Spread spectrum techniques for distributed multimeasurand optical fiber sensors

by

Ravikumar K. C.

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APPROVED:

Richard O. Claus
Dr. Richard O. Claus, Chairman

Dr. Ira Jacobs

Dr. Kent A. Murphy

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Spread spectrum techniques offer an alternate solution to the urgency for distributed optical fiber sensors. These techniques are based on the properties of pseudorandom sequences that have triangular autocorrelation functions with peaks only at regions of no delay. This affords an opportunity to give the desired signal a power advantage over many types of interference and noise.

A study in employing spread spectrum techniques for multiplexing optical fiber sensors is presented. A mathematical analysis of the system is conducted with due consideration given to performance issues. Simulations in software are conducted to characterize system performance. Hardware developed for this project operates at over 1 Mbps and is capable of simultaneously monitoring four sensors. Real time experiments conducted on these multiplexed sensors affirm the technical feasibility of the system. Configurations for viable applications of the system are also suggested.
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1.0 Introduction

Distributed sensors offer a practical solution to the increasing demand for monitoring large complicated structures. An object of current investigation is to manage several passive optical fiber sensors from a common optoelectronic source and receiver module. Many applications of fiber sensors like industrial systems, process control applications and smart structures require multi-sensor systems. Multiplexing of fiber sensors is an efficient and elegant way to do this. Multiplexing reduces the cost of the system by reducing the number of lasers and the number of detectors to be used. Passive multiplexing eliminates electrically active components in the immediate sensing environment.

Multiplexed sensors may be defined as those designed to collect information from a number of discrete sensing points, or regions. The determination of the spatial position of the sensor so that the detecting system can determine which sensor is providing the information is the essence of these investigations. The multiplexing capability is limited by the bandwidth and signal-to-noise ratio of the detected signals. An important factor of concern is the crosstalk between signals from independent sensors.

The large transmission bandwidth of a fiber channel makes multiplexing fiber systems an attractive feature. A single fiber is capable of supporting a large number of sensors and makes it a natural choice for many sensing applications. In addition, fibers are very insensitive to crosstalk from other fibers. Unintentional coupling of energy between fibers is insignificant as there is no appreciable energy in the outer regions of the cladding of the fibers.
Several multiplexing techniques exist for use with fiber sensors. Most of them owe their development to communication-based multiplexing techniques. The most common ones in use are spatial (SDM), time or temporal (TDM), frequency (FDM), wavelength (WDM) or chromatic, coherence, and polarization division multiplexing techniques. All these schemes allow signals from specific sensors to be separated at the demodulator.

Spread spectrum techniques for multiplexing which is the topic of this thesis was developed to allow an alternative to the ones mentioned above with many advantages over them. Correlation based-multiplexing techniques have been around for some time now, however their use has been limited mainly to communication systems. It has been used in the context of TDM with some success [7].

Spread spectrum schemes achieve multi-sensing capability by assigning a distinct signature waveform to each sensor in the system from a set of code sequences that have low correlations with one another. For this reason it is also known as Code Division Multiplexing (CDM)[9]. Here this idea is implemented using the favorable properties of the autocorrelation function for code sequences known as maximal length sequences. Maximal length sequences, discussed in more detail in Chapter 2, have a triangular autocorrelation function with peaks to this function occurring at regions of no delay. The detector receives the sum of all the sensor modulated code sequences. The detection scheme employed ensures that the sensor information is recovered by correlating the received signal with replicas of the assigned code waveform or signature sequence.

Correlation techniques have the advantage of being able to perform satisfactorily even under limited source power by improving the mean launched power, not the peak power as in many other multiplexing techniques, especially TDM. However this is more evident
in digital systems. They use high bit-rate pseudo-random sequences, hence, higher resolution is effected.

Many channels can operate simultaneously over a common frequency band. The common bandwidth required is ideally a few orders of magnitude above the sensor information bandwidth. This is mandated by the frequency of the analog sensor signal to be multiplexed.

The spread spectrum based system is scalable and modular. This characteristic of the ability to add sensors to the system with marginal increase in crosstalk is called soft capacity. Adding sensors to an existing network steadily increases the background noise. The limit to this adding on of sensors is dependent on the code sequence characteristics. It is interference limited, which is to say that the number of active sensors or the sensors which are contributing signals to the common channel limits the ability of the system to multiplex. If the activity cycle of the activated sensors in the system is low the chances are that the crosstalk is low. In applications where this cycle is low, this technique can be very effective.

Spread spectrum systems make efficient use of the spectrum as they are capable of providing asynchronous access to every sensor, though better performance is effected with synchronization. Moreover no guard band, as in FDM, and guard time, as in TDM, is necessary between the sensors. This represents saved bandwidth.

Chapter 3 discusses global design concerns pertaining to the implementation of the multiplexing scheme. A theoretical analysis is presented in chapter 4.0. Chapter 5.0 is a description of the hardware implementation with the results and chapter 6.0 discusses the software simulations conducted.

1.0 Introduction
From the applications point of view this technique can be a very convenient and marketable commodity. Optical fibers can be used simultaneously both as a communications and sensing medium. A method is described here in Chapter 7, where spread spectrum techniques can be applied in this area. Sensor networks are necessary to effectively develop smart materials. Chapter 7 also proposes a configuration on how this technique can be adapted to smart structure applications.
2.0 Multiplexing Options for Sensors

2.1.0 Conventional Techniques

A brief discussion of the more popular multiplexing techniques is carried out here. The advantages and disadvantages of the configurations are also discussed. This is followed with a detailed description of the spread spectrum technique.

2.1.1 Spatial Multiplexing (SDM)

This is a simple and straightforward approach. Separate fibers are used to interface with individual sensors. Each sensor interacts with a detector. SDM can be configured with a common light source and multiple detectors or with multiple sources and a common detector. The light source may be a common one with a star coupler distributing the power into several sensors or all the sensor outputs could be combined into a common detector by a star coupler. In order to use a single source and single detector synchronous switching may have to be employed. This involves successive connecting and disconnecting of the detector and source from the optical network. The operation of these switches on a time sharing basis makes it similar to TDM that is discussed next.

The crosstalk of this implementation is very good. There is little or no interaction of the detected signals with each other. The power budget is excellent and major losses are due to coupling and fiber losses. However it requires excessive hardware and is not flexible.
2.1.2 Time Division Multiplexing (TDM)

The sensors are temporally separated from each other with regard to the light source. This is achieved by arranging differing total optical propagation delays for the signals from each sensing signal. The light source also needs to be modulated by temporally spaced signals.

In the simplest configuration for TDM a repetitive short pulse of duration, such that the returning pulses from the sensors do not coincide at the detector, is used for modulation. The repetition rate is low enough to allow a pulse to return from the farthest sensor before the next pulse returns from the closest sensor.

This technique may be used in both coherent and incoherent systems. It requires fast electronics if the optical path differences are to be made short.

The limitations of this technique are that the maximum optical peak power of the pulses limits the dynamic range of the sensors. This can be rectified by using 'chirp' signaling. It requires fast electronics.

2.1.3 Wavelength Division Multiplexing (WDM)

For wavelength division multiplexing each sensor is assigned a particular wavelength with a well-defined spectral width. The parameter to be measured is detected by evaluating modulation at this wavelength. It normally uses many light sources of narrow spectral width to access individual sensors. In another implementation a WDM device is used in conjunction with a wide light source. The WDM device disperses the light into narrow wavelength components and couples them into the fibers leading into the sensors. The sensor signals are coupled into a single fiber and a WDM demultiplexer is used to demodulate the sensor outputs. The WDM device may be a grating.
This scheme has good optical budget performance and requires low speed electronics. The demerits are that multiple sources or detectors may be required. It is a problem to use this scheme with more than 10-20 sensors.

2.1.4 Frequency Division Multiplexing (FDM)

In this method the transmitted signal is modulated by electrical signals which contain as many distinct frequency components as the number of sensors in the network. The optical path through each sensor is made different by design to achieve unique optical delays. The output signal may be combined and received at a common detector. The output is the sum of phase-shifted signals containing the sensor signals. As the phase is determined by the frequency and the optical delay through individual paths, they can be demodulated by solving the resulting \( N \) simultaneous equations where \( N \) is the number of sensors in the network. The detected signal can be demultiplexed by frequency selective filtering.

One elegant method to do FDM is realized by sub-carrier frequency division multiplexing. Here the resultant of many sub-carrier modulation signals, dependent on the relative phase angles of the sensors is received at the detector and then the signals are separated by employing multi-channel phase sensitive detection and subsequently solving a set of simple simultaneous equations. Serious crosstalk can result here if the electronic and optical delays are unstable.

Another method is using frequency-modulated carrier wave multiplexing. The transmitted signal here is an optical carrier wave the frequency of which varies linearly for a period \( T \), after which it switches back to its original frequency, and repeats henceforth. When interferometers are connected with such a source with a differential path delay, the signals arriving at the detector will differ by a frequency proportional to the optical path
difference and the slew rate of the source. Various sensor signals may be isolated by filtering out the corresponding frequencies.

