR was invented by Barnoski and Jensen [1] for testing optical fibers. It is similar in operation to the time domain reflectometer (TDR) that has been used for many years to locate and diagnose irregularities and mismatches in RF cables. In the optical case the inherent backscatter provides a visualization of the attenuation and scattering properties of the fiber as a function of the length. Not only can the attenuation of the fiber be estimated but anomalies can be located and characterized.

An OTDR, shown in Fig. 1.1(a), operates as follows. A fast rise-time, narrow-width, strong light pulse is injected into the fiber. As the pulse propagates down the fiber, an almost constant fraction [1] of the intensity is continuously reflected (or backscattered) due predominantly [1] to Rayleigh scattering [2]. The backscattered light is fed, via the optical coupler, to the optical receiver where it is detected and amplified. Then the signal at the optical receiver output may be displayed on an oscilloscope. The time variation of the transmitted power P_f and the received power P_b is shown in

Fig. 1.1 a) Basic block diagram of an OTDR.

- b) Time dependence between optical pulses, of peak power P_o , launched to the fiber and backscattered signal power P_b . T_{rep} is the repetition rate. Arrow (a) shows the exponential decay. Arrow (b) shows an arc fusion splice. Arrow (c) shows a discrete reflection from a fiber connector. Arrow (d) shows a fiber anomaly. Arrow (e) indicates the fiber far-end reflection.

tested with this experimental OTDR system has a length of approximately 20 km thus the two-way optical loss is 40 dB. For the 20 km link the light pulse repetition rate has to be 5 KHz (since the backscatter duration is 200 μ s) and the receiver-signal processing units should have an input dynamic range of 40 dB optical (80 dB electrical).

1.5 Thesis Organization

The physical origin of backscatter, its quantitative description and the principles of OTDR operation are discussed in Chapter 2. In Chapter 3 there is a detailed description of the basic experimental OTDR components: 1) the optical transmitter, 2) the optical coupler and 3) the optical receiver. Chapter 4 discusses the noise sources present in an OTDR. A noise (SNR) analysis of the receiver is presented and receiver noise measurements are given. Further SNR improvement techniques are also described and the implementation of a parallel digital averaging technique for improving the receiver output SNR is presented. The fifth chapter contains the experimental results and the final chapter has the conclusions and future work recommendations.

In the cutback method [2] the fiber is cut into several pieces and gives the attenuation in situ, thus it is a destructive measurement and requires access to both ends of the fiber. In the insertion loss method [1] the power transmitted through a long piece of fiber is first measured and then the power transmitted through a much shorter geometrically identical piece of fiber is measured. The difference gives the total attenuation of the long piece. Although this method is not destructive, it requires access to both ends of the fiber and is less accurate than the cutback method [1]. The apparent advantages of an OTDR are that it does non-destructive attenuation measurements and it does not require access to both ends of the fiber. The OTDR trace provides an immediate and easy visual representation of the distributed loss of the fiber requiring access to only one end of the fiber link while the cutback and insertion loss methods cannot do so.

As far as fault locating is concerned, there is a technique which investigates the fiber backscatter in the frequency domain. This technique can be implemented with an Optical Frequency Domain Reflectometer (OFDR). The OFDR [3] does not appear to be advantageous because it is equivalent to OTDR for reflective fault location [4] and it is less sensitive to attenuation faults. An OFDR is generally more complex than a OTDR and needs a stable light source. The OTDR is considered to be more efficient than the OFDR for fault locating and attenuation measurements.

Finally it can be concluded that the examination of the fiber with an OTDR provides information about the fiber attenuation, fault location and fiber length that no other currently existing technique

can provide by itself.

1.4 Design Requirements for the Experimental OTDR

At the wavelength of $1.3 \mu\text{m}$, the commercially available multimode fibers have an average attenuation of less than 1 dB/km while at shorter wavelengths, as shown in Fig. 1.2, the attenuation increases and the same fibers have much higher loss in the order of 3 to 4 dB/km . Because of the low loss, the $1.3 \mu\text{m}$ region is very suitable for long distance fiber-optic communication links. A typical allowable loss between two repeaters of a multimode fiber link is 30 dB [3] (this loss depends on the transmission rate) which for the $1.3 \mu\text{m}$ wavelength range is approximately equivalent to a repeater spacing of 30 km . The dynamic range of an OTDR, or in other words, the distance over which the OTDR should be able to detect the backscatter or nonreflective faults such as fusion splices, has to reach beyond the middle of the link to be able to probe the fiber from both ends.

Hence, the design goal for this OTDR is a dynamic range of at least 15 km or a two-way affordable optical loss of 30 dB (60 dB electrical), by launching to the fiber under test a light pulse of -2 dBm (0.65 mW) peak optical power, having a width of $1 \mu\text{s}$ (100 m resolution¹). For this purpose very low noise detection-amplification devices are employed and a digital averaging method is introduced for the backscatter signal enhancement. The fiber-optic link which is

¹ The delay per unit length of light in glass fibers is 5 ns/m thus the $1 \mu\text{s}$ transmitted pulse width gives a resolution of 100 m since its spatial width is 100 m .

Spectral attenuation of various fiber types

Fig. 1.2: Fiber attenuation as a function of wavelength for various types of fiber. [17, pp. 43]

tested with this experimental OTDR system has a length of approximately 20 km thus the two-way optical loss is 40 dB. For the 20 km link the light pulse repetition rate has to be 5 KHz (since the backscatter duration is 200 μ s) and the receiver-signal processing units should have an input dynamic range of 40 dB optical (80 dB electrical).

1.5 Thesis Organization

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CHAPTER 2

PRINCIPLES OF OTDR OPERATION2.1 Introduction

An OTDR is the instrument used to observe and measure the light backscattered from an optical fiber in the time domain when a fast rise time narrow light pulse is injected into the fiber. This gives the fiber pulse response in backscatter mode. The backscatter response is used to measure the fiber attenuation and other fiber parameters as well as to locate and characterize discontinuities and faults in the fiber. Light which is directly reflected from fiber discontinuities is also detected but the backscattered light from a section of fiber is much weaker than the light reflected from a discrete discontinuity which exists in the same fiber section.

The comparison between backscattered and reflected light power is first discussed. Then two approaches made by Personick [5] and Rourke [1], which derive mathematical expressions for the backscattered light power are presented. Finally the basic diagram of an OTDR is discussed.

2.2 Fresnel Reflection and Backscatter

There are two kinds of fiber backward wave response which may be applied to fiber attenuation measurement and fiber fault location. These are the Fresnel reflection and the backscatter.

The Fresnel reflection [2] is the reflection of light at the boundary between two optical media for normal incidence. The reflection coefficient is given by

$$R = \left(\frac{n_1 - n_0}{n_1 + n_0} \right)^2 \quad (2.1)$$

where n_1 and n_0 are the indices of reflection of the two media.

For a glass-air interface where n_1 (glass) = 1.48 and n_0 (air) = 1 then $R = 0.04$ or -14 dB thus the reflected power will be 14 dB below the incident power. This is an ideal case. But the Fresnel reflection which is caused by the refractive index difference between the fiber core and the outside medium when the fiber is broken, is not always ideally produced. Indeed there are conditions (i.e. the fiber break is dipped into index matched fluid) such that no reflection occurs from the fiber break. In practice the fiber may be broken diagonally or the breaks may lay in water (for glass-water interface $R = 0.036$) thus the reflected power will generally fall short of the maximum 14 dB value. It is also insufficient to depend on only the Fresnel reflection to locate faults since there are faults, such as fusion splices and local stresses, which are non-reflective.

However there is always intrinsic scattering in a fiber. When a pulse of light is coupled into the fiber it encounters absorption and scattering sites as it propagates. As shown in Fig. 2.1, if the light power at a distance z from a source located at a beginning of the fiber is expressed as $P_f(z)$, then inside an incremental length Δz of the fiber at z , one portion of the forward travelling light ΔP_a is absorbed, while ΔP_s is scattered and a part of it ΔP_b is backscattered. The scattering is due to the nature of the fiber

Fig. 2.1: Attenuation components in an incremental length Δz of fiber.

$P_f(z)$ is the incident light power at distance z into the fiber,

ΔP_s is the scattered component,

ΔP_b is the backscattered component and

ΔP_a is the absorbed component of power at z .

structure [1] such as leaky rays, bends, microbends, core/cladding boundary imperfections, very thin cladding or joints, to nonlinear scattering [2] and mainly to Rayleigh scattering [2]. The Rayleigh scattering is caused by variations in refractive index which are small compared to the light wavelength [2], occurs roughly uniformly along the fiber and is the dominant source of the scattering mechanism in a fiber. Rayleigh scattering decreases in proportion to the fourth power of the wavelength thus at short wavelengths (i.e. $\lambda=800$ nm) where it contributes approximately 90% of the total fiber loss [2] that loss is high, greater than 5 dB/km. At longer wavelengths ($\lambda=1.3$ μ m) the Rayleigh scattering decreases rapidly, and contributes only ~30% of the total fiber loss [2] while that loss now is less than 1 dB/km.

It was experimentally proved [6] that the backscatter is Rayleigh scattered light which is trapped in reverse-travelling guided modes and is directed back toward the source. Hence the backscatter response of a fiber is determined by fiber construction and does not depend on the fiber face conditions. Investigation of the backscatter can provide information on the fiber distributed loss and on the location and loss of the faults that may exist in a fiber link. The backscatter coefficient is defined as [2].

$$R_b = S \cdot a_g \cdot v \cdot W / 2 \quad (2.2)$$

where S is the fraction of the scattered light which is recaptured travelling back through the fiber toward the source given by [5].

$$S = \frac{(NA)^2}{4n_1^2} = \frac{n_1^2 - n_2^2}{4n_1^2} \quad (2.3)$$

NA is the numerical aperture of the fiber

n_2 is the fiber cladding refractive index equal to 1.46

n_1 is the fiber core refractive index equal to 1.48

a_s is the Rayleigh scattering loss coefficient in Neper/m

v is the light group velocity in the fiber equal to 2×10^8 m/s

and W is the injected light pulse width in sec

For typical values, $W=500$ ns and $a_s=5 \times 10^{-4}$ Neper/m then $R_b=1.25 \times 10^{-4}$ or -40 dB thus the backscattered signal will be 40 dB below the forward travelling signal level. The backscatter is 26 dB below the ideally produced Fresnel reflection coefficient.

2.3 The Backscatter Response to a Probe Pulse

This section outlines the analysis reported by Personick, [5] and Rourke [1] of the backscatter response of a fiber to a pulse of light.

Suppose that a light pulse with peak power P_0 is coupled into the fiber. The forward-travelling pulse suffers a position-dependent total attenuation determined by the total attenuation coefficient $a_{ft}(x)$. This coefficient consists of two terms, a scattering term and an absorption term. The scattering term $a_{fs}(x)$ in low-loss fibers is due predominantly to Rayleigh scattering. The absorption term $a_{fa}(x)$ is due to [1] intrinsic absorption, impurity absorption (due to metal ions as Fe, Cu, V, Cr, Co and hydroxyl OH ion) and defect absorption (related to the thermal history of the glass). The fiber total forward attenuation coefficient is the sum of the forward scattering and absorption coefficients,

$$a_{ft}(x) = a_{fs}(x) + a_{fa}(x) \quad (2.4)$$

The differential mode attenuation, due to lossy mode attenuation followed by mode mixing, is neglected because the loss introduced by it is small [2].

As the pulse is propagating down the fiber the change in transmitted power ΔP_f in an incremental distance Δz , as shown in Fig. 2.1, is proportional to the power $P_f(z)$ at point z , according to

$$\Delta P_f / \Delta z = -a_{ft}(z) P_f(z) \quad (2.5)$$

The minus sign indicates that the power decays as the pulse propagates.

In the limit as $\Delta z \rightarrow 0$ (2.5) can be rewritten as

$$P_f = P_0 \exp \left[- \int_0^z a_{ft}(x) dx \right] \quad (2.6)$$

where the pulse is travelling from 0 (source) to z .

The Rayleigh-scattered power from the fiber at z will be.

$$P_s(z) = a_{fs}(z) P_f(z) \quad (2.7)$$

Of this scattered light only a fraction S will be recaptured travelling back down the fiber toward the source (transmitter). This fraction S is given by (2.3). Thus the backscattered light at z will be

$$P'_b(z) = a_{fs}(z) S P_f(z) \quad (2.8)$$

The backscattered light while propagating toward the source, suffers a position-dependent total attenuation $a_{bt}(z)$. After a time $t = z/v$, a portion of the backscattered power $P'_b(z)$ will arrive back at the source and in an OTDR will be directed to a photodetector.

Consequently, the power backscattered at position z , that eventually finds its way to the detector, is

$$P_b(z) = P'_b(z) \exp \left[- \int_z^0 a_{bt}(x) dx \right] \quad (2.9)$$

or using (2.8) and (2.6) in (2.9) then

$$P_b(z) = P_0 a_{fs}(z) \int_0^z \exp(-\int_0^x [a_{ft}(x) + a_{bt}(x)] dx) \quad (2.10)$$

since $a_{bt}(x)$ and $a_{ft}(x)$ are both positive.

In (2.9) $P_b(z)$ is the power which is backscattered from z at the receiver input, while $P_b(z)$ is the power backscattered from z , at z .

If the fiber attenuation is the same for pulses travelling in both directions, the attenuation coefficients are identical:

$$a_{ft}(x) = a_{bt}(x) = a_t(x) \quad (2.11)$$

For all fibers of practical interest, it can be assumed that the fiber attenuation is independent of position, thus $a_t(x)$ can be replaced by a_t and finally

$$a_t = a_s + a_a \quad (2.12)$$

where all attenuation constants are in Nepers/m

Equation (2.12) can replace (2.4) and holds for both forward travelling and backscattered waves. Thus (2.10) can be rewritten as

$$P_b(z) = P_0 a_s \exp(-2a_t z) \quad (2.13)$$

The backscattered energy from the fiber section from z to $z+dz$ will be

$$dE_b(z) = P_0 a_s \exp(-2a_t z) dz \quad (2.14)$$

The backscatter response from the fiber section of infinitesimal length dz , has a duration

$$dt = \frac{2 dz}{v} \quad (2.15)$$

Thus (2.14) can be rewritten as

$$dE_b(z) = P_0 a_s \int \frac{v}{2} \exp(-2a_t z) dt \quad (2.16)$$

and the backscattered power from distance z is

$$P_b(t) = P_0 a_s S \frac{v}{2} e^{-2a_t z} \quad (2.17)$$

If the transmitted pulse in an actual measurement has width W , then the total backscattered power at time t or from a distance $z=vt/2$ is

$$P_{b\text{TOTAL}}(z) = P_0 a_s S W \frac{v}{2} e^{-2a_t z} \quad (2.18)$$

$$\text{or} \quad P_{b\text{TOTAL}}(t) = P_0 R_b e^{-2a_t z} \quad (2.19)$$

where R_b is given by (2.2)

Equations (2.19) or (2.18) assume that $e^{-a_t vt}$ is approximately constant for intervals of W duration. In order to have approximately constant backscattered power over the spatial width of the exciting pulse in the fiber, the pulse width must satisfy [5]

$$W \ll \frac{1}{a_t v} \quad (2.20)$$

2.4 The Backscatter Technique for the Measurements of Fiber Parameters and Fault Location.

The backscatter method can be used to make the following measurements on low-loss Rayleigh scattering-limited fibers [1]:

1. The total attenuation coefficient a_t ;
2. The scattering coefficient a_s ;
3. The fiber length;
4. The location of defects, faults, connectors, splices (etc.);
5. The location and type of faults occurring in fiber fabrication;
6. The loss of defects, faults, connectors splices, etc.;
7. The effects of the external environment on fiber properties (bends, microbends, lossy coatings, strain etc.).

