Time-Domain Addressing of Remote Fiber-Optic Interferometric Sensor Arrays

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Abstract—This paper describes and analyzes a particular application of high duty-cycle time-division multiplexing to the separation and identification of signals from an interferometric sensor array. Using the method discussed here, the coherence length of the laser is no longer a severe design constraint. Also, the source phase-induced intensity noise which limits some other multiplexing methods may be overcome, leading to a higher sensitivity. The arrays of all-passive remote sensors exhibit minimal crosstalk between sensors, and have download insensitivity. A synthetic heterodyne demodulation technique prevents environmentally induced signal fading. Analysis includes coupling ratios for all directional couplers in the system, signal and noise spectra, minimum detectable phase shift, and the effect of ac coupling on noise and crosstalk. An experimental all-fiber implementation of a two sensor array has yielded a measured sensitivity of approximately 10 μrad/√Hz over a range of signal frequencies, and a crosstalk level of better than 55 dB.

I. INTRODUCTION

OVER THE LAST several years, interferometric fiber sensors have been developed for application in a wide range of fields [1]. Interest has been shown in arranging such sensors in passive element arrays. The majority of the recent publications on interferometric sensor arrays have focused on four multiplexing methods: frequency-division multiplexing (FMCW) [2], “coherence multiplexing” [3], [4], time-division multiplexing using unbalanced interferometers [5], [6], and time-domain addressing using an ultimately balanced interferometer [7].

Both the FMCW and coherence multiplexing methods require that there be unbalanced path lengths throughout the system. In the case of the FMCW system, the beat frequency which identifies each sensor arises from a path mismatch which delays one ramp relative to another. Theoretical calculations and experimental results agree that the minimum detectable phase shift in FMCW systems is limited by phase-induced intensity noise [8], to approximately 0.1 mrad/√Hz [9]. For the coherence multiplexed system, the paths which produce the signal are very well matched. However, because the optical signal is continuous, all of the various paths through the system contribute light at each detector. These unequal paths contribute to phase-induced intensity noise, which eventually limits the sensitivity of the system. Theoretical and experimental results [3] show that a coherence multiplexed system powered by a continuous wave single-longitudinal-mode laser diode is limited to a sensitivity on the order of 1 mrad/√Hz. It has been recently reported that by combining frequency-modulation of the laser diode and coherence multiplexing, it is possible to “heterodyne” the phase-induced intensity noise out of the signal frequency band [10]. This technique improves the sensitivity of the sensor array, but the requirement of unequal path delays between sensors, plus the fact that these delays can extend to undesirably large values, remain as disadvantages [11].

One method of implementing a time-domain addressing scheme was suggested by Henning et al. [5]. A long coherence length laser was gated at two different times, providing pulses with two different frequency shifts. The time differential between the two pulses was set to be equal to the propagation delay between adjacent arms in the sensors. The two pulses overlapped after passing through the sensor and provided a heterodyned signal. The use of an unbalanced interferometer is inherent in this technique so that the two pulses which enter the sensor at different times may interfere. Using unbalanced interferometers without compensation (described below) results in the generation of phase-induced intensity noise, which may limit sensitivity. The optical source must have a coherence length much greater than the relative delay through the interferometers, so for systems of this type a gas laser is required. In addition to the source problem for this array, the use of Fabry–Perot interferometers, as was suggested by Dakin et al. [6], results in an inherent crosstalk limitation due to the higher order reflections which naturally result.

We have previously described [7] an approach to passive interferometric sensing arrays using time domain addressing which is passive, avoids phase-induced intensity noise, permits equal path delays between sensors, and can be powered by laser diodes. In what follows, we present an analysis of performance of an array of this kind and measurements on an experimental system containing two sensing interferometers.

In this array, we use an unbalanced interferometer in each sensor location, and compensate for the path imbalance by placing a single equally unbalanced interferometer at one end of the array. The addition of the compensating interferometer allows any two paths composing...
two arms of an interferometer to be made equal. Thus, a long coherence length source is no longer required, and excellent sensitivity may be obtained. Since the source is gated, light arriving at the detector from various paths which do not contribute to the monitored sensor will be gated, light arriving at the detector from various paths long coherence length source is no longer required, and excellent sensitivity may be obtained. Since the source is two arms of an interferometer to be made equal. Thus, a system, as they do in FMCW and coherence multiplexed systems.

In Section II we describe the array topology. Section III discusses the requirements for optimizing optical power through the sensor array. The spectra of the signal and noise, and the signal-to-noise ratio resulting from the array and signal-processing electronics are presented in Section IV. In Section V, questions of crosstalk and noise related to detector coupling are treated. The experiments which have been performed to measure the performance of the time-domain addressed array are described in Section VI. Finally, in Section VII we summarize our results.

II. Sensor Array Geometry

Figs. 1 and 2 show two possible array configurations for time-division multiplexing. In the circuit of Fig. 1 the sensing interferometers, which are identical to the compensating interferometer, are placed in the rungs of the ladder structure. The arrangement of Fig. 2 has the advantage that the input and/or output bus contains the sensor coils, so that no additional delay line is needed to separate the pulses from adjacent sensors and the required number of optical components is minimized.