Some of the problems with this scheme are that it requires an unbalanced interferometer for it to operate, and this is a problem due to the finite coherence length of the source. Laser nonlinearities in the frequency modulation ramp make it difficult to extract exact phase information from the system.

Among other multiplexing methods, there are coherence and polarization mode multiplexing [2,7]. By employing two sensor interferometers of optical path difference (OPD) much more than the coherence length of the source, the interference associated with the sensors can be reconstructed by tuning the its OPD to match that of each sensor. This suffered a degradation in performance when more than two sensors were multiplexed.
Figure 2.1 Multiplexing formats
2.2.0 Code Division Multiplexing

2.2.1 Spread Spectrum Techniques

In recent years spread spectrum techniques for communications have proven to be a viable alternative to conventional multiplexing schemes with many advantages over them. Spread spectrum techniques rely on the pseudo-random properties of code sequences. The sensor signal is spread over a wide frequency range thus creating a noise like signal.

Spread spectrum methods afford an opportunity to give a desired signal a power advantage over many types of interference, even intentional noise. This power advantage is often proportional to the ratio of the dimensionality of the space of code sequences to that of the data signal.

The capacity of spread spectrum techniques is expressed by Shannon [5] in the form:

\[ C = W \times \log_2 \left( 1 + \frac{S}{N} \right) \]  

For \( \frac{S}{N} \leq 0.1 \)

\[ \frac{C}{W} = 1.44 \times \frac{S}{N} \]  

where \( C \) is the channel capacity, \( W \) is the bandwidth in Hertz,

\( N \) is the noise power, and \( S \) is the signal power.

The most often used code sequences are Maximal sequences (m-sequences), Gold sequences, and Kasami sequences. The type of code sequences employed, their length,
and its bit rate set bounds on the capability of the system. Therefore it is of paramount importance to study the properties of the sequences themselves.

Maximal sequences are the longest pseudo random codes that can be generated by a shift register of a given length. The properties of these sequences largely determine the performance of any system which rely on them. Dixon [8] gives a comprehensive treatment of these properties. They are listed below in detail, considering that it is the sequence that is employed in this thesis.

1. The number of ones in a m-sequence is one more than the number of zeros in the sequence.
2. The modulo-2 sum of an m-sequence and any phase shifted replica of the same sequence is another phase of the same m-sequence different from either of the originals (shift and add property).
3. The statistical distribution of ones and zeros is well defined and always the same.
4. Every possible state of an n-stage generator exists at some time exactly once during the generation of the cycle. An all zero state is however not allowed to occur.
5. The autocorrelation function is such that for all values of phase shift, the correlator value is -1, except for the 0 + 1 bit phase shift in which the correlation varies linearly from the -1 value to $2^{n-1}$ (sequence length).

The autocorrelation function of any sequence is defined by

$$R(\tau) = \frac{1}{T} \int_{0}^{T} c(t)c(t-\tau) dt$$  \hspace{1cm} (2.3)

Where $c(t)$ is the code sequence
For a maximal length sequence of length $N$ (generated by an $n$-bit shift register), due to its unique properties mentioned above, this function is two-valued and is given by

$$
N = \begin{cases} 
2^n - 1, & \tau = 0 \\
-1, & \tau \neq 0;
\end{cases}
$$

Figure 2.2 shows the autocorrelation function for a 127 bit long code sequence. It can be seen from the figure that for all values of delay but that of the undelayed and that with a delay of an integral multiple of the code length the function has a very low value. The autocorrelation function peaks at zero shifts and it increases linearly between the zero and plus and minus one bit shifts. Codes other than maximal length sequences have autocorrelational properties which are markedly different from those of the maximal ones. For these, as can be seen from Figure 2.2b, partial correlations of the phase-shifted replicas of the code itself give rise to minor peaks in the autocorrelation function. The functions in the figures are software generated from actual code sequences.

The difference between the region of maximum correlation (peak) and the next highest point of correlation is known as the index of correlation and gives a measure of the crosstalk the system can afford when the code is used. It can be seen that the index of discrimination is higher in the code for Figure 2.2a, than for Figure 2.2b.
Figure 2.2a. Autocorrelation function of a 127 bit m-sequence

Figure 2.2b. Autocorrelation function of a 127 bit non-maximal-sequence
2.2.1.1 M-Sequence Generator

A linear code sequence generator can be made up of a set of flip-flops connected as in a shift register. A combinational feeding back of the output and any state or states of the individual flip-flops leads to a number of states in the shift register which is a function of the length of the register and the combination of the feedback.

The generalized implementation of the generator is shown in Figure 2.3. Only certain combinational feedbacks in an $n$-length shift register generate a $2^n-1$ length pseudo random sequence known as the maximal length sequence. The feedback taps which achieve this are available in established tables published in many texts. Setting the taps as per these combinations results in cyclic m-sequences being generated by them.

![Diagram of M-Sequence Generator](image-url)

**Figure 2.3. Linear code sequence generator**
3.0 Global Design Considerations

This chapter considers major aspects in the design of spread spectrum systems. The modulation considerations, fiber sensors, noise considerations and receiver requirements are discussed.

3.1.0 Modulation

In order to transmit signals over the fiber it is necessary to modulate some property of the light with the information signal. This property maybe intensity, phase, frequency, or polarization, with either digital or analog modulating signals.

The terms AM, FM, PM, PolM, etc. are used in context with analog modulation, to represent amplitude, frequency, phase, and polarization, respectively. In digital terms it is common to use the term SK to mean Shift Keying, as in ASK, FSK, PSK and PolSK. In these schemes the values of either amplitude, frequency, phase or polarization switches back and forth between two values or more depending on the order of the signal space.

Intensity modulation (ASK) is the simplest of these to implement and is also the most often used. In this application we have chosen to intensity modulate the source. ASK is sometimes called on-off keying, though ASK is the preferred usage. The light source is switched between the on and off state. In the off state, the source is not usually turned all the way off due to certain implementation difficulties. What is important is the extinction ratio \( r \) defined as

\[
r = \frac{P_0}{P_1}
\]  \hspace{1cm} (3.1)
should be very low. Here $P_0$ is the OFF state amplitude and $P_1$ is the amplitude in the ON state.

AM or ASK for optical systems can be implemented in two ways, direct modulation or external modulation. Electro-luminescent sources like the laser diode can be directly modulated by varying the injection current with the desired waveform. However high speed on-off keying introduces many an undesirable transient phenomena in the laser. Some of the problems encountered are discussed here. Delay in turning on is caused if the diode is biased too low. If the current goes below the lasing threshold value and subsequently rises to a high state, there is a considerable delay in starting up the stimulated emission process. The spontaneous emission triggers the onset of lasing and hence the turn on time of the laser becomes a random variable.

Thermal effects also put a lower limit on the frequency at which the laser can be directly modulated. The on-off process is accompanied by a heating and cooling cycle which can play a detrimental role at frequencies above 10Mhz.

The random fluctuations which accompany high speed operations can cause mode hopping in the laser. Mode hopping results when any mode randomly captures the gain of the cavity for a while and then another.

At the onset of stimulated emission, carriers in the region are drawn into cavity, the light level drops and carriers have to build up sufficiently for the emission to prolong. Relaxation oscillations owe their origin to this "ping-pong" effect between the photon and carrier density. This oscillation would die out in a steady state system but exists to some extent in on-off keyed lasers. Line broadening or 'chirping' can result from oscillations in a
single longitudinal mode under CW operation and it is associated with the modulation induced changes in the charge density.

Figure 3.1. Direct ASK modulation of the laser

An alternative option, and one which is gaining popularity as a modulation technique, is external modulation. Here the laser is allowed to operate always above threshold and the drive current remains constant. The modulation is external to the source. Several approaches are possible to realize this. For intensity modulation, integrated optics modulators are available with high extinction ratios. Acoustic deflectors are commonly used in laser printers. They may be used for ASK if the bit rate is lower than 10 Mb/s.
For this research the signal was direct modulated. ASK was used although it can also classify under binary phase shift keying (BPSK) as a phase variation occurs simultaneously.

3.1.1 Sensor Modulation

The sensors used are all analog sensors, which means that the output of each of the sensors is an analog signal which intensity modulates the waveform emitted by the transmitter. Even if the sensor is a phase modulated one, this information is available as an amplitude modulation of the carrier.

The analog sensor signal is converted into a pulse-type signal where the amplitude of the pulse denotes the analog information. This results in a gated pulse amplitude modulated (PAM) waveform with natural sampling. With the use of pulses, it is to be expected that the bandwidth of the PAM signal will be wider than the original analog waveform. The sampling theorem determines the rate of sampling of the signal \( f_s \). If \( B \) is the highest frequency in the analog signal then \( f_s \) has to satisfy the condition,

\[
f_s \geq 2B,
\]

where \( 2B \) is called the Nyquist rate.