Each of these fiber parameters can usually be obtained from a single OTDR characterization or "fiber signature". The natural logarithm of the backscattered power is directly proportional to the total attenuation coefficient α_t . From (2.18) it can be seen that

$$\alpha_t = \frac{1}{2(z_2 - z_1)} \ln [P_b(z_1)/P_b(z_2)] \quad (2.21)$$

so that the fiber attenuation coefficient α_t can be obtained from measurements of the backscattered power at two different points z_1 and z_2 along the fiber.

A second approach can also be used, whereby a perturbation in the fiber can be identified by observing the backscatter response. Consider the OTDR response of an imperfect fiber. Label the probe pulse as $A p(vt - z)$ and the unperturbed fiber response as $b(z)$ where A is a constant. Assuming that the resulting system remains linear when a small disturbance $d(z)$ is added, then the backscattered power $P_b(t)$ can be written as

$$P_b(t) = A \int_z [b(x) + d(x)] p(vt - x) dx \quad (2.22)$$

Equation (2.22) is the sum of two convolution operations.

$$P_b = A \cdot b * p + A \cdot d * p \quad (2.23)$$

where $*$ denotes convolution.

The first term of (2.23) represents the unperturbed response and the second term is the response due to the disturbance $d(z)$. There are three classes of perturbations of special interest as shown in Fig. 2.2.

a) A very narrow probe pulse with $p(z) = 0$ for z greater than some small distance ϵ , while $d(z)$, the disturbance is of greater extent.

Fig. 2.2: A comparison of the spatial length (width) W of the laser pulse with the perturbation width Δz as it propagates along the fiber. a) The case where $W = 2\epsilon \ll \Delta z$, b) The case where $W \gg 2\epsilon = \Delta z$, c) The case where $W = \Delta z$.

Then

$$\begin{aligned}
 A d^*p &= A \int_z d(x)p(vt-x)dx \\
 &= A \int_z d(x)\delta(vt-x)dx \\
 &= A d(vt),
 \end{aligned} \tag{2.24}$$

Equation (2.24) states that the backscattered power is a faithful reproduction of the perturbation response (scaled by the probe pulse amplitude) when the perturbation is much larger than the width of the probe pulse.

b) A point defect with $d(z)=0$ for z greater than ϵ and $p(z)$ of much greater extent (probe pulse much wider than the defect length). For this case,

$$\begin{aligned}
 A d^*p &= A \int_z d(x)p(vt-x)dx \\
 &= A \int_z \delta(x)p(vt-x)dx \\
 &= A p(vt)
 \end{aligned} \tag{2.25}$$

Equation (2.25) states that the backscattered power is a faithful reproduction of the probe pulse when the perturbation is much smaller than the probe pulse width.

c) The spatial extent of $p(z)$ is approximately equal to that of $d(z)$. In this case details of the shape of the perturbation become important. In general the response will not be a faithful reproduction of either the probe pulse or the perturbation but will be wider than each of them.

Hence a narrow pulse yields the impulse response of the perturbation (the probe pulse samples the perturbation over a distance 2ϵ which is much less than the perturbation width). On the other hand a point defect can be identified by its resemblance to the probe pulse.

The above analysis indicates the potential of this technique for studying defects and faults in fibers. Point defects can be located and then examined to determine the source of the defect. Since extended defects, effect bound mode power (or propagating modes) in different ways, examination of the OTDR signature provides clues about the origin of the defect.

In practice the pulse widths which are used in long range (>15 km), 1.3 μm light wavelength OTDR's are more than 500 ns. That pulse width covers a distance of 50 m. The actual length of a typical fault such as a break, a splice or a bend can be less than a centimeter so it is much less than the pulse spatial width. This situation is investigated in subsection b) of the previous analysis.

It can be concluded that, since the backscatter from faults is a faithful reproduction of the probe pulse when the faults are much smaller than the pulse width, the backscatter should contain frequency components similar to those of the exciting pulse. It is well known [7] that the frequency content of a perfect fiber (no faults) backscatter extends from DC to a few tens of kHz. Therefore for practical fibers the backscatter frequency components are determined by the exciting pulse components. For a T sec wide exciting pulse then the backscatter signal bandwidth is about $1/T$ Hz.

2.5 General Block Diagram of an OTDR System

A basic block diagram of an OTDR is shown in Fig. 2.3. The transmitter consists of a light source and the pulser which enables the source to send light pulses. The pulse width can be varied. The light pulses are directed through the optical coupler to the fiber link input. The backscattered light is directed through the coupler to the receiver where the optical signal is first detected, converted to an electrical signal and then amplified using low-noise techniques. Then the amplified signal is processed through signal processing units in order to maximize the SNR. The method most often used for SNR improvement is boxcar averaging. Finally, the OTDR trace is displayed. A perfect fiber (no faults) signature or backscatter response is shown in Fig. 2.4(a). The logarithm of the same fiber response, shown in Fig. 2.4(b), displays a linear relationship between light power along the fiber and distance. A fiber signature with typical faults (splices) is shown in Fig. 2.4(c) and the logarithm of this trace would appear as Fig. 2.4(d).

Fig. 2.3: General block diagram of an OTDR.

Fig. 2.4: Typical displays of an OTDR system.

- a) Backscattered signal from a perfect fiber.
- b) The logarithm of (a)
- c) Backscattered signal from a fiber link with four arc-fusion splices.
- d) The logarithm of (c).

CHAPTER 3

DESCRIPTION OF THE OTDR SYSTEM3.1 Introduction

An experimental OTDR system implementing the parallel digital processing technique was designed and built. The block diagram of the experimental OTDR is shown in Fig. 3.1.

A laser diode is used as the light source. The laser driver supplies pulses of current to the laser diode. The laser which is operating above threshold in the lasing mode sends light pulses through the optical coupler to the fiber. The light pulses have peak optical power directly proportional to the peak current pulse applied to the laser diode. The relationship between laser emitted optical power and laser driving current is given by the characteristic curve of the laser. This curve, for the experiment laser, will be given in section 3.2.

The optical power emitted by a laser is also a function of the laser diode junction temperature. Small changes in the junction temperature can cause significant changes in the emitted power. For this reason a cooler device, based on the Peltier effect [8], is used to keep the laser diode temperature constant and low ($\sim 20^{\circ}\text{C}$).

The optical bidirectional coupler is a fused biconical-taper coupler [9]. This is a 4-port, 3 dB coupler, with the fourth port unused. The fourth-port fiber is dipped into index-matching oil (oil

Fig. 3.1: The block diagram of the experimental OTDR.

that has the same refractive index as the fiber core index). The laser light is directed through the coupler, with some loss, to the fiber link. The backscattered light is then directed, with an additional loss, from the fiber to the detector-receiver.

The receiver consists of a detector-preamplifier-amplifier circuit which transforms the backscattered light to an electrical signal and then amplifies it. The backscattered power is weak and the backscatter signal to noise ratio (SNR) is very low, typically reaching the -55 dB level at the end of a 20 km fiber trace. Thus the noise performance of the receiver is critical.

The optional logarithmic amplifier which follows the receiver takes the logarithm of the backscatter and transforms it from an exponential function to a linear function.

The digital averager is a signal processing unit which is used to average the receiver output in order to improve the SNR. This unit samples the backscattered signal, digitizes it and numerically averages it. The averaging may be as long as practically useful but the SNR improvement is eventually limited by the A/D converter quantization noise contribution. The design of the transmitter, optical coupler and receiver are discussed in detail in the following sections.

3.2 The Transmitter Section

The OTDR transmitter section has three main components: 1) the optical source; 2) the source driver circuit; and 3) a circuit for

temperature stabilization of the optical source.

3.2.1 Selection of Light Source

For OTDR applications, it is desirable to have optical sources capable of producing high peak radiance and moderate repetition rates with pulse durations in the range of 80 ns up to a few μ s because the pulse width is always limited as stated by (2.20). For the commercially available fibers, with attenuation of about 1 dB/km at 1.3 μ m, (2.20) states that the pulse width should be much less than 25 μ s.

In the spectral region around 1.3 μ m wavelength, a suitable pulsed source for a field use OTDR is the single heterojunction (In,Ga)As laser diode. The lasers are threshold devices, as shown in Fig. 3.2. Below threshold, the laser emits light almost linearly proportional to the current through the device, like an LED. Above threshold the emission of a laser diode is coherent and generated at high differential quantum efficiency dP_0/dI where P_0 is the emitted light power and I is the drive current.

The coupling efficiency of a laser to the fiber can approach 90% to 100%, by using various lens schemes, because the laser beam is directional.

Lasers do not maintain a constant optical output if the device temperature (the junction temperature) is changing. The threshold current of the laser changes by approximately $+1\%/^{\circ}\text{C}$. This 1% change

Fig. 3.2: The characteristic curve of a laser diode. I_{th} is the laser threshold current. Above I_{th} the light emission is coherent and generated at high differential quantum efficiency dP_o/dI .

is disastrous in terms of light output stability because its effect is magnified by the high differential quantum efficiency beyond threshold. For example a change of only 30°C can result in a complete loss of lasing output for threshold levels below a specific value. Clearly temperature stability must be considered in the design of the transmitter.

The laser diode light source used in the present experiment is a BNR laser diode mounted in a package with a fiber pigtail. In the same package there is a thermistor, a Peltier-effect junction as the cooling device, and a p-i-n diode facing the back mirror of the laser diode (the light emitted from the front mirror is coupled to the fiber). The pin-out diagram of this laser-package is shown in Fig. 3.3. The measured laser characteristic is shown in Fig. 3.4. The threshold current is 110 mA. the optical emitted power increases proportionally with current up to 450 mA of drive current, then saturates. In pulsed operation (as for OTDR), the laser linearity throughout the characteristic is not as important as the cooled peak power.

3.2.2 Laser Diode Driver-Pulser Circuit

For the OTDR a current in the range of 100 to 500 mA must be switched on and off at high speed through the laser in response to a low-level input TTL signal which is used to trigger the driver. The traditional circuit for controlling such currents is a bipolar transistor switch operated in the common emitter (CE) configuration as shown in Fig. 3.5(a). The speed of the CE switch is limited, however, compared with the speed of other configurations to be discussed. Bandwidth (BW) is traded off for current gain although this can be

<u>PIN NUMBER</u>	<u>FUNCTION</u>
1	Cooler +
2	Case Ground
3	Thermistor
4	Thermistor
5	Photodiode + bias
6	Photodiode - bias
7	Laser -
8	Laser +
9	No Connection
10	Cooler -

Fig. 3.3: Pin-out diagram of the packaged BNR laser diode.

Fig. 3.4: The P_o vs I characteristic of the BNR laser diode for pulsed operation. The threshold current is about 110 mA. The laser will degrade significantly and burn out if driven above 500 mA.

(a)

(b)

Fig. 3.5: Examples of laser driver circuits. a) A common emitter saturated switch. b) A differential amplifier current-routing circuit.

compensated somewhat by preemphasizing (overdriving) the base current during the turn-on period. In Fig. 3.5(a) preemphasis is accomplished by using a speed-up capacitor C_1 .

Speed can be limited by the fact that the laser diode drive appears to come from a current source (the reverse biased collector-base junction). Thus if the diode has a high capacitance, the laser drive current is initially diverted into charging this capacitance, thereby reducing the current available for light generation. This can be offset by a speed-up capacitor or by prebiasing the laser diode near its cut-in voltage. Alternatively, one must go to a low-impedance driver such as emitter-follower circuit to provide a voltage step rather than a current step. Speed can also be limited by the time required to remove minority carrier charge stored principally at the collector-base junction during saturation. This effect can be minimized by using a Schottky diode clamp between the collector and base of the drive transistor. Negative feedback thus provided limits the extent of saturation to about 0.3 V while the diode itself adds no additional minority carrier stored charge.

An important variation of the collector-loaded CE switch is the emitter-coupled or current-routing circuit shown in Fig. 3.5(b) [10] and [11]. Although the circuit resembles a linear differential amplifier it is operated outside the narrow range of linearity in the switching mode of operation. The usefulness of the current-routing circuit as a high speed circuit is based on nonlinear operation whereby one of the two transistors is cutoff in the steady-state condition. The transistor can be strongly overdriven but before it reaches

saturation, the feedback impedance (the impedance at the interface between the two stages is very low) is changed from that of the forward biased diode resistance (or the common-base input impedance) to the current source impedance which is very high. In other words, the emitter current is limited to the current I_0 supplied by the current source. The collector supply voltage has to be sufficiently positive and the load resistance sufficiently small to avoid causing saturation when the base is at its most positive potential. The circuit functions as if it were strongly overdriven at the input while at the same time remaining out of saturation. This effective configuration results in switching speeds faster than the CE configuration.

The current source should be built with a fast transistor having a low output capacitance C_{ob} so that the current supplied is not affected by the high speed switching operation via collector-to-base coupling.

The driver used in the experimental OTDR set-up is of the current-routing circuit type. Instead of using a reference voltage at the input of transistor Q_2 the base of Q_2 is driven with $\overline{V_{in}}$ (the complement of V_{in}). The driver circuit is shown in Fig. 3.6. It is capable of delivering 40 to 500 mA. At 500 mA the minimum achievable pulse width is 170 ns. The transistors Q_1, Q_2, Q_3 are type 2N2219A having an f_T of 300 MHz. In operation, the collector-emitter (V_{ce}) voltage of Q_3 is traded off for minimum possible power dissipation. The value of V_{ce} of Q_3 is kept slightly above the saturation level, while the power dissipated in the transistor remains within safe limits.

Fig. 3.6: Circuit schematic for the laser driver of the experimental OTDR system.

The driver current range (40 to 500 mA) satisfies the BNR laser driving requirements because the BNR laser can be operated at currents from 100 mA to 400 mA, for pulsed operation. The BNR laser was successfully tested for continuous operation at 300 mA, thus it can be safely operated at 400 mA pulsed.