The PAM modulated sensor signal can be represented by

\[
\omega_s(t) = \omega(t)s(t),
\]

where \( \omega(t) \) is the angular frequency of the modulating sequence

and \( s(t) = \sum_{k=-\infty}^{\infty} \prod \left( \frac{t-kT_s}{\tau} \right) \)
$s(t)$ is a rectangular wave switching waveform and $f_s = 1/T_s \geq 2B$.

3.1.2 Spectrum

The power spectral density of the modulated signal which is a PAM signal, is a $[(\sin(x))/x]^2$ function with line separation equal to the repetition rate of the coding sequence, i.e. if a 127 bit code is used with a code-rate of 1Mbps, the code would repeat every $(1 \times 10^6)/127 = 7874$ times per second. This factor influences the decision on code length. It is important to have the (code bit rate)/(code length) less than the lowest frequency of interest in the sensor signal. The code repetition rate should not fall in the sensor signal passband or the code and crosstalk cross products may fall into the demodulated signal band and thereby reduce receiver output signal-to-noise ratio.
3.2.0 Optical Fiber Sensors

Optical fiber sensors have the following advantages over conventional sensing techniques. Their immunity to electromagnetic interference, immunity to electromagnetic pulse effects, high sensitivity, light weight, small size, inherent electrical/optical multiplexing capability and large bandwidth have won them acceptance in many sensing applications.

Optical fiber sensors may be broadly classified in two major categories, intensity-based sensors and interferometric sensors. For this study one of each category has been used and multiplexed simultaneously.

3.2.1 Intensity-Based Sensors

Intensity-based sensors are the among the simplest form of fiber sensors. The optical power propagating in the fiber is attenuated proportionally to the measurand. Many sensors which fall into this category are easy to build and are reliable. Their low cost is another feature which favors their use. Small linear or angular displacements can cause amplitude modulations of these sensors.

The intensity sensor employed for this study is the intrinsic micro-bend sensor. An intrinsic sensor is one in which the measurand causes an amplitude modulation of the transmission in the optical fiber in the sensing region.

Microbend Sensors: All fibers are sensitive to micro-bending. Microbend sensors work on the principle of loss in the fiber due to bending[17]. On bending the power in the core leaks into the cladding. This causes attenuation in the fiber which is a function of the bending process. Figure 3.2b shows a typical reflective mode microbend sensor.
It has been observed that bending loss is maximum when the bending is applied periodically with a bend pitch $x$ where

$$x = \frac{C \pi r n}{NA}$$

(3.4)

where

$C$ is the core refractive index profile factor

$= 1.42$ for step index fibers; and 2 for graded index fibers

$r$ - core radius

$n$ - refractive index

3.2.2 Interferometric Sensors

These sensors operate on the principle of change in the phase term due to the environment. They are also called phase modulated sensors. Mach-Zender, Michelson, Sagnac, Fabry-Perot and many other such sensors fall into this category.

The Extrinsic Fabry-Perot Interferometer

The Fabry-Perot cavity is formed between two parallel ends of fiber [4]. One of them is single mode and acts as the input-output fiber. The other may be a multi-mode fiber and its end forms a reflector to the incident light. A hollow core fiber holds the two fibers together in its hollow tube. The Fresnel reflection from the glass-air interface of the single mode fiber at the cavity interferes with the reflection from the air-glass interface at the multimode end of the cavity. This occurs at the input-output, single mode fiber. The cavity length or the air gap between the two fiber ends varies with the strain on the hollow core fiber which holds the two fibers together. Any such change causes changes in the
phase of the sensing reflection with respect to the reference reflection. This causes interference fringes at the detector. Figure 3.2c gives an exaggerated view of an Extrinsic Fabry-Perot Interferometer. The light exits the fiber at the sensing region and re-enters the fiber which makes this an extrinsic form of sensor.
Figure 3.2a. General layout of reflective mode sensors

Figure 3.2b. Reflective mode microbend sensor with exaggerated details

Figure 3.2c. Fabry-Perot sensor system. Cross section shows the sensor.

3.0 Global Design Considerations
3.3.0 Noise

Noise in fiber optic systems is caused by various phenomena. The noise may be broadly classified as additive noise and multiplicative noise. Additive noise remains in the system even in the absence of the signal, but multiplicative noise is present only when the signal is present. Multiplicative noise arises due to the randomness within the signal itself or is produced due to certain device dependent characteristics.

Dark current noise, thermal noise, amplified spontaneous emission noise, electronic amplifier noise and crosstalk which are discussed in some detail below fall into the additive category. Modal noises, laser phase and amplitude noises contribute to multiplicative noise and signal shot noise (quantum noise).

3.3.1 Transmitter Noise

Intensity Noise: The power emitted by a laser diode in steady state conditions exhibits fluctuations around an average steady state level [2,5]. The variations are essentially due to the coupling of spontaneous photons into the stimulated lasing output.

The quantity Relative Intensity Noise (RIN) is used to describe the variance of the light-output amplitude divided by the average power. It is a form of noise to signal power. The term carrier to noise ratio (CNR) is used to describe the ratio between the powers of the modulated waveform signal and noise.
\[ RIN = \frac{\langle I^2 \rangle}{\langle I \rangle^2} = \frac{\langle I^2 \rangle}{I_0^2} \]  \hspace{1cm} (3.5)

if the amplitude modulation depth is \( m \), the CNR due to the laser amplitude fluctuations only is

\[ (CNR)_{RIN} = \frac{m^2}{2RINB} \]  \hspace{1cm} (3.6)

Where \( B \) is the receiver bandwidth.

**Phase Noise**

Quantum fluctuations in the laser affect the optical phase as well. The finite line width of each longitudinal mode is due to the random frequency shift during the emission which results due to phase fluctuation. The effect of the fluctuation of the mode power distribution in the output of the laser diode is called repetition noise.

The effect of the external optical reflections backed into the laser diode may lead to large changes in the noise and coherence properties of the laser. Strong instabilities, both in the power and spectrum may occur on account of this, depending on the distance of the reflector, feedback strength etc. In order to keep this noise low it is necessary to minimize any light reflected back into the laser from splices, connectors, etc., as it causes an increase in the level of spontaneous emission. RIN may be expected to be as low as -150dB/Hz if reflections are kept below -50 to -60 dB, or as bad as -110 dB/Hz for reflections of the order of -20 dB.
The reflection into the laser is a major concern here as reflection mode sensors are used for this thesis. The response of EFPI sensors can be improved by increasing the reflectivity of the reflecting surfaces by coating the end surfaces with metal. The reflectivity of glass to air interface is of the order of four percent. This can be improved to about 90% by using coatings. Although this is advantageous from a power budget point of view, it increases the laser noise. In the experiments conducted for this thesis no coatings were used, however this issue needs to be addressed when coatings are used. A remedy to this problem is the use of optical isolators.

3.3.2 Fiber-Noise

In fiber systems there are several sources of multiplicative noise. These noises are such that when the transmitted power along the fiber is zero, there is no noise due to it, hence they are multiplicative in nature. The two main contributors to this noise in a fiber system are modal noise and mode partition noise.

Modal noise is predominant in multi-mode fibers where there are many propagating modes. Modal noise is also known as speckle noise because a familiar speckle pattern is seen when many constructive and destructive interferences are visible across an aperture many wavelengths wide. When a multimode fiber is excited by a narrowband source, even little changes in the bend curvature, pressure, temperature, reflectivity from a splice, or connector, or frequency of the laser will cause a shift in the amplitude and phase at various points in the speckle pattern across the fiber crossection. Modal noise is a problem mainly with multimode fibers and narrowband sources, and mode partition noise manifests itself in signal mode fibers with wideband sources.
Mode partition noise is the result of the interaction between the non-zero bandwidth of the laser source and the dispersion of the fiber. The various wavelength components due to a wide band source arrive time-overlapped and form constructive and destructive interference patterns at the output.
3.3.3 Receiver Noise

The incident optical power is converted into electrical signals at the photodiode. The current in the receiver is given as

\[ I_p = R P_{in} \]

\[ I_p \] is the photocurrent;

\[ P_{in} \] is the incident optical power at the receiver

\[ R \] is the responsivity of the receiver

However a practical photodiode is not free from noise. The two main sources of noise at the photodiode are shot noise and thermal noise.

**Shot noise:** Shot noise is primarily due to two random current processes in the material of the photodiode. Electrical current consists of a stream of electrons which are randomly generated. These have been investigated by Schottky [2,6] and is given as

\[ I(t) = I_p + i_s(t), \]

where \( i_s(t) \) is the current fluctuation due to the shot noise. Here \( i_s(t) \) is a stationary random process with Poisson statistics which can be approximated by Gaussian statistics. They are dark current noise (*due to* \( I_D \)) which is typically in the range of a few nanoamperes, and quantum noise (*due to* \( I_p \))

\[ \sigma_s^2 = q(I_p + I_D) \Delta f \]

\( \Delta f \) is the Bandwidth of the receiver

and \( q \) is the charge on an electron.

The detection of light by a photodiode is a discrete process at the electron-hole level. An electron-hole pair results from the absorption of a photon and causes a current to flow.
**Thermal noise:** Random thermal motion of electrons in a resistor also manifests itself as a fluctuating current even in the absence of an applied voltage. Thermal noise can be included in the earlier equation, equation 3.8.

\[ I(t) = I_p + i_s(t) + i_T(t), \]

where \( i_T(t) \) is the current fluctuation due to the thermal noise. Here, \( i_T(t) \) is also modeled as a stationary Gaussian random process with a spectral density that is frequency independent up to \( f \sim 1 \text{ THz} \) (nearly white noise) and is given by

\[ S_T(f) = 2k_B T / R_L, \]

where \( k_B \) is the Boltzmann constant, \( T \) is the absolute temperature, and \( R_L \) is the load resistor. \( S_T(f) \) is the two-sided spectral density.

### 3.3.4 Correlational Receiver

For conventional processing in cellular systems, bipolar, antipodal and orthogonal signals, '+1' or '-1', are used to form the pseudo-random code. In incoherent processing '1' and '0' are used. Though it is possible in principle to do coherent optical processing, practical systems are not yet developed due to the high frequency of the optical carrier. The optimum (minimum probability of error) detector in additive white Gaussian noise (AWGN) consists of a bank of correlators or filters matched to each signal. The decision as to which signal was transmitted is determined by identifying the largest output of the correlator.

The received signal consists of added terms from each of the sensors. The receivers realize their processing gain by transforming the desired wideband signals into narrowband
and the undesired signals into wideband. The remainder of the process involves separating out the undesired signals.

The purpose of the correlator is to match a local reference signal to a desired incoming signal and thereby reproduce the embedded information bearing signal as the output. A perfect match is desired between the reference signal and the received code modulated signal. However, such an effect is not easily achievable. It is of some interest therefore to examine the effects of such mismatch on the performance.

If the desired signal is exactly matched with the receiver's reference, no noise is generated by the desired signal. The amount of noise depends on the output correlation or the degree of synchronization. When there is no synchronization (reference and the received signal are more than one bit apart) the output is all noise. Or if there are other sensors modulating delayed sequences, the output contains them.

For each \( r \) reduction in the synchronized signal due to code timing offset, there is a \( 2r \) increase in noise contribution on crosstalk, which results in a signal-to-noise function given as

\[
f(\tau) = A \left( \frac{T - \tau}{2\tau} \right), \tag{3.11}\]

Where \( A \) is the maximum amplitude of the signal;

\( T \) is the period of one bit;

and \( \tau \) is the timing offset.
The unwanted energy from other channels which appear in the desired signal is termed as crosstalk. In this application the quantification of crosstalk is an important parameter of interest. However, the crosstalk is dependent on the code sequences used. This is discussed in some more detail in Chapter 4.
4.0 Theoretical Analysis

Theoretical model:

The theoretical model is developed as in Figure 4.1. This system is used to come up with expressions for the noise in the system and most importantly, the crosstalk.

The system has a transmitter which is modulated by a pseudo-random sequence. The sequence is generated as shown in section 2.2. The transmitted waveform accesses k sensors which intensity modulate the code sequence.

Each sensor signal, $s_k$, modulates a distinctly delayed maximal length code sequence $[\alpha(t-\tau_k)]$. All the modulated code sequences are summed with additive white noise shown as $n(t)$ and presented to the detector. The detected signal $r(t)$ is correlated with the reference maximal length code signals corresponding to the desired sensor. The output of this receiver is the signal $Z$ which contains the desired signal.

In order to realize simultaneous outputs from all the sensors, a bank of correlators with their respective correlating reference signals is connected to the output of the common detector.
Figure 4.1 Theoretical model used to analyze the principle of operation of the spread spectrum based optical fiber sensor multiplexing system.
At the transmitter the laser is modulated by a spreading code. This code forms a carrier for the sensor signal. Let the spreading sequence be given as

\[ a(t) = \sum_{j=-\infty}^{\infty} a_{k,j} p_{T_c}(t - jT_c). \]  

\( \{a_{k,j}\} \) is the \( k \)th pseudorandom signature sequence corresponding to the \( k \)th sensor (unipolar equiprobable random variable)

\( a_{k,j} \in \{0,1\} \) the \( j \)th chip of the \( k \)th signature sequence (alloted to the \( k \)th sensor)

\( p_{T_c} \) is a unit pulse function of duration \( T_c \)

\[ p(t) = \begin{cases} 1, & t \in 0,T_c; \\ 0, & t \not\in 0,T_c; \end{cases} \]

\[ \sqrt{2P} \]

laser signal power

\[ a(t) \]

code sequence

transmitted laser signal

Figure 4.2. Source model
The sensor signal can be expressed as a sinusoidal function of the form

\[ S_k(t) = 1 + m_k \sin(\omega_k t + \phi_k), \]  

(4.2)

where \( m_k \) is the modulation index of the \( k \) th sensor,
\( \omega_k \) is the frequency in radians of the \( k \) th sensor,
\( \phi_k \) is the relative phase of the \( k \) th sensor.

The sensor signal modifies the carrier code signal. The modified signal is given by

\[ \sqrt{2} P S_k(t) \ast a(t) \]
\[ = \sqrt{2} P \left(1 + m_k \sin(\omega_k t + \phi_k)\right) \ast \sum_{j=-\infty}^{\infty} a_{k,j} P_r(t - jT_c), \]  

(4.3)

\[ P = \text{the signal power due to the laser.} \]  

(4.4)

It is assumed for this analysis that the signal power to all the sensors is the same. This is important to reduce the effects of the 'near-far' problem. The effects of this situation are discussed later in this section.

![Sensor model diagram](image_url)

**Figure 4.3. Sensor model**
At the common channel the modified signal sequences due to the other sensors connected to the system are added. Noise added to the channel owes its origin to the factors mentioned in section 3.2. They are mainly due to the laser intensity noise, the receiver shot noise, the receiver thermal noise, and the amplifier noise. This noise is modeled by a single Additive White Gaussian Noise (AWGN) term \( n(t) \). The major contributors to this noise are additive in nature. Multiplicative noise is neglected in this analysis as it is not very significant. The received signal consists of the sum of all the modified sensor signals and the component due to noise.

\[
r(t) = n(t) + \sum_{k=1}^{N} \left(1 + m_k \sin(\omega_k t + \phi_k)\right) * \sum_{j=-\infty}^{\infty} d_{k,j} P_{r_j}(t - jT_c),
\]

The received signal may be rewritten as the sum of three terms. The desired signal, noise due to the many factors discussed in section 3.3 and the crosstalk term due to the other sensors.

\[
r(t) = s_i(t) + n(t) + \sum_{k=1}^{N} s_k(t - \tau_k),
\]

where

\[s_i(t) = \text{the desired signal},\]

\[n(t) = \text{White Gaussian noise},\]

\[\sum_{k=1, k\neq i}^{N} s_k(t - \tau_k) = \text{crosstalk term due to the other sensors}.\]
Let us observe the response of the first sensor. At the correlation receiver the first sensor signal is demodulated by correlating the received signal with the signature sequence that was used to modulate the first sensor. This signal is the undelayed code sequence. From equation 4.6, the received signal with the term depicting the desired signal from the first sensor is given as

\[ r(t) = s_1(t) + n(t) + \sum_{k=2}^{N} s_k(t - \tau_k), \]

(4.7)

It must be noted that although unipolar code sequences are used to modulate the transmitted signal at the laser driver, the receiver uses the antipodal code sequence for correlation. At the transmitter, this is not possible as the modulated signal is a lightwave signal and its intensity waveform cannot be made negative non-coherently. However after the detector stage this signal is purely an electrical signal and may go negative. This bipolar form of the code sequence can be represented by

\[ a(t) - [1 - a(t)], \]

(4.8)
The code sequences modifying sensors other than the first are the original code sequence delayed by an integral bit of the sequence. For this reason the code for the first is represented by \( a(t) \) instead of \( a_1(t) \) as done earlier. The other sequences are given as \( a(t-\tau_k) \) where \( \tau_k \) is the delay corresponding to the \( k \)th sensor. The first term in equation 4.8 represents the unipolar code sequence, the second term, \(-[1-a(t)]\) is also unipolar but the zero's of \( a(t) \) are -1's and the 1's of \( a(t) \) are zero's here. Together the expression represents a antipodal code sequence.

The reference sequence has to be delayed by a time \( \tau_d \) equal to the round-trip delay of the signal through the system without any intentional delays. For simplicity in analysis this term is ignored here and assumed equal to zero. Incorporating this in the expression for the reference sequence will also yield identical results.
The received signal on correlation undergoes a multiplication operation given by

\[ Z = \int_{0}^{NT} r(t) \cdot [a(t) - [1 - a(t)]] \, dt, \]

\[ = A + \xi + \sum_{k=1, k \neq i}^{N} I_k, \]

(4.9)

where

(4.10)

\( A = \) component of output statistic due to the desired signal,

\( \xi = \) component of output statistic due to noise,

\( I_k = \) component of output statistic due to the kth sensor.