3.2.3 Temperature Stabilization of the Laser Diode

A small temperature increase significantly reduces the light power emitted from a laser diode when lasing. Thus a means of cooling the laser junction has to be adopted. An effective cooling circuit uses a Peltier-effect junction [8]. In the Peltier effect, if a direct current is passed through a circuit formed from two dissimilar metals or from a metal and a semiconductor, one junction gives off heat and is cooled and the other absorbs heat and becomes warm. The effect is reversible. Larger temperature differences are produced with metal-semiconductor junctions than with metal-metal junctions. A number of such junctions can be used to form a Peltier element which may be used as a cooling or heating element.

In the laser package a Peltier element is mounted in such a way that the cold side is in contact with the laser diode. A thermistor is also placed in close proximity to the laser diode in order to follow with accuracy the changes of its temperature. The thermistor is a resistor, usually fabricated from semiconductor material, that has a large negative temperature coefficient of resistance (the higher the temperature the less the thermistor resistance). The thermistor resistance to temperature characteristic is shown in Fig. 3.7.

Fig. 3.7: The thermistor resistance vs. temperature characteristic.

A schematic of the cooler circuit [3] is shown in Fig. 3.8. Circuit operation can be explained as follows. A change in temperature alters the voltage across the thermistor represented by resistance R_{TH} . This forces the differential amplifier U_2 to supply (or draw) current to (or from) the base of the output transistor Q_2 . The increase or decrease of the base current increases or decreases respectively the collector current of Q_2 . Transistor Q_1 is used as a safety valve for the protection of Q_2 which supplies the current to the cooler. When the collector current of Q_2 is too large, Q_1 turns on and allows the op-amp output current to flow through it, thus reducing Q_2 base current and finally limiting the cooler current to a maximum allowable value. An increase in current through the cooler forces the temperature to drop. The cooler current can be adjusted by the 5 k Ω potentiometer. This sets the reference voltage applied at the inverted input of the operational amplifier (pin 6). The difference between the thermistor voltage (pin 5) and the reference voltage (pin 6) determines the amount and direction of the base current into Q_2 . So any variation in the thermistor resistance (due to a temperature change) results in a change in cooler current. Thus the laser diode temperature remains constant. The cooler is essentially a feedback-control circuit where air is the feedback path.

Fig. 3.8: The Cooler circuit of the experimental OTDR.

3.3 Optical Directional Coupler

The optical coupler used in the experimental OTDR instrument is a biconical-taper coupler [9]. It consists of two fibers that have been twisted, heated and pulled into a taper and then fused in position as shown in Fig. 3.9. The coupling ratio of this type of coupler is adjusted by narrowing the waist of the biconical taper until the desired coupling ratio is obtained. The device passes that portion of light corresponding to low-order-modes directly from the input port to a single output port while it divides the rest of the light corresponding to high-order-modes, approximately equally between the two output ports (low-order modes travel near the fiber core axis while high-order modes penetrate more deeply into the cladding and are generally more lossy). To reduce the modal sensitivity derived from the separation of the light into higher- and lower-order modes portions, the coupler should be constructed in a way such that all the guided light takes part in the coupling action equally. This can be accomplished by reducing the waist of the biconical taper until almost all the light has converted to cladding modes (modes which travel in the cladding, mostly higher-order modes). Thus the input light converts via the down-taper into higher-order modes that are guided and mixed in the fused twisted region. The up-taper then gathers these modes together with the high-order modes predominating in the tapping fiber. The waist region acts as a mixing region in which the memory of the modal history of the light is in principle, lost. This coupling structure naturally leads to approximately modal-insensitive coupling. If two identical fibers are used, the coupler gives an approximately 3 dB optical power split.

Fig. 3.9: a) An illustration of the technique used to build an optical bidirectional 4-port coupler.
b) The coupler schematic.

The connections made to the optical coupler for use in the OTDR instrument are shown in Fig. 3.10. Port 1 is spliced to the laser pigtail, port 2 is spliced to the fiber cable and port 3 is spliced to the detector pigtail. Port 4 is terminated in an index matching fluid. (i.e. oil which has a refraction index approximately equal to the core refractive index of $n_1=1.48$).

The launched light at port 1 is transmitted to port 2 and 4. The coupling ratio between ports 1 and 4, or between ports 1 and 2, for an ideal case is 3 dB. Similarly the coupling ratios between ports 2 and 1 and between port 2 and 3 are 3 dB (ideal). Ideally port 2 should be isolated from port 4, and port 1 should be isolated from port 3. The coupler performance in terms of total input power (laser light launched at port 1) and total useful output power (backscattered power detected at port 3) is generally specified in terms of an insertion loss which in the ideal case is 6 dB (3 dB for 1 to 2 transfer plus 3 dB for 2 to 3 transfer).

The coupler used in the experiment, was manufactured by Canstar, model TC3-C and has a two-way total loss, with the splices loss included, equal to 8.6 dB. The coupling ratios, from ports 1 to 2, and 1 to 4 are not equal. The 2 to 3 and 2 to 1 coupling ratios are not equal as well. The coupling ratio between ports 1' and 2' (laser to fiber cable) was measured to be 3.6 dB, while the coupling ratio between ports 2' and 3' (backscatter power from fiber cable to detector) was 5 dB. Thus the total loss introduced by the coupler is 8.6 dB. The isolation between ports 1 and 3 as well as between ports 2 and 4 is more than 42 dB. The above experimental coupler appears to be 2.6 dB

Coupling ratios of the CANSTAR coupler:

from 1' to 2' is 3.6 dB

from 2' to 3' is 5 dB

Fig. 3.10: The connections of the optical bidirectional coupler to the OTDR system.

more lossy than the ideal coupler. This additional loss is probably due to the light which escapes from the cladding along the waist region of the coupler, where almost all light has converted to cladding modes.

If port 4 is not terminated then a portion of the light travelling from port 1 to port 4 is reflected. The magnitude of the reflection depends on the quality of the fiber end break. For a perfect break, as pointed out in section 2.2, the reflected power is well above the backscatter power level. Since it appears as a pulse at the front of the backscatter trace it is called front-end reflection. Its peak (always depending on the quality of the break) is much higher than the front fiber region backscattered power, consuming useful receiver dynamic range or even saturating the receiver. This unwanted front-end reflection can be reduced or eliminated if the fiber at port 4 is dipped into index matching fluid. In the experimental OTDR the front-end reflection is suppressed below measurable level.

3.4 The Receiver Section

One of the most important parts of an OTDR system is the optical receiver, which detects the backscatter light from the fiber cable and processes the resulting electrical signal. The optical receiver can be considered as an optical to electrical converter or O/E transducer.

The receiver, to be described in this section, consists of a photodiode and an associated amplifier along with necessary filtering. The function of the photodetector is to detect the incident light signal and convert it into an electrical current. The amplifier

amplifies this current while introducing the minimum amount of additional noise to corrupt the signal.

The fundamental goal in the design of an optical receiver is to minimize the amount of optical power which must reach the receiver in order to achieve a given signal to noise ratio (SNR). This minimum optical power is commonly referred to as the sensitivity. Other considerations also influence the details of the design, and in fact may result in a practical receiver with less than the optimum sensitivity. For example the sensitivity may be relaxed in order to achieve a wide dynamic range.

For most optical communication applications the optical detection element is either a p-i-n photodiode or an avalanche photodiode. The receiver sensitivity is determined by the individual properties of both the detector and the circuit in which it is located. In all receivers for which maximum sensitivity is required, the conversion of the optical signal into an electrical one of sufficient amplitude proceeds via a two- or three-step process as shown in Fig. 3.11.

Firstly, the incoming optical signal is converted into an electrical signal in the photodetector by a photoelectric effect.

Secondly, this signal may or may not be further amplified within the detecting element depending on the type of detector. If internal gain is present, it is usually accompanied by the generation of some additional noise. As we shall see later in connection with avalanche

Fig. 3.11: Generalized photodetection - amplification process at the optical receiver indicating all noise sources introduced during the process.

photodiodes, this gain mechanism can in fact be quite noisy, so much so that there are several instances where little or nothing is to be gained by way of improved SNR when an avalanche photodiode (APD) is used [12],[13],[14],[15].

Thirdly, the photocurrent generated in the photodiode will undergo amplification, usually in a FET or bipolar front-end amplifier, where again additional noise will be added, and the signal and noise will both be filtered or processed so as to maximize the SNR.

In order to convert usefully as much of the incoming light energy or power into electrical current, the photoelectric effect should be as efficient as possible. In the following sub-section, the nature of the photoelectric effect will be briefly discussed, along with other photodiode parameters which determine the receiver performance. Then the design of a front-end circuit (preamplifier) will be presented. Finally our experimental preamplifier will be presented and its parameters will be given.

3.4.1 Optical Detection. Selection of Optical Detector

Two important factors determine the efficiency of the optical to electrical conversion during the photoelectric effect:

- 1) The quantum efficiency η of the photodetector, which is defined as that fraction of the incident photon flux, that generates electron-hole pairs, which are collected by the detector contacts or in other words η is the number of carriers (electrons) per incident

photon.

2) The responsivity R of the photodiode is the ratio of the photocurrent I_{ph} to the incident optical power P_o .

Since each photon has energy $h\nu$, the responsivity R can be related to the quantum efficiency η by [16].

$$R = \frac{I_{ph}}{P_o} \quad (3.1)$$

$$= \frac{N_e q}{t} \bigg/ \frac{N_{ph} h\nu}{t}$$

$$= \frac{\eta \lambda}{1.24} \quad \text{A/W if } \lambda \text{ is in } \mu\text{m} \quad (3.2)$$

where N_e is the number of generated electrons

N_{ph} is the number of incident photons

q is the electron charge

Under normal bias, a photodiode [17] (reverse-biased p-n or n-p junction) has a depleted semiconductor region with a high electric field that serves to separate photogenerated electron-hole pairs. Each photon which is absorbed within this region produces an electron-hole pair that is quickly separated by the field and the two charge carriers move to the respective contact regions where they recombine. Their movement results in current impulses with a time duration equal to the transit time of the carriers from the generation point to the edge of

the depletion layer. The life-time of the charge carriers determines the quantum efficiency of the detector. Short lifetime materials (direct gap materials, defined as having a lifetime shorter than the time required to diffuse to the edge of depletion layer) recombine before they reach the depletion layer edges, thus do not contribute to conduction while long lifetime carriers (indirect gap materials such as Silicon and Germanium) are able to move to the edge of depletion layer where they recombine thus contribute to conduction. The lifetime of the carriers depends on the depletion layer region where the photon is absorbed while the absorption process is a function of wavelength and material.

The thickness of the active region is a trade-off between the competing effects of fast transit time, requiring a narrow depletion region, and the combination of quantum efficiency and low capacitance, which require a wide depletion region.

The p-i-n photodiode is one of the most common photodetectors because the depletion layer thickness (the intrinsic -i- "high resistivity" layer between p^+ and n^+) can be tailored to optimize both the quantum efficiency and the frequency response.

Avalanche photodiodes (APD) contain a region of high electric field, where avalanche multiplication of the photogenerated carriers occurs. The current delivered to the diode contacts is many times larger than the primary current generated by the incident light, because the multiplication gives rise to internal current gain. The APD requires much higher bias voltage than a p-i-n, to maintain the

high electric field. The current gain is not a linear function of the applied voltage and is sensitive to temperature.

When there is no incident light, the reverse currents which arise in the device (p-n, p-i-n or APD) are: 1) generation recombination currents in the depletion region, and 2) surface leakage currents. The reverse currents are called dark current, I_D . In APDs a part of the dark current is multiplied by the internal gain. The total dark current of the p-i-n photodiode introduces shot noise having a mean square amplitude $\langle i^2 \rangle_{pin} = 2qI_D B$ where B is the receiver bandwidth. The shot noise an APD introduces is $\langle i^2 \rangle_{APD} = 2q[I_{DM} \langle M^2 \rangle B + I_{DN} B]$ [11] where I_{DM} is the multiplied part of the dark current, I_{DN} is the non-multiplied part and $\langle M^2 \rangle$ is the mean square value of the carrier multiplication factor M . Also $I_D = MI_{DM} + I_{DN}$. The dark current is dependent upon temperature and rises as temperature increases. Low dark current values require a small depletion layer area and minimum depletion volume.

Photodiodes, either p-n, p-i-n or APDs are nearly ideal detectors for the 0.8-0.9 μm range when made from Silicon, as evident in Fig. 3.12(a) [16]. They have nearly 100% quantum efficiency, can be made with low capacitance and have adequate speed for nearly all applications.

For longer wavelengths as shown in Fig. 3.12(b) quantum efficiency drops very low for Si detectors. Germanium diodes can be used for detection at long wavelengths ($\lambda > 1 \mu m$) with quantum efficiencies up to 70%. Due to smaller bandgap Germanium diodes have

(a)

(b)

Fig. 3.12: a) Quantum efficiency and responsivity for various photodetectors [16, pp. 751].
 b) Quantum efficiency of longer wavelength photodetectors (Ge and III-V compounds such as GaInAs) [17, pp. 4-21].

substantially higher dark current per unit volume than silicon detectors.

Finally, semiconductor materials formed from elements in groups III and V of the periodic table have been fabricated into photodiodes (p-i-n diodes and APD). These compounds have such absorption and ionization properties [11] that allow small device structures, fast carrier transit times and high quantum efficiency. Dark currents in these materials are smaller than Ge but larger than Si.

For longer wavelength systems, 1.2-1.4 μm , a comparison between APDs and p-i-n's and the various material options such as Ge or III-V compounds, will be crucially dependent on the dark current as well as on the preamplifier stage noise contribution [12]. Germanium, however, has a high bulk leakage (dark) current. A low capacitance and leakage current could be obtained more successfully from a III-V device [14].

The low-noise advantages of a p-i-n diode in addition to the economical and operational disadvantages of an APD when operated in a portable OTDR (such as operation voltages greater than 25 V and additional circuitry for the control of avalanche gain) led us to the selection of a GaInAs p-i-n photodiode (a III-V compound semiconductor material) interfaced with a GaAs MESFET preamplifier circuit. This p-i-n FET package is manufactured by Lasertron, model Q-DF-090-T. The p-i-n photodiode has a responsivity of 0.87, a capacitance of 0.5 pF and a dark current less than 15 nA, and can be operated from -5 V power supply. The preamplifier circuit specifications will be discussed in section 3.4.3.

3.4.2 The Preamplifier Design in General

The fundamental goal in the design of an optical receiver is to minimize the amount of optical power which must reach the receiver in order to achieve a given SNR or for a given amount of optical received power to achieve the maximum SNR. The minimum detected optical power in order to achieve a given tolerable SNR is commonly referred to as the sensitivity and is usually measured in dBm of optical power. It depends upon the detector type and characteristics as well as the design of the preamplifier. The preamplifier or front-end is defined as the first stage of amplification following the photodetector. It will be the dominant source of noise added to the signal and hence its design will be the principal factor in determining the sensitivity of the receiver.