![Receiver model diagram](image)

**Figure 4.5. Receiver model**
Expression for the desired sensor signal

\[
A = \int_0^{NT} S_1(t) \ast [a(t) - [1 - a(t)]] dt \tag{4.11}
\]

\[
= \int_0^{NT} \sqrt{2P} S_1(t) \ast a(t) \ast [a(t) - [1 - a(t)]] dt \tag{4.12}
\]

\[
= \sqrt{2P} \int_0^{NT} S_1(t) \ast a(t) \ast a(t) dt - \sqrt{2P} \int_0^{NT} S_1(t) \ast a(t) \ast [1 - a(t)] dt \tag{4.13}
\]

\[
= \sqrt{2P} \int_0^{NT} S_1(t) \ast a^2(t) dt - \sqrt{2P} \int_0^{NT} S_1(t) \ast [a(t) - a^2(t)] dt. \tag{4.14}
\]

when \(a(t) = 0, \ [-1 - a(t)] = -1\) and when \(a(t)=1, \ [-1 - a(t)] = 0; \ a(t) \in \{0,1\} \) and \(-[1 - a(t)] \in \{-1,0\}\). From this it can be seen that \(a^2(t) = a(t)\).

Therefore the second term in equation 4.14 goes to zero. Using the autocorrelation function for maximal length sequences [equation 2.3], the first term in equation 4.14, can be simplified further. The waveform \(S_1(t)\) is slowly varying compared to \(a(t)\), therefore in the interval \([0, NT]\) its value is a constant.

\[
\int_0^{NT} S_1(t) \ast a^2(t) dt = S_1(t) \int_0^{NT} a^2(t) dt \tag{4.15}
\]

\[
= \frac{1}{2} S_1(t)
\]

The power is half as the duty cycle of the code carrier is 50%. 

---

4.0 Theoretical Analysis
Expression for noise

Noise is modeled as a Gaussian random process, \( n(t) \), hence any linear operation on it results in an Gaussian expression. Therefore \( \xi \) is a Gaussian random variable. An expression for its mean and variance is derived here.

\[
\xi = \int_0^T n(t)a(t) dt 
\]

(4.16)

The expected value of \( \xi \) is the mean of the expression

\[
E[\xi] = E\left[ \int_0^T n(t)a(t) dt \right]
\]

(4.17)

\[
= \int_0^T E[n(t)] a(t) dt
\]

(4.18)

\[
= 0; \text{ since the mean of } n(t) \text{ is zero}
\]
The variance is given by

\[
E[\varepsilon^2] = E\left[\int_0^T n(t)a(t)\,dt \int_0^T n(s)a(s)\,ds\right]
\]

(4.19)

\[
= E\left[\int_0^T \int_0^T n(t)n(s)a(t)a(s)\,ds\,dt\right]
\]

(4.20)

\[
= \int \int E[n(t)n(s)a(t)a(s)]\,ds\,dt
\]

(4.21)

\[
= \int \int \frac{N_0}{2} \delta(t-s)a(t)a(s)\,ds\,dt
\]

(4.22)

\[
= \frac{N_0}{2} \int a(t)a(t)\,dt
\]

(4.23)

\[
= \frac{N_0}{2} T, \quad (a(t)*a(t) = 1)
\]

(4.24)

where \(N_0\) is the noise power spectral density.
Expression for crosstalk

\[ C_k = \int_0^T s_k(t - \tau_k) a(t) dt \]  \hspace{1cm} (4.25)

\[ = \int_0^T \sqrt{2} P s_k(t - \tau_k) a(t - \tau_k) a(t) dt \]  \hspace{1cm} (4.26)

\[ = \sqrt{2} P \int_0^T s_k(t - \tau_k) a(t - \tau_k) a(t) dt \]  \hspace{1cm} (4.27)

\[ = \sqrt{2} P \int_0^T a(t - \tau_k) a(t) dt \int_0^T s_k(t - \tau_k) dt, \]  \hspace{1cm} (4.38)

The crosstalk term depends strongly on the correlational properties of the code sequence employed as mentioned in section 3.3.4. If the sequences used are maximal length sequences, then ideally as given in equation 2.2.3a, the autocorrelation function is two valued. For m-sequences this value is N when there is perfect correlation and -1 for all others. This gives a crosstalk ratio of 1/N. The ratio is the index of discrimination of the autocorrelation function of an m-sequence. The quantity gives an amplitude related crosstalk figure.

\[ N = 2^n - 1; \text{ therefore crosstalk } = -20 \log_{10} [2^n - 1] dB. \]

Under the conditions that all the sensor signals are of comparable magnitude, this expression is approximately equal to zero for all non zero values for \( \tau_k \),

\[ \text{as } \int_0^T a(t - \tau_k) a(t) dt \equiv 0, \]

The effective signal-to-noise ratio (SNR) is the ratio of the square of peak amplitude of the autocorrelation function \( N=P^2 \) to the variance of the amplitude of the cross correlation
for the k-1 uncorrelated interfering sensors. The SNR increases in proportion to the increase in bit rate of the code sequence, with respect to the sensor signal bandwidth.

Unequal Signal Power

So far in the analysis equal power to the sensors has been assumed. If the sensors are modulated by unequal powers then there is the danger of what is known in communications jargon as the "near-far" problem. In the above analysis the signal power from the source as seen by the sensors is assumed to be same (P) for all sensors.

Here we briefly analyze the situation when this not the case.

Let the powers be $P_1, P_2, \ldots, P_n$ for the n sensors.

We will now modify equation 4.25 which derives an expression for crosstalk with new values $P_1, P_2, \ldots, P_n$:

$$C_k = \int_0^T s_k(t-\tau_k)a(t)dt$$

$$= \int_0^T [\sqrt{2P_2}s_k(t-\tau_k)a(t-\tau_k)a(t) + \sqrt{2P_3}s_k(t-\tau_k)a(t-\tau_k)a(t) + \ldots + \sqrt{2P_n}s_k(t-\tau_k)a(t-\tau_k)a(t)]dt$$

$$= \left[\sqrt{2P_2} + \sqrt{2P_3} + \ldots + \sqrt{2P_n}\right] \int_0^T s_k(t-\tau_k)a(t-\tau_k)a(t)dt$$

(4.39)

As can be seen from the above expression, the crosstalk can become significant if the power in any one or more of the signals is appreciably high.
The system performance suffers when the received powers are unequal because even with good (quasi-orthogonal) signal constellations, the output of each correlator contains a spurious component which is linear in amplitude with respect to the other sensing elements in the network. As the number of sensors in the system increases, the signal-to-noise ratio at the receiver deteriorates.
5.0 System Design and Experiments

Figure 5.1 gives a block diagram of the system developed to implement spread spectrum techniques for optical fiber sensor multiplexing. However, this configuration could not be implemented directly. Prior to the actual system implementation, a low-speed model was developed. The low-speed model is used to demonstrate spread spectrum techniques very broadly.

![Block Diagram of spread spectrum multiplexing scheme.](image)

Figure 5.1. Block Diagram of spread spectrum multiplexing scheme.

Among the difficulties involved with high-frequency operation is the interfering noise picked up from the surroundings. Breadboard-wired circuits are very prone to be
Among the difficulties involved with high frequency operation is the interfering noise picked up from the surroundings. Breadboard-wired circuits are very prone to be influenced by the surroundings when operated at high frequencies. The wire leads of the components act as antennas to ambient radiation. The low frequency model avoids the complexities which arise from operating at higher frequencies. This preliminary model deals with the need for fast receivers and laser diodes. The low speed model is discussed briefly as it helped lay the foundations for the actual implementation discussed in section 5.1.2.

5.1.0 Low Speed Operation

At low frequencies the optical delay to delay the sequence by even one bit is very large. The length of fiber typically needed to delay a kilohertz waveform by a period is calculated here. The speed of light in the fiber is a function of the refractive index of silica. The speed of light in air = $3 \times 10^8$ meters/sec and the refractive index of silica is 1.45. Therefore the speed of light through silica is $3 \times 10^8/1.45 = 2.069 \times 10^2$ m/s. Operating at 1kHz implies that the pulse width or bit period is one milli-second. To obtain a delay of even one bit or pulse means a fiber length of about 200 km [$2.069 \times 10^8$ m/s divided by $10^{-3}$s].