The most straight-forward [11] but not necessarily the optimum method of designing a front-end is to terminate the input to the preamplifier with a load resistor R_L , as shown in Fig. 3.13(a), such that in conjunction with the input total capacitance C_T , the bandwidth of the input admittance is equal or greater than the detected signal bandwidth B . This requires

$$R_L \leq \frac{1}{2\pi B C_T} \quad \text{where } C_T = C_d + C_s + C_a \quad (3.10)$$

C_d is the diode capacitance

C_s is stray capacitance

and C_a is the preamplifier input capacitance

(a)

(b)

(c)

Fig. 3.13: Examples of preamplifiers (front-ends) for an optical receiver. a) The straight forward approach. b) The integrating, hi-impedance front-end. c) The transimpedance front-end.

High speed detection requires small values of R_L , while low noise performance requires large R_L . If R_L is 50 Ω it will introduce Johnson noise many times higher than that of an optimum bipolar or FET front-end. Thus the simplicity of this approach is paid for by a considerable increase in the circuit noise over that potentially achievable.

The second approach [11] to the design of a front-end, is to reduce all sources of noise to the absolute minimum. This is accomplished by selecting low-capacitance high-frequency devices, low dark current detectors and minimizing the bias resistor thermal noise by making R_L large. This approach is referred to as the high-impedance front-end shown in Fig. 3.13 (b). When the input resistance R_{in} is large then the input admittance is dominated by the capacitance C_T and the signal tends to be integrated [11]. There are two drawbacks in this approach. 1) An equalizer is needed to compensate for the initial $R_{in}C_T$ roll-off and since R_{in} and C_T varies from unit to unit individually equalized amplifiers are required. 2) This design has a reduced dynamic range compared to other approaches [11], such as the transimpedance pre-amplifier. However the high-impedance front-end can achieve the ultimate in low-noise front-ends but with reduced dynamic range [11][18].

The third approach involves using a transimpedance preamplifier where the load resistance R_L becomes the feedback resistance of the transimpedance amplifier. Shown in Fig. 3.13(c) it is basically a current-to-voltage converter [19]. In the limit of a large loop gain the relation between output voltage and input current is given by

$$V_{out} = -Z_F \cdot I_s \quad (3.11)$$

where Z_F is the effective feedback impedance

The transimpedance preamplifier is capable of wide bandwidth [19], provides greater dynamic range than the high impedance approach and has a noise performance which can approach that of the high-impedance design [11][18]. The "apparent" load resistance (the closed-loop input resistance) because of high loop gain, is low ($Z_{in} = Z_F / 1 - K$ where K is the gain, from the Miller theorem) thus minimizing the input time constant of the circuit and resulting in an extended frequency response. The open-loop input impedance (the amplifier input impedance) should be high so that essentially the photocurrent I_s flows through R_F and the output signal is defined by $I_s \cdot R_F$ since the input node is an AC ground [19]. The transfer gain of this configuration is then R_F volts/amp at all frequencies for which the forward loop gain remains $\gg 1$ [19].

The basic sources of noise for the transimpedance pre-amplifier are detector leakage, the input transistor and the feedback resistor R_F . Since the maximum value of R_F is limited in order to keep the amplifier stable [19][11] then the noise of the transimpedance preamplifier will always exceed that of a high impedance preamplifier. A capacitance C_F which is much less than C_T [17] may be added externally or be considered as the feedback resistor stray capacitance. This capacitance is necessary for stability and its absence may yield a small damping constant producing unwanted oscillations.

From the three preamplifier designs, which are discussed above, the transimpedance design offers wider BW, high gain, wider dynamic range and less need for equalization compared to the integrating front-end design, and the straightforward design.

However, its noise contribution exceeds the integrating front-end noise.

3.4.3 The Experimental Preamplifier

The detector-preamplifier of the experiment OTDR is a Lasertron pin-FET receiver module, (pin-out diagram shown in Fig. 3.14(a)) with a high performance, low capacitance, low leakage 1.0 to 1.6 μm GaInAs pin detector, hybridized with a 1 μm gate GaAs MESFET transistor transimpedance preamplifier. The module input is a 50 μm core graded index fiber coupled to the pin detector. The transimpedance preamplifier provides optical dynamic range of more than 25 dB. Its sensitivity for a 10 MHz BW signal is -50 dBm.

The GaInAs pin detector has a responsivity of 0.85 A/W at 1.3 μm and a capacitance of 0.45 pF at a bias of -5V. The pin diode dark current is specified to be 10 nA at 25°C. The detector dark current increases with temperature causing the sensitivity to decrease due to higher shot noise.

The GaAs MESFET (a FET with a Schottky barrier gate electrode) preamplifier is a transimpedance cascode feedback circuit, shown in Fig. 3.14(b), which can achieve 100 MHz bandwidth. The use of a GaAs

Fig. 3.14: a) The Lasertron transimpedance preamplifier pin-out diagram.
b) The Lasertron transimpedance preamplifier circuit diagram and the FET model at high frequencies.
c) The 4 MHz low-pass filter.

MESFET in the first stage of the preamplifier offers excellent noise performance. These devices have a high transconductance g_m and a small gate capacitance C_g where

$$C_g = C_{gs} + C_{gd}$$

where

C_{gs} is the gate to source FET capacitance

and C_{gd} is the gate to drain FET capacitance

As we will see in the next chapter in order to minimize the amplifier noise, g_m/C_T^2 should be maximized and MESFETs can fulfill this requirement. Capacitance C_T includes the photodiode capacitance, the preamplifier input capacitance C_a and stray capacitances C_s . Capacitance C_a consists of the gate capacitance C_g and the Miller capacitance. Due to capacitance C_{gd} there is a feedback path from output to input. Using the Miller theorem, the contribution can be expressed in terms of voltage gain A_v and C_{gd} , C_{gs} capacitances as follows [24].

$$C_a = C_{gs} + (1 - A_v)C_{gd} \quad (3.12)$$

The common source FET amplifier has inverting gain thus the input capacitance C_a is significantly increased due to the Miller capacitance term $(1 - A_v) C_{gd}$. Minimization of C_a requires a circuit configuration chosen to minimize the Miller capacitance. This can be achieved by following the common source front-end stage with a stage having a low input impedance such as the common-base configuration. The common source has a voltage gain

$$A_v = g_m R_D$$

where R_D is the drain resistance

Where a common-base stage follows, the effective drain resistance is R_D in parallel with the common-base input resistance which is very low, thus the term $(1-A_v) C_{gd}$ is much reduced. The cascode configuration yields very low input capacitance C_a . The following common-base configuration offers high voltage gain (approximately equal to the common emitter, non-inverting gain) thus producing a high open-loop gain for the transimpedance front-end.

The output stage of the front-end is an emitter follower circuit which serves also to match the high common-base output impedance to a low impedance, since the transimpedance preamplifier should have a low output impedance (50Ω). The feedback path is from the common emitter output to the common source FET front stage gate. The feedback resistor doesn't load the preamplifier output because of low output impedance. The feedback resistor R_f is $40 K\Omega$ and the front-end 3 dB bandwidth is 90 MHz, as given in the data sheets.

The backscattered signal bandwidth, for a $1 \mu s$ launched laser pulse, is 1 MHz. Since shorter light pulses, down to 200 ns, will also be launched for better spatial resolution (20 meters), the chosen receiver bandwidth is 4 MHz. A low pass filter (LPF) shown in Fig. 3.14(c) has to follow the Lasertron preamplifier module in order to limit the front-end bandwidth from 90 MHz to 4 MHz. The Lasertron transimpedance transfer gain is equal to R_f or $40 K\Omega$ over the 4 MHz bandwidth, with a ± 1 dB fluctuation as reported in the Lasertron data sheets and as verified experimentally later.

3.4.4 The Post Amplifier Stages

Since the backscatter is very weak, in the range of -40 dBm to -90 dBm, the amplification introduced by the transimpedance front-end is inadequate. Two additional voltage gain stages are used. Each stage is a video amplifier circuit using the Motorola IC NE592. The circuit is shown in Fig. 3.15. The first stage has a voltage gain of at least 200 as given in the data sheets for the NE592 while the second stage has gain variable from 10 to 200. Resistor R sets the gain which needs to be variable to avoid saturation of the second stage when larger than -50 dBm optical backscatter signal levels are amplified. The second stage maximum peak-to-peak output voltage is less than 2 volts. In order to have an output signal of 8 V p-p amplitude, two transistor amplifier stages follow with variable gain. Finally an emitter follower circuit buffers the post amplifier output from the next stages of the receiver and drives them through a low impedance source.

3.4.5 The Range Switching Circuit

The backscattered signal from a 20 km fiber link with a 1 dB/km fiber attenuation has an optical dynamic range of 40 dB. In other words the backscattered signal from a distance of 0 km will be 10^4 times larger, in terms of optical power or in terms of signal current or voltage, than the backscattered signal from a distance of 20 km. This is because the 0 km backscatter suffers 0 dB attenuation, while the 20 km backscatter suffers, because of its 40 km round trip, 40 dB attenuation.

Fig. 3.15: The post amplifier circuit schematic.

Since digital averaging will be introduced to improve the SNR of the backscatter, an A/D converter is used to digitize that signal. The input dynamic range will be determined by the A/D converter word length. For an 8-bit A/D, with 256 discrete levels of quantization, the ratio of the maximum possible input voltage to the minimum one determines the A/D input dynamic range, which in terms of voltage ratios (or optical power ratios) is 24 dB. For a 20 km link the backscatter signal range is 40 dB, so that the signal has to be divided into two ranges covering 20 dB each. The first range (0 to 10 km) can be digitized directly but the second range (10 to 20 km) has to be amplified in order to use the A/D input range effectively. That is, we want to set the 10 km backscatter signal to the maximum available A/D input voltage, so that the 20 km backscatter signal, which is 20 dB below the 10 km signal, can have a value at least twice the A/D quantization step. Now the two 20 dB ranges can be efficiently digitized by the 24 dB input dynamic range A/D.

The switching circuit shown in Fig. 3.16 consists of two high speed analog MOSFET switches, two high speed switch drivers and some digital circuitry to decide when to switch on or off the first or the second part of the backscatter waveform. For optimum operation and to acquire the fastest available switching times, the analog MOSFET DG308 switches, made by Siliconix, have to be driven with MOSFET logical levels (High=+15V, Low=0V). They need ± 15 power supply operation. The analog switch series resistance is low, about 100 Ω . Under these conditions the switch "on" delay is 150 ns from the time the pulse is applied to the digital input of the switch while the switch "off" delay is 200 ns. These analog switches are made to have switch "off" times

Fig. 3.16: The range switching circuit schematic.

longer than switch "on" times and they are termed as "break before make" switches.

The drivers for the analog switches are common emitter transistor circuits. The transistors which are used are 2N2369A NPN high speed saturated switches with an f_T of 500 MHz and collector currents up to 100 mA. The levels applied at the inputs of the drivers are TTL (0 to 5V), while the output levels (the collector voltage) of the drivers are MOS levels (0 to 15 V). These levels are applied then to the digital inputs of the analog switches.

The digital circuitry consists of a shift register, 74LS96, a few inverter gates, 74LS04, and decade counters 74LS192 which divide the C_1 clock by 250. All clock circuitry is drawn and explained in the digital hardware of the averager section 4.4. Then the counters "borrow" pulse strobes the shift register to shift. The shift register is operated in a closed loop mode exchanging the positions of "1" and "0" at Q_0 and Q_1 when it is triggered with the inverted "borrow" counter pulse. The Q_0 and Q_1 shift register outputs drive the two analog switch driver inputs.

In this way the backscatter signal is divided into two ranges and each range can be separately processed without affecting the other. The second range is then amplified with a NE592 video amplifier stage with variable gain from 4 to 200.

Finally, the first range of the 20 km backscatter (0 to 10 km) is amplified by the first two voltage video post amplifier stages by

at least 2000 (voltage gain) while the second range is amplified by at least 8000 (2000 for the first two stages and 4 for the single additional stage after the switching). The low frequency cut-off of the total amplifier is determined by the coupling capacitors and the input-output impedances of the amplification stages and is approximately equal to 100 Hz. The high frequency cut-off of the amplifier is determined by the 3 dB cut-off frequency of the low pass filters which are used to limit the receiver BW since both the transimpedance preamplifier module and the NE592 video amplifiers have an operational BW which exceeds frequencies of 40 MHz. Both the preamplifier module and the NE592 amplify from dc.

3.5 Summary

The design philosophy for an experimental OTDR system has been presented.

First the laser transmitter was discussed including the laser diode specifications, the laser driver circuit and the cooling device along with the cooler circuit. Then the optical directional coupler operation was briefly explained and the coupler specifications were given. Finally a pin-FET front end was chosen for the OTDR receiver. The preamplifier Lasertron module, the post amplifier and the range switching circuit were described.

In the next chapter the noise sources in an OTDR system are discussed and a brief front-end noise analysis will be made. Noise and SNR measurements will be presented and SNR enhancement techniques

including digital averaging will be discussed. Finally the hardware for the digital averaging technique will be designed and its operation will be explained.

2

CHAPTER 4

SIGNAL AND NOISE CONSIDERATIONS4.1 INTRODUCTION

The primary difference between conventional electrical and optical receivers is that in an optical system there is quantum noise, in addition to other common noise sources, which becomes shot noise in the detected signal current.

A second major difference between optical and electrical receivers is that the optical receiver is a current driven system, and for this reason optical front-ends are designed to minimize the noise current. This immediately implies that optical front-ends will have high impedance inputs and that 50 Ω input impedance amplifiers, while having a low noise voltage, are unsuited because of the high noise current associated with the 50 Ω . Usual considerations for the design of a low-noise preamplifier may not apply to an optical front-end.

In this chapter, all noise sources and their contribution to the total noise will first be discussed. Then a brief noise analysis of the front-end will be presented to point out the importance of various parameters in the total noise, and to support the selection of a p-i-n/FET front-end for the OTDR. Experimental noise current measurements follow and finally the SNR's for the backscatter signal from various distances are calculated. The need for SNR improvement is obvious since the SNR's for signals beyond 8 km are negative and at

longer distances, reach the -55 dB level. Methods for SNR improvement are discussed and the digital averaging technique is finally adopted. The implementation of the digital averaging technique is presented in detail and the operation of the digital averager is explained. Finally the logarithmic amplifier circuit and the "logging" operation advantages are presented.

4.2 Noise Sources in the OTDR System

In an OTDR system the noise that is added to the signal is introduced both by the receiver and by the transmitter of the system. The transmitter noise is much less important than the receiver noise, and at signals which return from fiber distances longer than 5km, the SNR is totally dependent on the receiver noise [21]. The noise from the optical receiver can be classified into three groups according to its origin. These are:

- a) Quantum noise which is always present in the signal;
- b) Noise introduced by the photodetector. This is a function of the photodetector type, design and temperature;
- c) Noise introduced by the pre-amplifier.

It is the object of the preamplifier design to ensure that, at minimum signal levels, the noise from the preamplifier is small compared with that from the other two sources, or failing in this, to

ensure that it is simply minimized. As far as transmitter noise is concerned, it was proved [21] that it dominates at near-end and mid-range fiber distances (0 to 5km) while it is negligible at longer distances.

The following subsections will discuss in detail, noise sources present in the transmitter and in the receiver.

4.2.1 The Transmitter Noise

The transmitter noise is the fluctuation in the output power of the laser successive pulses defined by, [21]

$$\sigma_{P_0} = \langle (P_0 - \langle P_0 \rangle)^2 \rangle^{1/2} \quad (4.1)$$

where,

σ_{P_0} is the transmitter noise standard deviation

P_0 is the laser pulse amplitude

and $\langle \rangle$ denotes mean value

Various semiconductor lasers at 1.3 μ m have been tested [21] with the laser case temperature being kept constant.

Under these conditions,

$$\sigma_{P_0}/P_0 = 0.05 \quad (4.2)$$

which is equivalent to an electrical, signal-to-transmitter noise ratio of 26dB. Although different batches of diodes may yield different performance levels this test is very useful in establishing the relative contributions of the various noise sources.

While the signal-to-transmitter noise ratio is constant [21], the backscatter signal amplitude falls exponentially with distance. Since the backscatter is directly proportional to the laser pulse peak amplitude P_0 , (2.19), then the transmitter noise standard deviation σ_{P_0} decreases exponentially in order to have the signal-to-transmitter noise ratio constant. It is expected that, at some distance z from the front end of the fiber link the transmitter noise standard deviation σ_{P_0} , will be equal to the receiver noise standard deviation σ_n . Beyond that point the receiver noise dominates. If the backscatter signal-to-total noise (transmitter and receiver noise) ratio is plotted versus the distance, as shown in Fig. 4.1, the total SNR remains constant until the receiver noise starts to dominate and then the SNR decreases. The knee in the SNR curve occurs at about 2-4km. However, in the experimental OTDR the transmitter noise is unimportant, because in the front range of the backscatter (0 to 2km), where it dominates, the backscatter signal has an adequate SNR (more than 10 dB) for the detection of splices.

Fig. 4.1: The backscatter signal to noise ratio at the receiver output of the experimental OTDR system as a function of the fiber link length.

4.2.2 The receiver noise

The noise in the optical receiver is introduced during photodetection and pre-amplification. There is also noise added to the signal at the post-amplifier but its importance to the overall receiver noise performance is negligible. The photodetector and the pre-amplifier (front-end) design determine the optical receiver noise performance and the OTDR dynamic range.

The basic optical pre-amplifier shown in Fig. 4.2 (a) can be modelled for noise estimation purposes [11] as in Fig. 4.2 (b). The noise sources included in this model, include contributions from the current flowing in the photodetector, i_d , thermal noise associated with the biasing resistors i_1 , and noise sources of the preamplifier i_a and e_a . The signal current generated by the incident light is indicated by $I_s(t)$. The capacitance C_d is the depletion capacitance of the photodetector and C_g is the stray capacitance associated with the interconnection of the detector to the input of the preamplifier. The resistor R_L represents the resistive portion of the circuit used to bias the detector and the input transistor. The generators i_d , i_1 , i_a and e_a characterize the various sources of noise in the system and Y_a is the preamplifier input admittance. The contribution to total noise from subsequent amplification (post-amplifier) can be neglected [11].

The noise sources can be grouped into two categories:

Fig. 4.2: a) Schematic of an optical preamplifier.
 b) The optical preamplifier is modeled with noise sources and a noiseless amplifier.

- 1) The photodetector noise, including the quantum noise or the optical signal dependent noise and the signal independent photodetector noise current; and
- 2) The preamplifier noise including biasing circuit noise as well as the preamplifier noise voltage and current sources.

The total input noise current density, when a pin photodiode is used, can be written as [11]

$$\frac{d\langle i^2 \rangle}{df} \Gamma = 2qI_s + 2qI_D + \frac{4KT}{R_L} + \frac{d\langle i_a^2 \rangle}{df} + \frac{d\langle e_a^2 \rangle}{df} |Z_{in}(\omega)|^{-2} \frac{A^2}{Hz} \quad (4.3)$$

where q is the electron charge

I_s is the photocurrent

K is the Boltzman constant

and $Z_{in}(\omega)$ is the total input impedance

The total input admittance can be written as [11],

(4.4)

$$Y_{in}(\omega) = Y_a(\omega) + 1/R_L + j\omega(C_d + C_s)$$

or $Y_{in}(\omega) = 1/R_{in} + j\omega C_T$ thus $Z_{in}(\omega) = R_{in} + 1/j\omega C_T$

where $C_T = C_d + C_s + C_a$, C_a is the preamplifier input capacitance and R_{in} is the resistive portion of the total input impedance. For a transimpedance preamplifier with a feedback resistor R_f and a bias resistor R_L , the third term of (4.3) has to be $4KT/R_f \parallel R_L$ instead of $4KT/R_L$. For a high input impedance preamplifier there is no feedback resistor R_f and R_L is chosen sufficiently high ($> 500k\Omega$) in

order to have a small contribution to the overall noise. It can be seen that the noise current density of a transimpedance preamplifier is higher than that of the high impedance one due to the additional factor $4KT/R_F$.

The first two terms of (4.3) belong to the first category of noise sources, the photodetector noise. The first term $2qI_s$ is the signal dependent noise or quantum noise which becomes shot noise at detection. It results from the randomness associated with the rate of arrival of the photons at the detector.

The second term $2qI_D$ is shot noise resulting from the constant dark current I_D .

The last three terms of (4.3), belong to the second category of noise sources, the preamplifier noise. The third term $4KT/R_L$ is thermal noise due to the bias resistor R_L .

The fourth term $d\langle i_a^2 \rangle / df$ is the preamplifier input noise current density due to the shunt current noise source i_a while the fifth term $d\langle e_a^2 \rangle / df |Z_{in}(\omega)|^{-2}$ is the preamplifier input noise current density due to the series voltage noise source e_a . The fifth term is a frequency dependent term and it turns out [11] that it is the limiting factor in the optimization of broadband optical receivers.

If the noise sources are assumed to be independent of frequency [11] over the range of interest which is the receiver bandwidth B , then using (4.3) the total mean square input noise is given by,

$$\langle i^2 \rangle_T = 2qI_s B + 2qI_D B + \frac{4KTB}{R_L R_F} + \frac{d\langle i_a^2 \rangle}{df} B + \frac{d\langle e_a^2 \rangle}{df} \int_0^B |R_{in} + \frac{1}{j\omega C_T}|^2 df \quad A^2 \quad (4.5)$$

It is commonly accepted [14][13] that the total receiver noise for low dark current photodetector such as a pin photodiode, is determined by the preamplifier circuit noise while the total receiver noise is strongly dependent on the photodetector dark current, when Ge APD is used as a photodetector, because Ge APDs, as shown in Table 4.1 [23] introduce a large amount of dark current. The dark current noise term at $\lambda=1.3 \mu\text{m}$, is much smaller for InGaAs pin photodiodes than Ge APDs as shown in Table 4.1. (In this table the InGaAsP APDs are not yet commercially available while the Si pin/APD photodiodes are for short wavelengths, $\lambda < 1 \mu\text{m}$).

4.3 Noise Considerations for pin-FET Front-End

In this OTDR system, because the received signal (backscatter) level is very low (tens of nW to pW), the quantum noise is small compared to the preamplifier noise. Even if Ge APDs offer the advantage of internal gain for the detected signal, while pin diodes offer no gain, at $\lambda=1.3 \mu\text{m}$ their noise is so high that a low noise preamplifier using a pin photodiode yields a better noise performance [12][13][14] than a preamplifier with a Ge APD.

When the preamplifier is built with a FET transistor, in a common source configuration as shown in Fig. 4.3(a), then the noise sources of the circuit according to (4.3), are as follows.

- 1) Thermal noise due to R_L . For a transimpedance design it will be thermal noise due to R_f and R_L . This noise current is the third term of (4.5).

TABLE 4.1

PARAMETERS OF STATE-OF-THE-ART PHOTODETECTORS

	<u>Si</u> <u>PIN/APD</u>	<u>Ge</u> <u>APD</u>	<u>InGaAsP</u> <u>APD</u>	<u>InGaAs</u> <u>PIN</u>
η	0.8	0.8	0.8	0.8
t_r (ps)	150	150	100	60
C_d (pF)	<1	<1	<1	<1
I_{DN} (nA)	1-10	100-500	0.01-1	1-10
I_{DM} (nA)	0.001	50-200	0.02-1	
k	0.01-0.03	0.7-1.0	0.3-0.4	
M_{max}	300	40	40	

Total Dark Current $I_D = I_{DN} + MI_{DM}$ Where M is Avalanche Gain.

η is the quantum efficiency, t_r is the rise time, and k is the excess noise factor of an APD.

Fig. 4.3: a) A pin-FET front-end circuit.
b) A high frequency model for the pin-FET front-end.

- 2) Shot noise due to FET gate leakage, given by $2qI_g$ where I_g is the gate leakage current. This noise current is the fourth term of (4.5) and because FETs have very small I_g (<10 nA) it is negligible.
- 3) Thermal noise associated with the FET channel transconductance which at the FET input is given by [11] $4KT\Gamma/g_m$, where g_m is the FET transconductance and Γ is a numerical factor which for a GaAs FET is ~ 1.1 [11]. This noise voltage is the largest part of the fifth term of (4.5).
- 4) Thermal noise due to R_D , part of which is induced to the input but its contribution to the input series noise voltage generator is relatively small [24] thus it can be neglected.
- 5) A frequency dependent flicker noise which induces to the FET input a noise component part of the fifth term of (4.5). This noise is negligible for frequencies approximately above 10 KHz for good devices [26][26] such as MESFETs [22]. At lower frequencies flicker noise for MESFETs is quite large varying from 10^{-19} to 10^{-17} A^2/Hz [27], but its contribution to the total receiver noise current (for a receiver bandwidth of a few Megahertz) is very small because the frequency range where the flicker noise dominates is a few thousandths of the total receiver bandwidth. Flicker is also limited by the receiver low cut-off frequency which is about 100 Hz.

Hence, for a FET front-end the principal noise sources are the bias/feedback resistor thermal noise and the channel transconductance noise and the total FET transimpedance front-end noise current can be written as:

$$\sigma_T^2 = \langle i^2 \rangle_T = 2qI_{SB} + 2qI_{DB} + \frac{4KTB}{R_L R_F} + \frac{4KT\Gamma}{g_m} \int_0^B \left| R_{in} + \frac{1}{j\omega C_T} \right|^{-2} df \quad A^2 \quad (4.6)$$

R_{in} is the resistive portion of the total input impedance thus it is a parallel combination of R_L and of the front-end input resistance. A transimpedance front-end, has a low-input resistance [19] thus the total input resistance of the front-end should be low. On the contrary the front-end input impedance due to capacitance C_T is very high, thus $R_{in} \ll 1/j\omega C_T$ or $R_{in} + 1/j\omega C_T \approx 1/j\omega C_T$ and (4.6) can be rewritten as:

$$\sigma_T^2 = \langle i^2 \rangle_T \approx 2qI_{SB} + 2qI_{DB} + \frac{4KTB}{R_L R_F} + \frac{4KT\Gamma B^3}{g_m} (2\pi C_T)^2 \quad A^2 \quad (4.7)$$

It can be seen from (4.7) that the limiting circuit noise for a FET varies as B^3 and also as C_T^2/g_m . Therefore high transconductance and low input capacitance are required for the FET front-end noise minimization.

In bipolar transistor front-ends, the dominant noise sources are the shot noise associated with the base and collector current and the thermal noise of the base resistance. Examining the minimum noise expressions for a bipolar front-end [11], the noise seems to vary as

the square of the bandwidth B in contrast to the cubic dependence for the FET and linearly with the capacitance compared to the quadratic dependence for the FET.

The fact that the bandwidth dependence of the noise of the bipolar front-end is "slower" than for the FET front-end [11] results in the bipolar one producing lower noise current at high frequencies whereas the FET front-end is superior at low frequencies. It is widely accepted and has been proved [23][11][26][27] from various researchers that a FET has a superior noise performance than a bipolar transistor for frequencies up to 100-200 MHz while bipolar transistors have superior behaviour above 200 MHz when both are used in an optical front-end design with a pin photodiode.

The front-end circuit and its specifications, for this OTDR receiver, are given in section 3.4.3. Additional specifications of the Lasertron front-end/receiver module can be used to calculate the total front-end noise current using (4.7).

- 1) The GaInAs pin dark current is $I_D=30$ nA and its capacitance is $C_d=0.6$ pF.
- 2) The GaAs MESFET transconductance $g_m=40$ mS and its gate leakage current $I_g=10$ nA. For GaAs FETs the numerical factor $\Gamma=1.1$.

- 3) The FET total input capacitance is not given, so it is assumed [26]
 $C_{in} \approx 2 \text{ pF}$.
- 4) The transimpedance resistor $R_F = 40 \text{ K}\Omega$.
- 5) The bias resistor R_L is much larger than R_F , so $R_F // R_L \approx R_F$ and the thermal noise due to bias circuit is determined by the smaller resistor R_F .
- 6) The receiver BW is 4 MHz.

Also the Boltzman constant $K = 1.381 \times 10^{-23} \text{ J/K}^\circ$, the absolute temperature $T = 298^\circ \text{K}$ (25°C) and the electron charge $q = 1.6 \times 10^{-19} \text{ C}$.

By omitting the quantum noise term ($2qI_s$) from (4.7), the pin-FET preamplifier total noise current σ_A (where $\sigma_T^2 = 2qI_s B + \sigma_A^2$) is calculated to be

$$\sigma_A^2 = 3.84 \times 10^{-20} + 1.64 \times 10^{-18} + 5 \times 10^{-21} \text{ A}^2 \quad (4.8)$$

$$\text{or} \quad \sigma_A = 1.3 \text{ nA}$$

where $3.84 \times 10^{-20} \text{ A}^2$ is due to the pin dark current

$1.64 \times 10^{-18} \text{ A}^2$ is due to the feedback resistor

and $5 \times 10^{-21} \text{ A}^2$ is due to the FET transconductance

It is obvious in (4.8) that the dominant source of noise in this Lasertron design is the feedback resistor thermal noise.

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4.4 Experimental Noise Measurements

The Lasertron front-end/receiver total noise current (equation 4.7) with quantum noise omitted) was also measured. The instrument set-up is shown in Fig. 4.4. An LED source with a flat response (± 0.5 dB) over the bandwidth of interest (0 to 20 MHz), is driven with a constant dc current. The continuous light emitted from the LED is detected at the pin-FET receiver. An ammeter is connected between the -5V supply and the detector bias pin (1) of the Lasertron receiver package to measure the photocurrent through the p-i-n diode. The output of the Lasertron (pin 7) is connected to a spectrum analyzer.