To get over this problem, the low frequency model uses two lasers instead of delays. Each laser provides optical power to a sensor. One of them is modulated with the original pseudo-random bit sequence and the other by a delayed version of the same sequence. The delay is exactly an integral multiple of a bit.
Figure 5.2 gives the layout of low frequency model. Seven stages of an eight bit parallel-out serial shift register are used in conjunction with XOR gates to develop a 127 bit m-sequence generator (maximal sequence generator). A clock frequency of 1kHz was chosen for the pseudo random bit sequence (PRBS) generator. The output (PRBS) is buffered and used to modulate a Seastar laser (L1) which operates at 1300nm. A delayed version of the same sequence (PRBS delayed by two pulse widths exactly) is used to modulate another identical laser, (L2). The outputs of the lasers are launched into interferometric sensors (Extrinsic Fabry-Perot Interferometers) connected as in Figure 5.2, (S1) and (S2) respectively. So now we have S1 powered by L1 and modulated by the PRBS (PN1); and S2 powered by L2 and modulated by the delayed PRBS (PN2). The outputs of each of the sensors are monitored on the oscilloscopes O1 and O2. The outputs of the sensors are optically coupled using a 2x1 coupler. The coupled output is detected by a detector which converts the optical signal to an electrical signal E. This signal is correlated with PRBS (PN1) to obtain the output of sensor S1. This process is the correlation stage. Similarly this signal is mixed with the PRBS (PN2) to get the output of sensor S2. The outputs at this stage still contains the 1Kb/s bit sequence and the spread undesired signal. This is removed from the output by passing it through a low pass filter. The receiver is discussed in more detail later in section 5.2.3.

The system described gave an opportunity to monitor the sensor signal along with the demultiplexed sensor signal from the correlator. The response observed was very satisfactory and provided valuable information about the operation of the system. It was noticed that the performance of the system (signal to noise ratio) deteriorated when the laser was operated not close to its threshold. It was also seen that the performance improved with an increase in bit rate. These and several other observations provided a thorough insight into the operation of the system. This paved the way for the development of the actual model discussed in the next section.
Figure 5.2. Layout for demonstrating multiplexing by spread spectrum methods at low speeds.
5.2.0 High Speed Implementation

The real aim of multi-sensor networks is to use minimum hardware and yet at the same time access the information from a maximum number of sensors. The model described in this section achieves this goal. Unlike in the earlier model, this model uses delays. The length of the delays is inversely proportional to the frequency of operation (bit rate of the code sequence). The bit rate of the code sequence should be several orders of magnitude greater than the highest frequency component of the detected signal for effective recovery of the sensed signal through the demodulator. The system model is described in Figure 5.3.

5.2.1 Transmitter Design

System operating wavelength was chosen at 1300nm. The source employed is a BT&D laser Transmitter: XMT3150-PT. This is ideal for single-mode fiber-optic systems. The laser is equipped with a built-in modulation driver which can be modulated using a serial input bit stream. The laser bias driver supplies bias current to the laser which is nominally equal to the laser's threshold current. A laser bias monitor provides an analog output voltage which is proportional to the laser bias current.

The pseudo-random bit code is generated by a shift register made of digital flip-flops having a linear combinational feedback. The design is identical to the description of the pseudo-random sequence generator given in section 5.1.0 and 2.2.1.0.

The delayed signal generator circuitry consists of another shift register and an electronic delay. In order to delay the sequence by a fraction of a bit, the delay circuit delays the clock signal to the second shift register. An external electronic delay is set to compensate
for the optical round trip delay of the sequence. This delay is a necessity considering that we desire maximum autocorrelation with the transmitted sequence. This is a means for synchronizing the reference code sequence with the transmitted signal. The flip-flops comprising the shift register have serial and parallel data out pins. Delayed versions of the signature sequences are available at the several parallel taps of the shift register.

5.2.2 Optical layout design

The couplers employed for the implementation are single mode 2 by 2, 3 dB couplers with excess loss of about 0.3 dB. The sensors are separated by optical delays of a bit of the signature sequence. As the frequency of operation is 2Mb/s, the optical path to be traversed for the realization of a bit delay is

\[ D = \frac{\text{speed of light in the fiber}}{\text{bitrate}} \]

\[ = \frac{\text{speed of light in free space} \times \text{refractive index of the fiber}}{2 \times 10^6 \text{bits/sec}} \]

\[ = 3 \times 10^8 \text{ m/sec} \times 1.45 \]

\[ = 2 \times 10^6 \text{ bits/sec} \]

\[ = 218 \text{ meters} \]

As the sensors in use are reflective in nature the optical transmission traverses twice through the delay and hence only half the delay length is enough to realize the same effect. The optical delays corresponding to these distances were measured with an OTDR to obtain precision lengths. Three fiber delay coils, one of 218 meters and two of 109 meters each were thus constructed.
The code sequence-modulated laser output is launched into a 2x2 coupler, C1 in Figure 5.3. One arm of the coupler is split further by another 2x2 coupler (C2). One of the output arms of this coupler leads to the 109 meter delay, while the other is linked directly to a reflective Fabry-Perot sensor, S1. The delayed arm of this coupler, C2, is connected to sensor S2. The other arm of coupler C1 is connected to the 218 meter-long delay coil. The output end of the delay coil is split in two by the coupler C3. One of the two arms of this coupler is connected to sensor 3 and the other through a 109 meter long delay to sensor 4. By this configuration the effective optical delay for the path through sensor 1, 2, 3 and 4 is 500ns, 100ns, 1500ns and 2000ns, respectively. This corresponds to bit delays of one, two, three, and four bits for sensors 1, 2, 3, and 4, respectively.

The other input arm of the coupler C1, in other words the return path, is coupled with the detector.

![Figure 5.3a. Optical system layout with four sensors multiplexed by spread spectrum techniques.](image)

5.0 System Design and Experiments
Figure 5.36 Layout with four sensors multiplexed by spread spectrum techniques.
5.2.2.1 Power Budget

An important parameter issue is the power penalty. For semiconductor lasers the OFF state power \( P_o \) depends on the bias current \( I_b \) and the threshold current \( I_{th} \). When \( I_b < I_{th} \) the power emitted during the 0 bits is due to spontaneous emission, and \( P_1 \) on state power corresponding to bit 1. When the laser is biased near or above the threshold \( P_o \) can be a significant portion of \( P \),

\[
I_1 = RP_1; \\
I_0 = RP_0; \\
\text{Receiver sensitivity } P_{rec} = \frac{(P_1 + P_0)}{2};
\]

(5.1)

Enough power has to reach the detector through the optical system to ensure acceptable performance of the signal. The receiver sensitivity puts a limit on the minimum power that can be reliably detected.

The main contributors to the loss in the system are the connectors, splices and couplers. The fiber losses are not significant at the distances in the implemented system, however, they may play a part in a real world application where the multiplexed information has to be transported over longer distances. Under these circumstances one could resort to repeaters or in line optical amplifiers like the EDFA. In reflective networks a major source for the power penalty is the reflection coefficient. Plain glass to air interfaces have a reflection coefficient of only four percent. The single largest cause for the power penalty in this configuration is the use of reflective mode sensors. This loss is inherent in such a system, as the reflection coefficient at the glass air interface is only of the order of four percent.
As is noted from equation 4.39, it is desirable to have equal powers to each sensor. The system laid out as in Figure 5.3 achieves this condition in principle. The following table gives the actual measured power at the points numbered on the figure.

<table>
<thead>
<tr>
<th>Location as in Fig 5.3</th>
<th>Power in microwatts</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>528</td>
</tr>
<tr>
<td>2</td>
<td>205</td>
</tr>
<tr>
<td>3</td>
<td>80</td>
</tr>
<tr>
<td>4</td>
<td>87</td>
</tr>
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</tr>
<tr>
<td>9</td>
<td>48</td>
</tr>
<tr>
<td>10</td>
<td>42</td>
</tr>
<tr>
<td>11</td>
<td>480 nw</td>
</tr>
</tbody>
</table>

It may be noted that since the duty-cycle of the laser modulation is only about 50% (the actual ratio is $\frac{2n-1}{2n-1}$), where n is the number of stages in the shift generator used as a code generator) the power at 1 is only 528μW. The power available at the sensors are sensor 1 :: 80μW; sensor 2 :: 50μW; sensor 3 :: 85μW; and sensor 4 :: 42μW.

Table 5.2 gives the loss across different sections of the system based on the measured values shown in Table 5.1.
Table 5.2

<table>
<thead>
<tr>
<th>Locations of interest in Fig 5.3</th>
<th>Actual Loss (dB)</th>
<th>Ideal loss (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1-2</td>
<td>4.0</td>
<td>3.0</td>
</tr>
<tr>
<td>1-6</td>
<td>3.4</td>
<td>3.0</td>
</tr>
<tr>
<td>2-3</td>
<td>4.1</td>
<td>3.0</td>
</tr>
<tr>
<td>2-4</td>
<td>3.7</td>
<td>3.0</td>
</tr>
<tr>
<td>4-5</td>
<td>2.4</td>
<td>~0</td>
</tr>
<tr>
<td>6-7</td>
<td>1.6</td>
<td>~0</td>
</tr>
<tr>
<td>7-8</td>
<td>2.9</td>
<td>3.0</td>
</tr>
<tr>
<td>7-9</td>
<td>5.4</td>
<td>3.0</td>
</tr>
<tr>
<td>9-10</td>
<td>0.6</td>
<td>~0</td>
</tr>
</tbody>
</table>

The effects of dispersion are not prominent in systems operating at bit rates less than 100Mb/s.