The total front-end/receiver input noise density can be expressed as

$$\frac{d\langle i^2 \rangle_T}{df} = 2qI_s + \frac{d\sigma_A^2}{df} \quad \text{A}^2/\text{Hz} \quad (4.9)$$

where I_s is the photocurrent due to the continuous light,

$2qI_s$ is the quantum noise density

and σ_A^2 is the Lasertron receiver noise power including pin dark current, bias/feedback resistor thermal noise and FET front-end noise.

When the LED is off, the spectrum analyzer displays the Lasertron receiver noise power density $d\sigma_A^2/df$. When the LED is switched on, the noise power is increased to $d\langle i^2 \rangle_T/df$ due to the additional quantum noise $2qI_s$.

Fig. 4.4: The experimental set-up for the noise current measurement of the Lasertron pin-FET preamplifier.

The difference in dB per Hz bandwidth between the two measurements is

$$10 \log \left[\frac{2qI_s + d\sigma_A^2/df}{d\sigma_A^2/df} \right] \quad (4.10)$$

A knowledge of I_s allows σ_A to be calculated. Measurements were made from 40 KHz to 20 MHz at a photo current level $I_s = 5.56 \mu A$. The measurements have a 0.5 dB accuracy. The calculated values for the Lasertron preamplifier noise current density σ_A' (where $\sigma_A'^2 = d\sigma_A^2/df$) are listed in Table 4.2.

TABLE 4.2

THE LASERTON PREAMPLIFIER NOISE CURRENT DENSITY σ_A'

<u>FREQUENCY</u>	<u>σ_A' (pA/$\sqrt{Hz}$)</u>
40 KHz	1.74 $\pm$ 0.2
50 KHz	1.08 $\pm$ 0.1
60 KHz	0.77 $\pm$ 0.06
100 KHz	0.58 $\pm$ 0.04
8 MHz	0.58 $\pm$ 0.04
10 MHz	0.67 $\pm$ 0.05
14 MHz	0.67 $\pm$ 0.05
16 MHz	0.77 $\pm$ 0.06
18 MHz	0.67 $\pm$ 0.05
20 MHz	0.67 $\pm$ 0.05

The average noise current density is approximately 0.7 ± 0.05 pA/ $\sqrt{Hz}$ and for a 4 MHz receiver BW, the total noise current for the Lasertron

All the operations performed for a sample at one point, namely digitization—read from RAM—add—store to RAM—increase i , have to be completed, before the next sample is taken. Hence the sampling interval cannot be less than a value determined from the system hardware delays. Therefore the objective is to design the averaging system in such a way that it meets the sampling theorem requirements. The backscattered signal has a BW of 1 MHz so the sampling rate has to be at least 2 MHz. The averager was designed to have a 400 ns sampling interval or 2.5 MHz sampling rate. The number of samples for a 20 km trace is 500 since a 1 km round-trip requires 10 μ s.

It can be shown [28] that digital averaging increases the SNR in proportion to the square root of the number of additions or

$$\text{SNR}_A = \text{SNR}_B + 20 \log \sqrt{N} \quad (4.12)$$

where SNR_A is the SNA after averaging N times

and SNR_B is the SNR before averaging

However, in the digital averager the improvement is limited by quantization noise introduced during digitization in the analog to digital (A/D) converter. Originally SNR_A increases along with the averaging time but at some N value [30] starts to decrease as quantization noise enlarges. The value of N at which this occurs depends on the A/D converter word length. For an 8-bit word A/D, it is shown in [30] that the maximum improvement in SNR is 50 dB which is obtained after averaging approximately 2^{19} times. Because a 40 dB improvement is required to reach 15 km, then the 8-bit A/D meets the system requirements.

directional coupler two-way loss must also be taken into account. The laser pulse peak power emitted directly from the laser package pigtail is 5 mW (optical) or 7 dBm. The optical coupler loss from the laser to the fiber is 3.7 dB and from the fiber to the detector/front-end is 5 dB, thus the total coupler loss is 8.7 dB. This value also includes the loss in the splices to the laser and detector pigtails, and to the fiber link. Thus the equivalent optical power actually launched into the link is approximately -2 dBm or 0.63 mW. The laser pulse width is 1 μ s. The Rayleigh scattering coefficient at $\lambda=1.3 \mu$ m is $a_g=1.5 \times 10^{-4}$ Np/m. The capture fraction S is given by (2.3) or $S \approx 0.005$ for the fibers which are used. Finally, $v=2 \times 10^8$ m/s and the total attenuation is $a_t=1$ dB/km or 2.3×10^{-4} Np/m. From (2.16), the backscattered light from 0 km is $P_b(0)=50$ nW (optical). The photocurrent is $I_s(0)=P_b(0) \cdot R$ and since the Lasertron pin responsivity R is 0.8 A/W then $I_s(0)=40$ nA.

The SNR at 0 km may now be calculated. From (4.11) $\sigma_A=1.4$ nA. The quantum noise calculated from $\sigma_Q^2=2qI_sB$ for a bandwidth $B=4$ MHz, has a mean square value $\sigma_Q^2=\langle i_Q^2 \rangle=5.7 \times 10^{-20} \text{ A}^2$ or the quantum noise current is $\sigma_Q=0.2$ nA. The total noise current then is $\sigma_T=1.6$ nA. Hence,

$$\text{SNR}(0)=10 \log \left[\frac{I_s(0)}{\sigma_T(0)} \right]^2 = 27 \text{ dB (electrical)}$$

At 10 km, the SNR is 40 dB less than at 0 km

$$\text{SNR}(10)=-13 \text{ dB}$$

At 20 km the SNR is 80 dB less than at 0 km

$$\text{SNR}(20)=-53 \text{ dB}$$

The above calculations give the SNRs at the preamplifier input or output, considering all noise sources introduced at detection and preamplification. Although the noise added in the post-amplification section, was considered negligible, there is a small SNR deterioration due to the following amplification stages, approximately equal to 1.3 dB according to the NE592 video amplifiers (used as post-amplifiers) data sheets. Taking this deterioration into the account the receiver output SNRs become

$$\text{SNR}(0) = -25.7 \text{ dB}$$

$$\text{SNR}(20) = -54.3 \text{ dB}$$

4.6 Methods for SNR Improvement

By averaging a periodic signal corrupted by additive uncorrelated noise, the desired signal component is enhanced relative to the noise component [28]. The electrical SNR can be increased in proportion to the number of additions N . There are two well known methods to implement the averaging process. The boxcar averager or boxcar integrator and the digital averager.

4.6.1 The Boxcar Integrator

The boxcar integrator [29] involves opening a gate repetitively at a specific instant on the input waveform and integrating the resultant output. This output will be the average height at the selected point (one channel) of the signal. A slow variation of the gate delay time leads to a slow output signal whose shape approximates the true shape of the input signal with reduced noise. The degree of

signal recovery from noise, depends upon the number N of samples taken at each point on the input waveform, and if the number is large then the system will take a long time to complete the signal averaging process. Slow DC drift or flicker noise then becomes a serious problem. However, even with small N the signal from the boxcar is a replica of the input signal with somewhat reduced noise [29].

From each backscattered trace period, only one point (channel) is selected and averaged (integrated) during the next N periods. The boxcar averages over N samples at the same point (channel) in the trace and then, while the averager stores this average value, the boxcar moves to the next point and again averages over N samples and so on.

The boxcar technique is time consuming. If M is the number of sample points making up one period of the signal and N is the number of periods over which the averaging is performed then the boxcar requires NM periods to average N times.

The speed of the averaging could be drastically improved if all M points are selected in one period and averaged over N periods. A multichannel digital averager, which will be discussed below is able to perform N times averaging in N periods

4.6.2 The Digital Averager

The digital averaging technique or multichannel technique is a process wherein the noisy signal is first sampled and digitized, and then the corresponding samples from each cycle are added together in

form of digital words, in order to increase the SNR.

In more detail, M samples are taken from one period of the repetitive signal and digitized. Those M digital words are stored in a Random Access Memory (RAM). During the next signal period, another M new samples from the same points of the signal are digitized, added to the corresponding previous cycle samples and finally the sum is stored in the RAM replacing the previously stored samples. In every cycle, the M new samples are added to the corresponding already stored batch of M sums. Therefore the RAM at the end of every cycle contains M digital words, each one of those, is the result of the addition of N digitized samples, taken from one point of the trace, where N is the number of signal periods being swept, including the current one. Thus averaging a repetitive signal N times requires a time interval N periods long, independent of the number of samples taken from each signal period.

There are two parameters that determine the quality of the digital averager: the sampling rate which determines the resolution of the system and the averaging time which determines the SNR improvement. The sampling time is related to M while the averaging time is proportional to N .

The first operation in the digital averaging technique is the sampling and digitization of the analog trace. The sampling rate should be at least twice the maximum frequency component of the analog signal, as the sampling theorem states. On the other hand the sampling interval is limited by the hardware delays of the averaging system.

All the operations performed for a sample at one point, namely digitization—read from RAM—add—store to RAM—increase, have to be completed, before the next sample is taken. Hence the sampling interval cannot be less than a value determined from the system hardware delays. Therefore the objective is to design the averaging system in such a way that it meets the sampling theorem requirements. The backscattered signal has a BW of MHz so the sampling rate has to be at least 2 MHz. The averager was designed to have a 400 ns sampling interval or 2.5 MHz sampling rate. The number of samples for a 20 km trace is 500 since a 1 km round-trip requires 10 μ s.

It can be shown [28] that digital averaging increases the SNR in proportion to the square root of the number of additions or

$$\text{SNR}_A = \text{SNR}_B + 20 \log \sqrt{N} \quad (4.12)$$

where SNR_A is the SNA after averaging N times

and SNR_B is the SNR before averaging

However, in the digital averager the improvement is limited by quantization noise introduced during digitization in the analog to digital (A/D) converter. Originally SNR_A increases along with the averaging time but at some N value [30] starts to decrease as quantization noise enlarges. The value of N at which this occurs depends on the A/D converter word length. For an 8-bit word A/D, it is shown in [30] that the maximum improvement in SNR is 50 dB which is obtained after averaging approximately 2^{19} times. Because a 40 dB improvement is required to reach 15 km, then the 8-bit A/D meets the system requirements.

4.7 The Experimental Digital Averager

4.7.1 Description of the Averager

The block diagram of averager circuit shown in Fig. 4.5(a) consists of a sample-and-hold circuit, an 8-bit A/D, a few 4-bit latches, 8-bit tri-state output latches, 4-bit full-adders, decade and hexadecimal programmable counters and a few 1K x 4-bit RAMs. The clock circuit is not indicated in Fig. 4.5 but it will be discussed separately later.

The averaging circuit operates as follows.

Assume that all latches, adders, counters and RAM are cleared during "switch-on". A synchronization circuit, part of the timing circuitry, switches the averager on (activates the clocks C_1 to C_8) at the beginning of a backscatter signal cycle, when an "average" command is given to the system.

Clock C_1 strobes the S/H - A/D circuit to sample and digitize the first sample S_1^1 , of the first point P_1 from the first cycle as shown in Fig. 4.5(b). Counter (1) which operates as the memory addresser indicates memory location M_0 . RAM is in READ mode.

Clock C_2 strobes latches (1) and (2). Latch (1) latches the 8-bit word for S_1^1 while latch (2) reads the value for M_0 from RAM which is zero because RAM has been originally initialized (cleared).

Fig. 4.5: a) The block diagram of the averager operating in the averaging mode.

b) The sampling of the backscatter waveform.

As soon as latches (1) and (2) transfer their input data to their outputs, the adder starts adding.

Clock C_3 strobes latch (3) to latch the adder output, which is S_1^1 , but the latch (3) output is still in the tri-state condition.

Clock C_4 switches the RAM to WRITE mode.

Clock C_5 enables the latch (3) output and S_1^1 is written to RAM in location M_0 .

Clock C_6 switches RAM to STAND-BY mode (idle) while the latch (3) output returns to tri-state, and RAM during stand-by switches to READ mode.

Clock C_7 increments the counter (1) and it indicates the memory location M_1 .

Here ends the first sampling interval. The time separation between the clock pulses is the maximum required for proper completion of each function (i.e. add, store, etc.).

After the clock C_7 pulse, the same pulse sequence of 400 ns total duration, starting from clock C_1 , is applied to the second point P_2 , from the first cycle and finally S_1^2 is stored in the M_1 location, and so on, until the first sample S_1^{500} , of point P_{500} from the first cycle will be stored in the location M_{499} .

One averaging cycle of duration 200 μ s (500 points $\times$ 400 ns) has been completed. At that time counter (1) returns from 499 to 0 (it is preprogrammed to do so) indicating again M_0 while counter (2) counts one ($N=1$) averaging time.

The second averaging cycle starts now, with the same clock sequence as before (C_1 to C_7). the second sample S_2^1 of the first point is added to S_1^1 and stored in location M_0 and so on until the second sample S_2^{500} of the last point is added to S_1^{500} and stored in location M_{499} . Counter (1) returns from 499 to 0 and counter (2) counts two ($N=2$) averaging times.

When the desired number of averaging times is completed then counter (2) disables the clocks and the 500 sums of the samples of each point are stored in the 500 RAM locations. These data remain in the RAM and can be displayed at any time through a digital-to-analog (D/A) converter using a display clock circuit.

4.7.2 Operation of the Clock Circuit

The clock circuitry for the averager system produces eight clocks (C_1 to C_8) as shown in the timing diagram, Fig. 4.6. The clocks perform the following functions.

Fig. 4.6: Timing diagram for the averager clocks.
 C_1 to C_8 operate in the averaging mode
 C_9 and C_{10} operate in the display mode.

- 1) Clock C_1 strobes the S/H circuit and the A/D conversion process. It also strobes the counters associated with range-switching (section 3.4.5). The C_1 pulse width is 20 ns.
- 2) Clock C_2 , having pulse width 30 ns, strobes latches (1) and (2).
- 3) Clock C_3 strobes latch (3) to latch and has a 30 ns pulse width.
- 4) Clock C_4 is the RAM "WRITE ENABLE" ($\overline{WE}$) pulse with a width of 120 ns.
- 5) Clock C_5 is the latch (3) "OUTPUT CONTROL" ($\overline{OC}$) pulse that switches latch (3) output from tri-state to active. Its width is 104 ns.
- 6) Clock C_6 is the RAM "CHIP SELECT" ($\overline{CS}$) pulse that switches RAM from active to stand-by where the RAM output is in high-impedance (hi-z) state. Its width is 40 ns.
- 7) Clock C_7 , having pulse width 35 ns, strobes the RAM addresser counter (1).
- 8) Clock C_8 clears latches (1) and (2). It is an inverted C_6 clock.

These eight clocks are generated by a master 10 MHz clock. The master clock generator circuit consists of two inverters for the 360° phase shift required for oscillation, a 10 MHz crystal and resistors and capacitors. As shown in Fig. 4.7(a) the 10 MHz master clock (CK), after a start/stop control circuit, is divided by 4 and then it is fed

Fig. 4.7: (a) Block diagram of the circuit which generates ten clocks $C_1 \dots C_{10}$ for the averager from the 10 MHz master clock CK .
 (b) Timing diagram for the master clock and the four original clocks.

to a Delay Flip-Flop (D-F/F) (54LS174). The resulting waveforms are also shown in Fig. 4.7(b). The top waveform is the master clock CK; next is the divided-by-4 clock, then the inverted-by-4 and finally the four D-F/F outputs CK1, CK2, CK3 and CK4. These four waveforms are called "original clocks" and all eight clocks are generated from them with variations in pulse width, introducing additional delays and gating. Monostables 74121 or 74123 (dual) that set the clock pulse width, delays which are introduced by inverter gates 74LS04 or 74S04 and NAND, NOR AND gating using 74LS00, 74LS02, 74LS08 and 74LS10 constitute the rest of the clock/timing circuitry. Schottky IC's are used throughout for high speed operation.

The "original clocks" are first AND gated with the inverted master clock (CK) using a 74LS08. Then C_1 is generated by CK1, C_2 by CK2, C_3 , C_4 and C_5 by CK3 C_6 and C_7 by CK4. All the delays between subsequent clock pulses are long enough to allow completion of the operations, avoid output to output connections between two or more IC's and avoid wrong writing to the RAM or wrong reading from the RAM. The pulse widths are wide enough to ensure proper strobing to all latches, RAMs and counters. Finally the total duration for one point processing is less than 400 ns thus the averager design finally meets the sampling rate requirements.

4.7.3 The Initialization and Synchronization Circuitry

The two previous sections discussed the basic operation of the averager and the clock circuitry that controls it. Additional circuitry is required to synchronize the beginning of averaging with

beginning of the backscatter signal cycle, to clear all latches, counter, adders and RAM, to start the averaging by enabling the clocks to run in proper order and finally to disable the clocks at the proper time when the desired averaging has been completed.

The laser pulser circuit is strobed by a 5 KHz pulse train generated by dividing the 10 MHz master clock (CK) by 2000, as shown in Fig. 4.8(a). This synchronizes the pulser to the averager. The laser pulser and the laser diode operate separately from the averager/clock circuitry. When the laser is on and backscatter is detected, and it is desired to start averaging, then with the externally applied "start averaging" command the clock circuit should remain idle, wait for the laser pulse that coincides with the beginning of the backscatter trace and, when this pulse is detected, start averaging by letting the clocks run. These operations are performed by the circuitry of Fig. 4.8(a). The external "on" command through F/F(1) and F/F(2) and AND gate allows the master clock to run to the counter $\div 4$ and to the four D-F/F where the four "original clocks" are produced.

During the first averaging cycle the clear inputs of latches (1) and (2) are kept low in order to clear the RAM. At the end of the first cycle the address counter (1) strobed by the 500th C_7 clock pulse switches from 499 to 500. That 500 is decoded by means of a NAND gate and initializes the address counter (1). Thus the addresser counters count from 0 to 499 and then goes back to 0. Counter (2) counts all those initialization pulses from NAND gating in order to keep track of the number of averaging N and when the desired N is completed then counter (2) triggers F/F(3) to stop averaging by preventing the master

Fig. 4.8: (a) The initialization and synchronization circuit diagram of the averager system.

Give the external
command to start
the averaging process.

Fig. 4.8 (b): Initialization process for the onset of averaging.

clock to enter the eight clock generation block, thus disable all eight clocks.

The externally applied initialization which occurs each time the averager is "switched on" consists of:

- 1) Clearing F/F(1), F/F(3) and F/F(4);
- 2) Clearing the four "original clocks" generator, the D-F/F;
- 3) Loading the desired preset N to counter (2).

4.7.4 The Averager Circuitry

The averager system was designed to perform averaging 2^{16} times. Since an 8-bit A/D is used, the maximum word length which can occur when averaging 2^{16} times is a 24-bit word. Hence a 24-bit word length RAM is required and all data buses, except the 8-bit one from the A/D to latch (1) and to the adder are 24 bits. The RAM address bus, since 500 points are sampled and stored, should be 9 bits. The IC's that constitute the main averager circuit are as follows.

- 1) The A/D converter is an 8-bit TRW monolithic video, flash A/D converter evaluation board, model IDC1007PCB with a minimum of 20 megasamples per sec conversion rate. It will accept and digitize a 0 to 1 V analog signal. Its dynamic range is 48 dB (electrical).

- 2) One quad D-F/F with clear, the 74LS175, is used for latch (1) and four hex D-F/Fs with clear, 74LS174, are used for latch (2).
- 3) Six Schottky 4-bit full-adders, 74S283, with fast carry are used for the adders.
- 4) Three Schottky, octal D-type, tri-state output latches, 74S374 are used for latch (3).
- 5) Six 1K by 4-bit fully static RAMs are used in parallel. This model is a high speed product from National, the NMC2148HJ-2 with 45 ns access time. All six RAMs have the same address bus.
- 6) Three synchronous programmable up/down, 4-bit binary counters 74LS193 form the RAM addresser. They count up to 499 and then clear. There are also the "averaging times" counters, five 74LS192 synchronous programmable up/down decade counters. The "averaging times" number N is selectable and may be any number from 0 to 2^{16} .

4.7.5 The Display Section

When the N times averaging is finished then the RAM contains the sum of N samples for each point, for all 500 points stored at 500 different memory locations, 24 bits each. Those 24 bit words should be divided by 2^{16} (N) in order to have the original averaged trace.

Division by 2^{16} is equivalent, in a digital system, to shifting to the left by 16 bits or for a 24 bit word division by 2^{16} can be accomplished by reading only the first most significant 8 bits, D_{23} to D_{15} , and ignoring the bits D_{15} to D_0 . Hence the D_{23} (MSB) to D_{15} (LSB) 500 8-bit words after Digital to Analog (D/A) conversion and proper display timing reproduce the originally sampled trace but with reduced noise. From Fig. 4.9(a) it is seen that the 8 most significant bits of the RAM are read by the octal latch (4) every 400 ns, during a 100 ns READ mode cycle (while at the remaining 300 ns RAM is in STAND-BY mode). Then these 8-bits are fed to a D/A and the D/A output may be displayed on an oscilloscope screen. The octal latch is a 74LS374 one with its output continuously enabled ("control output" is low). The D/A is an 8-bit, high speed current-output device, National model DAC0800, with 0.1% non-linearity and 100 ns settling time.

The display clocks are generated by CK4 with monostables and delays. The addresser counters (1) operate as in the "average mode" of operation strobed by C_1 . The "display mode" clock timing diagram is shown in Fig. 4.9(b).

Fig. 4.9: a) The block diagram of the display circuit.
b) Timing diagram for the display mode.

4.8 Overflow Detecting Averaging Process

The experimental averager memory word length is 24 bits, and the A/D is 8 bits. In the sampling-digitization process the maximum backscatter signal amplitude (the one at 0 km) corresponds to the A/D maximum analog signal input (1V) and a digitized output consisting of all "ones" or 11111111. The maximum averaging time in order to avoid overflow at memory location M_0 is $2^{16} + 257$ times. (If the number 11111111 or 255 is multiplied by $2^{16} + 257$ then an all "ones" 24-bit word is obtained).

If the averaging continues beyond $2^{16} + 257$ times, then an overflow occurs at M_0 , during the next averaging cycle ($2^{16} + 258$). The next sample for P_1 is added to the all "ones" 24-bit word which is stored at M_0 and the sum is a 25-bit word with the MSB equal to "one" and all bits except the least significant 8 are "zero". According to binary addition when an overflow occurs, M_0 clears and accumulation and adding of the new samples starts again from zero. When the averaging continues beyond $2^{16} + 258$ times and because the backscatter is a decaying exponential, then all memory locations from M_1 to M_{499} overflow sequentially, thus clear and start accumulating new samples.

However this "overflow, clear, start from zero" process does not destroy all previous averaging improvement. Consider a noiseless backscatter trace, as shown in Fig. 4.10(a). Since the display circuit of the averager introduces a division by 2^{16} this type of averaging performs the following operation.

$$S_n^i(nT) = S_{n-1}^i(nT) + x_n^i(nT) / 2^{16} \quad (4.13)$$

Fig. 4.10: (a) A noiseless backscatter trace; ϵ is the averager memory limit.
 (b) The averager display of the averaged backscatter trace of (a), during the continuous averaging process.

with $S_0^i=0$ and $n=1,2,\dots,N$

where $S_K^i(nT)$ is the sum of samples of point P_i at the n th averaging cycle and T is the averaging cycle.

$x_n^i(nT)$ is the sample of P_i at the n th cycle and N is the number of averaging times.

From (4.13) it can be seen that the averaged signal amplitude increases continuously as the averaging continues. When adding (or averaging) 2^{16} times has been completed then the averaged waveform is a replica, with greatly reduced noise, of the input backscatter signal, since a division by 2^{16} is introduced at the display. If averaging continues up to $2^{16} + 257$ times then the M_0 location is ready to overflow. Averaging once more causes an overflow at M_0 to occur and the OTDR trace returns to zero at that point. As averaging continues, $M_1, M_2, \dots, M_{499}$ return to zero in succession, while the memory locations which have already overflowed start to accumulate new samples and the signal builds up again. The new averaged trace seems to break into two portions. The first portion which builds up after the overflow seems to be a continuation of the second one, where overflow has not occurred yet, but, with its dc (ϵ) removed. With this dc subtraction the amplitude of the first portion is within the limits of the memory.

Consider two points P_1 and P_K which are KT sampling intervals apart ($T=400$ ns) as shown in Fig. 4.10(a) or (b). The quantized values of P_1 and P_K are x and y respectively, and they are related by the equation $y=xe^{-aKT}$ where a is the attenuation coefficient, K is an

integer from 1 to 500 and T is the sampling interval. Memory overflow occurs each time, after continuous adding, the value of each point of the averaged trace exceeds the 24-bit word length or in terms of analog signal exceeds a maximum value ϵ .

At the first averaging cycle ($N=1$) M_0 contains x and M_{K-1} contains y . We assume that after averaging N_1 times, M_0 overflows, and after N_K times M_{K-1} overflows. After averaging N_1-1 times, M_0 contains $(N_1-1)x$ which is less than ϵ by a number R_1 thus $\epsilon = (N_1-1)x + R_1$. The number R_1 is the residue from the division ϵ/x and it is less than x . Hence:

M_0 contains $(N_1-1)x$

M_{K-1} contains $(N_1-1)y$ and

the vertical difference between $P_1(M_0)$ and $P_K(M_{K-1})$ is $(N_1-1)(x-y)$.

After averaging N_1 times, M_0 overflows. M_0 should contain $(N_1-1)x + x + R_1 - R_1$ or $\epsilon + x - R_1$ and because the dc(ϵ) is removed M_0 contains $x - R_1$, M_{K-1} contains $N_1 y$ and the vertical difference between M_0 and M_{K-1} is $\epsilon - N_1 y + (x - R_1)$ or $(N_1-1)x + R_1 - N_1 y + (x - R_1)$ or $N_1(x-y)$.

After averaging N_K-1 times,

M_0 contains $[(N_K-1)-N_1]x + (x - R_1)$

M_{K-1} contains $(N_K-1)y$ which is less than ϵ by a number R_2 thus $\epsilon = (N_K-1)y + R_2$ and R_2 is the residue from the division ϵ/y and it is less than y .

The vertical difference between M_0 and M_{K-1} is, $[(N_K-1)-N_1]x + (x - R_1) - R_2$ or $(N_1-1)x - \epsilon - R_2$ or $(N_K-1)(x-y)$.

After averaging N_K times then

$$M_0 \text{ contains } [(N_K-1)-N_1]x + (x-R_1) + x$$

M_{K-1} should contain $(N_K-1)y + y$ or $(N_K-1)y + y + R_2 - R_2$ or $\epsilon + y - R_2$ and because the dc (ϵ) is removed M_0 contains $y - R_2$.

The vertical difference between M_0 and M_{K-1} is,

$$[(N_K-1)-N_1]x + (x-R_1) + x - (y-R_2) \text{ or } N_K x - \epsilon - (y+R_2) \text{ or } \underline{N_K(x-y)}.$$

It can be seen that the vertical difference between P_1 and P_K while averaging continues, increases proportionally to the number of averaging times even if overflow occurs. After averaging N_K times, the M_0 content is $X_{N_K} = (N_K - N_1 + 1)x - R_1$ and the M_{K-1} content is $Y_{N_K} - y - R_2$. Since an overflow has occurred at both locations then the real values for X_{N_K} and Y_{N_K} , if the memory capacity were unlimited, would be $X_{N_K}(\text{real}) = X_{N_K} + \epsilon = N_K x$ and $Y_{N_K}(\text{real}) = Y_{N_K} + \epsilon = N_K y$. The exponential relationship between $X_{N_K}(\text{real})$ and $Y_{N_K}(\text{real})$ is obvious since each one is directly proportional to x and y respectively.

From this analysis it is concluded that memory overflow does not disturb the averaging process and it does not distort the averaged signal. The overflow is "detected" by the system, the dc component (ϵ) is removed and averaging continues. Hence, this averaging method can be called simple addition averaging with "overflow detection".

In reality the backscattered signal has additive noise which at approximately 7.5 km becomes equal to the signal level and beyond that point the SNR becomes negative. The noise fluctuations are many times

larger than the pure backscatter signal amplitude for distances beyond 10 km. Even if the range switching circuit is used and additional amplification of the second range (10 to 19.5 km) is introduced, the additional amplification is limited because of the maximum 1V A/D analog input signal amplitude and because of the large noise fluctuations. For example at 18 km the noise fluctuations are about 400 mV and the backscatter signal is approximately 0.7 mV. Amplification can only double the amplitude of the second range without saturating the A/D converter. The signal at 18 km with this amplification becomes 1.4 mV but it is still less than the A/D quantization step (4 mV). Note that at least 200 mV dc offset is added to the backscatter + noise signal before quantization in order to keep it in the 0-1V operating range of the A/D. Even if the pure signal is less than the A/D quantization step the superposition of the noise fluctuation and the dc offset allow the signal to be quantized.

Since a division by 2^{16} is performed prior to display, then after averaging 2^{16} times the displayed signal is a replica of the input signal but with greatly reduced noise. An 8-bit D/A is employed for display. All backscatter points that have amplitude similar to the D/A quantization step will be represented by one level. This would make the system useless for distances beyond 14 km because only two or three levels are used for a 6 km range or a 12 dB (optical) range.