This summarizes the optical circuitry of the system.
5.2.3 Receiver Design

The detector used is a Electro-Optical Systems, ICAE5-003-20 MHz, a detector with a built-in preamplifier of gain $10^5$. Another stage of amplification was incorporated at the output to match the impedance of the correlator. This transimpedance stage outputs a voltage signal which contains the signal $r(t)$ as expressed by equation 4.6.

The method of detection is direct (envelope) detection. At the detector, the detected signal is transformed into a current signal which in turn is converted into a voltage waveform using a transimpedance pre-amplifier. This signal is correlated with the reference sequence corresponding to the desired sensor signal. An electronic gating circuit performs the balanced synchronous detection at the photodetector output.

The reference signal to the correlator is provided by an electronic delay circuit. The delay is used to delay the code signal which modulates the laser. This delay is not to delay the signal by a bit period, but to account for the round trip optical delay of the signal through the network. It compensates for the inherent delay due to the couplers and the fiber to the sensors. This delay is the round trip delay through the first undelayed sensor as the optical path is designed with symmetrical lengths for all paths. Bit shifted versions of the original code are tapped off from consecutive parallel output points of the shift register.

The delayed reference code sequence is inverted to convert all the '1's to '0's and all the '0's to '1's. The uninverted and inverted versions of the code are used to switch the received signal to form bipolar, signals of the received signal as shown in Figure 5.4a.

The two switched signals are fed into the positive and negative inputs of a difference amplifier (Figure 5.4b.). The output of the difference amplifier is the correlated signal. This signal undergoes low-pass filtering at a fourth order Butterworth filter. The cut-off
for the filter is 500Hz. This is chosen as the frequency of operation of the sensors as the operating range of the sensors is well within that range.

A bank of four such correlators form the receiver system for the four sensors. The reference signals for the correlators are bit delayed versions of the original, with the original used as the reference for sensor 1, one bit delayed sequence for the second, two bit delayed sequence for the third and finally a three bit delayed sequence for the fourth sensor.
Figure 5.4. Schematic of correlation receiver components
5.3.0 Experiments

Experiments done in real-time

The following experiments were conducted to establish the operation of the system.

Sensor 1 as in Figure 5.3 was perturbed and the three sensors, sensor 2, sensor 3 and sensor 4, were inactive and disconnected from the system. This is done by disconnecting the connections at 5, 8 and 10 in Figure 5.3. To avoid any back reflections from the open endfaces of the fiber these ends are doused in index matching gel. This experiment does not result in zero crosstalk into sensor 1 if the others are still connected but inactive. Ideally for zero crosstalk the sensors should not only be inactive but also disconnected from the circuit. This so because there is a little crosstalk from the delayed sequences itself as given in equation 4.38. This gives a finite value even if the modulation index $m_k$ of the $k$th sensor in the expression for the sensor modulation, equation 4.2, is nonzero(i.e. there is no sensor signal is absent.

Experiment two involved observing the demultiplexed output of sensor 1 at the output of the correlator, in the absence of sensor 1 but with all the other three sensors simultaneously perturbed. Sensor 1 is disconnected by disconnecting link 3 in Figure 5.3 and applying index matching gel at that end.

Finally experiment three is conducted with all four sensors being active simultaneously and the demultiplexed output of sensor 1 is observed. This case is used to compare with the case when all sensors are simultaneously active to compare the cross talk.

Similarly these three experiments are repeated on the other three sensors. The results of these experiments are oscillographs shown in Figures 5.5 to 5.9.
Figure 5.5. The oscillographs on the left show the detected signal and the one on the right show the output of the receiver, which is the desired signal corresponding to the first sensor, sensor 1. The three cases are: a) when only sensor 1 is active and the other three are idle; b) when all the other three sensors are active and sensor 1 is idle; and c) when all the four sensors are active and sensor 1 is observed.
Figure 5.6. The oscillographs on the left show the detected signal and the one on the right show the output of the receiver, which is the desired signal corresponding to the second sensor, sensor 2. The three cases are: a) when only sensor 2 is active and the other three are idle; b) when all the other three sensors are active and sensor 2 is idle; and c) when all the four sensors are active and sensor 2 is observed.

5.0 System Design and Experiments
Figure 5.7. The oscillographs on the left show the detected signal and the one on the right show the output of the receiver, which is the desired signal corresponding to the third sensor, sensor 3. The three cases are: a) when only sensor 3 is active and the other three are idle; b) when all the other three sensors are active and sensor 3 is idle; and c) when all the four sensors are active and sensor 3 is observed.
Figure 5.8. The oscillographs on the left show the detected signal and the one on the right show the output of the receiver, which is the desired signal corresponding to the fourth sensor, sensor 4. The three cases are: a) when only sensor 4 is active and the other three are idle; b) when all the other three sensors are active and sensor 4 is idle; and c) when all the four sensors are active and sensor 4 is observed.
The oscillographs in Figure 5.5 to 5.8 demonstrate the performance of the multiplexing scheme using spread spectrum techniques. All the sensors show satisfactory performance. It is seen that the system has a high robustness under noisy conditions.

5.3.1 Discussion of Results

The experimental results are largely qualitative in nature. They confirm the technical feasibility of the spread spectrum based fiber sensor multiplexing system. The implementation gives an estimate of the performance of the system in accordance with theoretical predictions. Figures 5.5a, 5.6a, 5.7a, and 5.8a demonstrate that the sensor signal in isolation, that is, in the absence of any other sensor signal can be reconstructed by demodulating the received signal at the receiver.

The output shown in Figures 5.5b, 5.6b, 5.7b, and 5.8b is the crosstalk into sensor channels 1, 2, 3, and 4 respectively. They essentially reveal the effect of crosstalk due to other sensors in the networked system, and indicate that the crosstalk is minimal. These plots give a qualitative idea of the crosstalk when three sensors are active, simultaneously, in an array. The noise floor on these outputs is a high frequency component and could eliminated if the system electronics is made EMI safe. The proximity of the filter circuit to the high frequency section is another reason for this noise.

Figures 5.5c, 5.6c, 5.7c, and 5.8c show the output of sensors 1, 2, 3 and 4 respectively, under the influence of crosstalk from the other three sensors in the network. It can be seen that this response is as good as the response due to the sensors operating in isolation as shown in Figures 5.5a, 5.6a, 5.7a, and 5.8a.

It may be noted that despite attempts to provide equal signal power to individual sensors, points 3, 5, 8 & 10 in Figure 5.3 which are connected to the sensors do not receive equal
powers. This is a reason for the crosstalk in the sensors and is described by equation 4.39. To accommodate the unequal powers in the individual sensor signals, a compensation scheme with compensating coupler ratios can be utilised in the system. The weaker sensor signal can then have a higher share of the launched mean power by connecting it to the higher ratioed arm of the coupler. This offers a solution to what is typically called the "near-far" problem in cellular communication terminology.
6.0 Computer Simulations and Results

As the experiments conducted on the built hardware are limited by physical requirements, and did not provide qualitative results, the entire model developed was simulated in software. The simulations are based on the analysis in Chapter 4 and the design in Chapter 5. The simulations were conducted using MATLAB, on the SPARC 10 workstation.

Algorithm followed for simulations.

1. Input:
   a. length of shift register, n
   b. taps required for m-sequences, T[]
   c. bit rate of operation, R
   d. sensor information: read number of sensors, individual frequency, phase, modulation index
   e. signal-to-noise ratio, SNR

2. Generate m-sequence of length N=2^n-1

3. Generate bit-stream of bit rate R with above code sequence

4. Generate sensor functions using equation 4.2

5. Multiply the code bit-stream with the sensor function, by assigning a unique differential bit delay to each sensor.

6. Generate as many random values approximating Gaussian Noise of signal-to-noise ratio SNR as the number of samples, using the Box Mueller algorithm

7. Add (couple) all modulated sensor signals together with generated additive Gaussian white noise

8. Generate reference signals which are replicas of the original bit-stream.

9. Evaluate the expression Z [Equation 4.9]
10. Repeat 1 to 9 for various conditions

The noise model is developed from the Box-Mueller algorithm for generation of Gaussian random variables.

The assumptions made in this model are listed below:

- Fiber losses are negligible
- Multiplicative noise is negligible
- Sensor signals are sinusoidal

The crosstalk for these simulations is estimated by using the elimination technique. The desired sensor channel into which the crosstalk from the other sensors is calculated is not activated. In other words, the sensor is not perturbed whereas all the other sensors are activated by perturbing them. In an ideal situation where there is no crosstalk there should be no power in this channel after multiplexing. The power observed under these circumstances is the absolute value of the crosstalk. Crosstalk measurements are made by estimating the ac power transferred in the sensor channel under consideration.