If averaging 2^{16} times is repeated instead then the displayed waveform is actually amplified two or more times and the 1.4 mV backscattered signal from 18 km will become 2.8 mV or 4.2 mV and so on. In this manner the quantization levels for the last 6 km range are utilized more effectively and the recovered signal range is extended.

The number of repetitions of the 2^{16} averaging cycle is limited by the accumulation of the A/D quantization noise. The standard deviation of the A/D round-off error is given by [28].

$$\Delta(N) \equiv \sqrt{\Delta_n^2/N + \frac{4N}{7} \Delta_q^2} \quad (4.14)$$

where Δ_n^2 is the noise variance

and Δ_q^2 is the quantization noise variance

Hence, the SNR_A after averaging N times can be written as [30].

$$\text{SNR}_A = -10 \log \left(\frac{10^{-(\text{SNR})_B/10}}{N} + \frac{4}{7} N \Delta_q^2 \right) \quad (4.15)$$

where SNR_B is the SNR before averaging

For an 8-bit A/D the maximum SNR improvement, as stated in [30], is 50 dB and is equivalent to approximately 2^{20} times averaging or 3.5 minutes averaging time. With the experimental digital averager more SNR improvement can be obtained, as reported in Chapter 5 Fig. 5.9, with longer averaging up to 5.5 minutes.

4.9 The Logarithmic Amplifier

The main problem in the backscatter signal quantization process is that it is an exponential waveform and the quantization levels are not equally distributed along the trace. In the front range n levels may cover a 2 km distance or 4 dB (optical), while in the far end range these n levels may cover a 10 km distance or 20 dB. Hence, a logarithmic amplifier is required to transform the exponential trace

into a linear one and then the resolution will be the same over both ranges. This can be seen in Fig. 4.11(a). In addition, a recent study [27] on the effects of logarithmic conversion (logging) of signals showed that SNR improvement is expected at very low and negative SNRs.

The logarithmic amplifier circuit shown in Fig. 4.11(b) is a well-known design [31] with a few modifications in order to increase its bandwidth. The circuit operation depends on the precise logarithmic relationship between the collector current and emitter-base voltage of transistor Q_1 . The op-amp U_1 , with Q_1 as the feedback element, generates a logarithmic output voltage for a linear input current. The op-amp U_2 is used as the reference current supplier. Q_1 and Q_2 are a matched pair of transistors integrated in the same chip, LM394. The op-amps U_1 and U_2 are of the type LH0032C with wide bandwidth (70 MHz) and very low input bias current because of the J-FET differential input. The logarithmic converter bandwidth is approximately 1 MHz and its input dynamic range is about 35 dB. The log output is 99% accurate for any input current between 100 nA and 500 μ A.

The logarithmic amplifier follows the OTDR receiver post-amplifier output buffer. This buffer serves also to introduce a dc offset to the receiver output signal before it is applied to the logarithmic amplifier (log-amp). Since the log-amp output is given by [31].

$$E_{out} = \frac{-KT}{q} \frac{R_1 + R_2}{R_2} \ln \left(\frac{E_{in} R_3}{E_{ref} R_{in}} \right) \quad (4.16)$$

Fig. 4.11: (a) A comparison of the distribution of the A/D quantization steps in an exponential and in a linear trace.
 (b) Circuit schematic of the logarithmic amplifier of the experimental OTDR.

then the input signal to the log-amp must be positive when the reference voltage is positive. A positive dc offset voltage is added to the signal waveform in order to make the signal + noise waveform positive. The dc offset introduces some distortion [27] to the logging conversion. If $x(t)$ is the signal and $\log[x(t)]$ is the logarithm of it, then with a D volts dc offset, the log-amp output will be $\log[x(t)+D]$. For $x(t) \gg D$, which occurs at short distances, then $\log[x(t)+D] \approx \log[x(t)]$ but when $x(t) \ll D$, at long distances (beyond 14 km) then $\log[x(t)+D] \approx \log D$ which is a constant distortion.

The logarithm of the backscatter exponential is a linear trace having a negative slope. The distortion introduced by the dc offset tends to decrease the slope of the linear trace to zero at long distances as seen in Fig. 5.8. This distortion is the only disadvantage of the log-amp.

The log-amp has two advantages when it is introduced before digitization. Firstly, it improves the low SNRs. Secondly, by converting the exponential backscatter to a linear one it increases the resolution of far range (>14 km). It extends the OTDR reach but it introduces some distortion to the far range signal. This distortion can be removed before trace display.

In this OTDR system, the use of the log-amp is optional as determined by the settings of the switches SW1 and SW2 in Fig. 3.1.

CHAPTER 5

OTDR PERFORMANCE MEASUREMENTS

In this chapter, the results from the testing of a 19.5 km fiber-optic link with the experimental OTDR system are presented. The link was made from nine reels of fiber spliced together with arc-fusion splices. The fiber was probed from both ends and photographs of the fiber signature were taken. The peak laser pulse power effectively launched into the fiber, including the forward and return loss of the coupler and optical losses from the splices, was -2 dBm or 0.63 mW. The fiber link average loss was approximately 0.7 dB/km. With the splice loss included the total link loss was measured, using the insertion loss technique, to be about 20 dB. Thus the backscattered signal had a 40 dB (optical) dynamic range. The expected backscatter SNR at 19.5 km is about -55 dB. Since the SNR improvement through averaging is limited by the A/D quantization noise, the averaging time was approximately limited to 2^{20} times or about 5.5 minutes averaging.

The effects of averaging are shown in the sequence of photographs in Figs. 5.1-5.11.

The first photograph, Fig. 5.1, shows the backscatter trace of the 19.5 km link without averaging. A 20 μ s time interval is equivalent to a 2 km one-way distance. The fiber far-end Fresnel reflection is indicated. The fiber front-end reflection is suppressed, as reported in section 3.3. The first three splices at 2, 4 and 6 km can be identified but beyond 6.5 km, the backscatter signal is buried in the noise.

Fig. 5.1: Experimental OTDR display of the backscattered signal from the 19.5 km fiber link (without averaging).

Fig. 5.2: Experimental OTDR display of the first range of the backscatter from 0 to 10 km (without averaging).

Next this trace is divided into two ranges of length 10 km and 9.5 km, respectively, and each range is processed separately. The first (0 to 10 km) or front range is shown in Fig. 5.2 and the second (10 km to 19.5 km) or far range is shown in Fig. 5.3. The second range has an additional amplification of 6 dB (electrical) over the first range, in order to have the 10 to 19.5 km backscatter signal cover the logarithmic amplifier dynamic range. The far-end Fresnel reflection is indicated.

The logarithm of the first range signal is shown in Fig. 5.4. The logarithmic conversion is expected to enhance the backscatter signal and improve the negative SNRs. The noise fluctuation enlargement at longer distances (comparing the "logged" trace of Fig. 5.4 with the exponential trace of Fig. 5.2 where the noise fluctuation is constant along the trace) indicates that an "amplification" is introduced, at those distances, to the backscatter + noise signal because of the exponential to linear conversion. Hence the "logged" trace of Fig. 5.4 can be quantized more efficiently than the exponential trace of Fig. 5.2, because the quantization steps are equally distributed. It can be seen that the "logging" operation tends to decrease the signal dynamic range by suppressing the high level signal values and by boosting the low level signal values. The "logged" signal fits more efficiently into the A/D input dynamic range.

The logarithm of the second range signal appears in Fig. 5.5. The 10 to 19.5 km backscatter is completely buried in the noise even after the exponential to linear conversion. Only the far-end

Fig. 5.3: Experimental OTDR display of the backscatter from 10 to 19.5 km amplified by 6 dB over the first range signal (without averaging).

Fig. 5.4: Experimental OTDR display of the logarithm of the 0 to 10 km backscattered signal (without averaging).

Fig. 5.5: Experimental OTDR display of the logarithm of the 10 to 20 km backscattered signal (without averaging).

Fig. 5.6: Experimental OTDR display of the averaged trace of the 0 to 10 km backscatter after 20 sec. averaging.

reflection appears at 19.5 km. A dc offset equal to the noise fluctuation is added to both "logged" traces before their application to the A/D input in order to keep them within the A/D quantization limits ($0 \rightarrow 1V$).

The averaged trace of the first range signal, after 20 sec averaging, is shown in Fig. 5.6 and, after 1 minute averaging (two overflows had occurred), is shown in Fig. 5.7. The three high-loss splices at 2, 4 and 6 km, which also appear on the non-averaged trace of Fig. 5.4, are very clearly shown. A fourth, very low-loss splice appears at about 8 km, since the 20 sec averaging is enough for the recovery of the 0 to 10 km backscatter signal from noise.

The averaged trace of the second range signal after 3.5 minutes averaging (one overflow had occurred) is shown in Fig. 5.8. Four splices are identified by arrows at 10.3, 13, 14 and 15.5 km. The slope of the averaged trace decreases as the distance increases beyond 15 km and finally becomes very small because of the distortion introduced by the dc offset before logging.

Figure 5.9 shows that a more distant splice can be identified at about 18 km (indicated by the arrow), but not very clearly, with an averaging time of 5.5 minutes (four overflows had occurred). The decrease in slope of the portions of the "chopped" waveform of Fig. 5.9 is due to the dc offset distortion.

Fig. 5.7: Experimental OTDR display of the averaged trace of the 0 to 10 km backscatter after 1 min. averaging (two overflows had occurred).

Fig. 5.8: Experimental OTDR display of the averaged trace of the 10 to 19.5 km backscatter after 3.5 min. averaging (one overflow had occurred). The four splices at 10.3, 13, 14 and 15.5 km are indicated by arrows.

Fig. 5.9: Experimental OTDR display of the averaged trace of the 10 to 19.5 km backscatter after 5.5 min. of averaging (four overflows had occurred). The 18 km splice is indicated.

With proper scaling Figures 5.6, 5.7 and 5.8 allow the measurement of the fiber link loss and the splice losses. The resolution of the OTDR for the splice location is approximately 100 m since the launched probe pulse is 1 μ s wide.

Figures 5.10 and 5.11 display the same link signature when light is launched from the other end at 19.5 km. Figure 5.10 shows the 19.5 to 9.5 km range, and Figure 5.11 shows the 9.5 to 0 km range. A correlation of Fig. 5.6 with 5.11 and of Fig. 5.8 or 5.9 and 5.10, ensures the accurate location of the splices. However, it is entirely possible that splices which appear on the signature taken from one end can appear with lower loss or even disappear on the signature taken from the other end. This could occur when two fibers having different NA are spliced together so that the splice loss for one direction of transmission is different from the loss in the opposite direction.

It can be concluded that this experimental OTDR is able to detect non-reflective faults such as arc-fusion splices at distances 18 km from the transmitting end. The logging of the backscatter signal before digitization extends the OTDR range to 18 km. If the backscatter exponential trace is directly digitized without introducing the logarithmic conversion then the OTDR range is approximately limited to 14-15 km.

Fig. 5.10: Experimental OTDR display of the averaged trace of the 19.5 to 9.5 km backscatter when light is launched from the 19.5 km end. The averaging time is 1 min. (two overflows).

Fig. 5.11: Experimental OTDR display of the averaged trace of the 9.5 km to 0 backscatter when light is launched from the 19.5 km end. The averaging time is 3.5 min. (one overflow).

CONCLUSIONS AND RECOMMENDATIONS

An OTDR system for testing multimode fibers at 1.3 μm light wavelength has been developed and tested. This system employs a digital averaging technique which provides a SNR improvement of more than 50 dB and consequently extends the range of fault location to 18 km (about 20 dB one-way loss).

A long wavelength semiconductor laser diode was used as the pulsed light source. Because these laser diodes have relatively limited peak emitted optical power, the fiber backscatter is very weak, therefore very low-noise detection and amplification techniques are required in order to have a large OTDR dynamic range. A pin-FET (hybrid) front-end was used at the receiver which has a very low noise performance with a total noise current density of 0.7 pA/ $\sqrt{\text{Hz}}$. Low-pass filtering (3 dB at 4 MHz) was also introduced to limit the noise. Since the backscatter SNRs were still very poor for distances beyond 6 km, digital signal averaging was used for SNR enhancement. The logarithmic conversion process which was introduced before digitization was experimentally proven to be very important for the recovery of the backscatter signal from noise even though it introduces a small amount of distortion within the OTDR dynamic range. This distortion will be eliminated in a final system.

The experimental averaging system, as reported in Chapter 5 Fig. 5.9, performed averaging longer than 3.5 min. and more SNR improvement was obtained even if, based on [30], SNR deterioration is to be expected. The accumulation of more samples in this averaging process did improve the signal beyond the commonly accepted limit. The reasons for this improvement are: 1) The backscatter signal even if smaller than the A/D quantization step (for distances beyond 13 km) covers all the A/D input dynamic range because of the dc offset and the noise fluctuations; 2) while the noise fluctuations make it possible for the low-signal values to be efficiently quantized, they cancel out because of averaging; 3) the overflows allow the averaging to continue beyond the memory limits. The continuous adding of samples "amplifies" the very low backscatter signal to a value larger than the A/D quantization step, therefore it increases the resolution of the averaged trace and improves the SNR. Further study is needed to clarify these points fully, but this is beyond the scope of the present study.

Finally the logarithmic conversion process together with the "overflow detecting" averaging process makes the OTDR capable of detecting splices at 18 km (about 20 dB one-way loss).

Several improvements can be made in the experimental OTDR system in order to make the system more efficient and easier to operate. The following recommendations are made

- 1) A semiconductor laser with a higher output power should be used to extend the OTDR range and reduce distortion due to dc offset when the log-amp is used.
- 2) A microprocessor is required for the reconstruction of the "chopped" averaged waveform and for other functions. Thus an averager ~~memory-microprocessor memory~~ interface circuit has to be developed for the averager memory control. Also the start-stop averaging controls may be supplied by the microprocessor.
- 3) The OTDR display should be calibrated in order to be able to read the link loss and the splice losses from the averaged fiber signature. Appropriate microprocessor software is also required for more accurate measurement of the splice loss and its location.
- 4) Further experimental work should be done to determine whether it is better to perform the logarithmic conversion before or after digitization of the backscatter in order to reduce the dc offset distortion.
- 5) Because the pin-FET front-end noise current is due predominantly to the thermal noise of the feedback resistor $R_F=40\text{ K}\Omega$ larger values of R_F are required in order to decrease the receiver noise. This is not possible with the present experimental OTDR because the pin-FET front-end is a commercial hybrid circuit which is packaged and sealed. Therefore a pin-FET front-end with large R_F ($500\text{ K}\Omega$ to $1\text{ M}\Omega$) should be built. It should have at least a 1 MHz bandwidth since a $1\text{ }\mu\text{s}$ launched laser pulse is used.

- 6) Further study should be done on the display process in order to reproduce the averaged trace in a more efficient way using all 24 bits of the RAM instead of using only the most significant 8 bits.

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