The sensors operate at frequencies of less than 100Hz. This reduces the sampling frequency and helps in reducing the run time for the simulation. The bit rates of the code sequence is maintained at 1MHz. Various parameters of the system were varied in order to study their effects on the performance. The experiments were performed with the system elaborated in section 5.1.2.

The simulations and an analysis of the results are presented in the next few pages.
**Experiment 1:** Effect of increasing the number of sensors on the system. This simulation is conducted with increasing the number of sensors on the system.

- Signal-to-noise ratio = 10;
- code length = 15;
- modulation depth of sensors = .25;

![Graph](image)

**Figure 6.1. Effect of increasing the number of sensors on the system**

The performance of the system is limited by the number of sensors in the system as can be seen from Figure 6.1. This is in accordance with the theory developed in chapter 4. The number of sensors which can be multiplexed is limited by the length of the sequence used.
**Experiment 2:** Effect of extinction ratio on performance. In this simulation the extinction ratio of the laser signal varied. This is done by adding a constant dc signal to the transmitted coded signal in steps of .2 volts. When the signal amplitude is one, and the \( dc \), the extinction ratio is given as in equation is \( r = \frac{P_0}{P_1} = \frac{dc}{1+dc} \). The experiment was conducted under the following parameters.

- Signal-to-noise ratio = 15
- code length = 15
- Number of sensors = 7; connected as in vector [q],

where \( q = [1 \ 0 \ 0 \ 1 \ 0 \ 1 \ 0 \ 0 \ 1 \ 0 \ 0 \ 1 \ 0 \ 1 \ 1] \). A '1' represents a connected sensor and '0' represents the absence of a sensor in the respective bit position of the code sequence.

**Figure 6.2. Effect of laser modulation extinction ratio on performance**

It is seen from Figure 6.2 that the crosstalk into the desired signal increases with an increase in extinction ratio.
Experiment 3: Effect of unequal powers in the sensors. This simulation was conducted by observing the power into the channel of the first channel when that sensor is inactive and many of the others are. The signal power in the other active channels is increased in steps of .2, as a ratio of the signal power in sensor 1. The results of this test are shown in Figure 6.3.

\[
\text{Signal-to-noise ratio} = 15; \\
\text{code length} = 15; \\
\text{modulation depth of sensors} = .25;
\]

![Graph](image)

**Figure 6.3. Effect of unequal powers in the sensors**

From Figure 6.3, it is seen that the crosstalk into a sensor channel increases if the power in the other channels is significantly larger than the signal in the sensor under consideration. This result agrees with the analysis in chapter 4 (Equation 4.39).
**Experiment 4:** An important parameter of interest is the effect of asynchronous correlation. This was studied by simulating different scenarios of correlation mismatch. The delay is varied in steps of a tenth of a bit period.

- Signal-to-noise ratio = 15;
- code length = 15;
- modulation depth of sensors = .25;

![Graph showing effect of asynchronous correlation on performance](image)

**Figure 6.4. Effect of asynchronous correlation on performance**

The power from neighboring sensor channels falling into the sensor channel under consideration increases with the slip in synchronization. This can be seen from Figure 6.4 which is an illustrates this phenomenon. Figures 6.5a-h illustrate how the channel signal in sensor channel 3 of the simulation disappears from the output and finally sensor 4 of the next channel appears as the delay totals a bit period. In this run of simulations the delay was increased in steps of a seventh of a bit period.
delay = 0/7th of the bit period

delay = 1/7th of the bit period

delay = 2/7th of the bit period

delay = 3/7th of the bit period

Figure 6.5a-d. Detected signal being correlated with a reference signal which is the original delayed by 2 bit periods: (b) to (d) show synchronized correlation. (a) Sensor 3 signal is received when the bit delay is exactly 2 bits from the original. (b) to (h) the received signal when the reference signal is delayed by fractions of a bit period.
Figure 6.5e-h. The detected signal being correlated with a reference signal which is the original delayed by 2 bit periods. (e) to (h) show synchronized correlation. (h) Sensor 4 signal is received when the asynchronism is matched by an additional bit delay.
7.0 Applications and Conclusions

Two possible novel applications of this approach in multiplexing are proposed here. One application is to smart structures and the other is to sensing along communication channels.

7.1.0 Applications

7.1.1 Smart Structures

A smart structure is one which is adequately well instrumented as to allow the observance in real time, of the state of the structure in all its critical aspects. The purpose of smart structures is to monitor manufacturing processes, measure loading and environmental parameters and to asses the structural integrity with time.

Recent development of fiber composite materials have made it possible to directionally transfer the structural response of a material to another point at which it can be monitored. Depending on the structure, its design and, loading criteria, fiber networks can be tailored to sense and carry the information from any section of the structure. The lamination concept in today's structural industry has made this application more attractive.

A distributed fiber system is an inherent feature of any feasible smart structure design. In a multi-sensor situation in developing a smart structure, one would ideally need as many sensors to monitor a matrix of locations as the number of nodes in the network. This would put an immense demand on the hardware and processing if multiplexing was not
implemented. Multiplexing of this matrix of sensors would reduce this complexity, but one would still need $N$ sensors to do monitor as many points. A topology is proposed here which uses spread spectrum multiplexing to achieve this capability with the number of sensors being reduced by an order proportional to the square root of the number of nodes. The system uses microbend sensors.

The sensing fibers are embedded in the composite structure as shown in Figure 7.1. The design proposed here uses $2N$ sensors to sense $N^2$ locations. The sensing locations form a matrix as shown. The launched light is distributed through a star coupler, into the $2N$ fibers leading into the sensors. Microbend sensors are used for this implementation. Their effective sensing region is the entire length of the embedded fiber.

![Diagram of sensing fibers and components](image)

**Figure 7.1.** Model for developing smart structures using spread spectrum fiber sensor multiplexing techniques.
7.1.2 Mono-Fiber Communications and Sensing

The same fiber may be used for transmitting sensor and communications signals. A scheme is proposed here as to how this may be achieved.

In a communication system where code division multiple access methods are being used, the sensor model proposed can be incorporated. This involves the use of tap off couplers from the communication channel. The coupler may be of unequal splitting ratio with most of the power being on the main channel and a low percentage dependent on the power budget of the system coupled into the sensor.
Figure 7.2. Spread spectrum communications and sensing on the same fiber

In this topology the sensors used are necessarily reflective mode sensors. The sensor information is available only in the reflected path. It would increase the noise in the communication channel if the sensor signal is allowed to propagate into the communication channel receiver. The separation between the sensors is determined as an integral multiple of a bit delay in the code sequence. This gives the best autocorrelation
peak. At the receiver of the system, a delayed version of the sequence, delayed by the propagation through the length of the fiber, is used to correlate with the reflected signal. This system will work even if one of the communication channels use the same code sequence as it's signature sequence.

Such a configuration can strive to be useful in situations where the fiber cable needs to be monitored under harsh environments like underground or undersea.

Some constraints are foreseen in this topology. Among them is the sacrifice of bandwidth to the analog sensors. All the optical fiber sensors encountered so far have analog signal outputs. This increases the power budget on the otherwise digital communication link. Signal-to-noise ratio is a major consideration in analog systems and cannot be handled as efficiently as in digital systems.

7.2 Conclusions

Over the last decade optical fibers have gained wide acceptance as a measurement and instrumentation medium. To use its several advantageous properties it is necessary to adapt quickly to various industrial and technological requirements. It must be possible to incorporate newer technologies into this area to maximize benefits. One such is the spread spectrum technology.

Here a optical fiber based sensor array has been multiplexed using the spread spectrum technology. The dependence of the performance of such a system on various system parameters is analyzed. A mathematical foundation is laid to understand the theory of
spread spectrum system as applied to the multiplexing scheme proposed. Hardware implementation is supplemented by simulation of the scheme in software and its characteristics studied. The system establishes a reliable method for multiplexing an array of four Extrinsic Fabry-Perot Interferometers.

The simulation of the system in software is used to demonstrate the dependence of the scheme on several parameters of concern. Finally a few applications of this technique of multiplexing are proposed.

Future work could incorporate external modulation of the laser and electro-optic correlation of the output. This will eliminate the need for fast lasers and detectors. Another area of improvement is to use signals of known frequency to modulate the sensors and observe the output in the frequency domain to obtain crosstalk figures.
8.0 References


Vita

Ravikumar K. C. was born on February 15, 1967 in Bangalore, India to Vijaya and Sukumara Menon. He received his Bachelor of Engineering Degree in Electrical Engineering, in 1990 from Bangalore University, India. He worked as a Systems Engineer at Powerica Ltd., Bangalore for about two years before he enrolled for his Master's in Electrical Engineering at Virginia Polytechnic Institute & State University, Blacksburg, Virginia.

Mr. Ravikumar K. C. is now employed as a Design Engineer in the semiconductor industry in Austin, Texas.