A series of pulses with repetition rate \( \frac{1}{\tau} \) is launched into the input bus of the fiber-optic sensor array. Each sensor consists of an unbalanced interferometer which generates two pulses from a single input pulse. The time delay \( \Delta T \) between the two pulses is greater than or equal to the width of the pulse. In the sensors, phase information is impressed differentially with respect to the two arms of the interferometer. The pulses from the two arms do not overlap and hence do not interfere. After traveling back through the output bus the pulses enter another interferometer, which has a path difference \( \Delta T \) between the two arms to compensate for the initial path difference. Two pulses which have traveled equal length but different paths (through a long arm in one interferometer and a short arm in the other) interfere, and convey information about the differential phase modulation occurring along the two routes through the sensor and compensator. If desired, the compensating interferometer may be placed at the input of the system and the interference take place at the output of the sensors.

Every pulse from this array can provide information from a sensor, except the first and last pulses, so that the output duty cycle can be high. For each input pulse into an \( N \) sensor array of this type there will be \( (N + 2) \) output pulses. By setting the repetition rate of the input pulses to \( \tau = (N + 1) \Delta T \), the output duty cycle is maximized at \( N/(N + 1) \), which approaches unity for a large array. In contrast, only every second pulse from the circuit of Fig. 1 can carry information from the sensors. The other pulses contain information presumably of no interest, and may also contain phase-induced intensity noise, depending on the length of the delay lines. This array achieves its maximum duty cycle of 0.5 when the delay between interferometers is \( \Delta T \) and the repetition rate is set to \( \tau = 2N\Delta T \). For these optimized cases, there are pulses resulting from the overlap of the first and last pulses from different input pulses which will contain a large amount of phase noise-induced intensity noise. Alternatively, the generation of phase-induced intensity noise may be eliminated by setting \( \tau = (N + 2) \Delta T \) and \( 2(N + 1) \Delta T \) for Figs. 2 and 1, respectively. In either case, the non-signal-bearing pulses are not routed to the signal processing electronics.

The signal which is detected consists of phase modulation impressed on the optical carrier by the variation of the fiber refractive index due to the pressure of an acoustic wave. The modulation of the fiber index by an acoustic wave was described by Bucaro et al. [12]. The optical pulses generally sample the signal at a frequency much higher than the acoustic frequency. The pulses from different sensors are separated at the output of the system using either an optical switch before the detector or elec-
tronic switch after detection. The pulses which have sampled the acoustic signal from a given sensor are routed to signal processing electronics, which produce the non-fading signal for measurement.

III. OPTICAL POWER OPTIMIZATION OF THE SENSOR ARRAY

In analyzing our time-domain addressed system, we first specify the coupling constants used in the sensing array and in the compensating interferometer. We assume that it is desirable to have equal sized pulses from all rungs of the ladder array in order to obtain uniform sensitivity from all sensors.

Consider the sensing portion, or ladder, of the system shown in Fig. 2. For an array of \( N \) sensors there will be \( N + 1 \) rungs in the ladder. The rungs are numbered \( 1 \) through \( N + 1 \) with a variable \( i \) beginning with \( i = 1 \) for the rung nearest to the light source. Directional couplers connect each rung to the input and return bus.

We assume that light must couple to go from the input bus to the ladder rung, and again from the rung onto the output bus, as is shown in Fig. 2. As was discussed in a previous publication [3], couplers on the input and output bus of the same rung should be set to the same coupling ratio to maximize signal power. Thus, we assume that the directional couplers on the input and output bus for rung \( j \) have the same power coupling ratio \( \kappa_j \). Light returning from sensor \( j \) will have passed through couplers \( 1 \) through \( j \) on both the input bus and return bus. Couplers \( 1 \) through \( j - 1 \) will each have transmitted \( 1 - \kappa_i \) of the power in the pulse to sensor \( j \).

In general, there will be some loss in the fiber, splices, and couplers in the system. We will assume that there is the same amount of loss in each similar section of the array, regardless of its relative position. We designate the power transmission (excluding coupling) through a rung of the ladder by \( \alpha_i^2 \), and the power transmission through the section of the ladder between two rungs (on the input and output bus) by \( \alpha_i^2 \). The peak power returning from rung \( j \) to the central processing location is thus given by

\[
P_{\text{in}, \text{return}} = \kappa_j^2 \alpha_j^2 P_{\text{in}}, \quad j = 1
\]

\[
P_{j, \text{return}} = \kappa_j^2 \alpha_j^2 \left[ \prod_{i=1}^{j-1} (1 - \kappa_i)^2 \right] \alpha_{j-1}^2 P_{\text{in}}, \quad j \geq 2 \tag{1}
\]

where \( P_{\text{in}} \) is the peak power of the input pulse into the array. By assuming that we want the same power from each rung of the array to return to the central processor, i.e., \( P_{i, \text{return}} = P_{j, \text{return}} \) for all \( i \) between \( 1 \) and \( N + 1 \), we derive a relation between the coupling ratios:

\[
\kappa_j = \frac{\alpha_b \kappa_{j+1}}{1 + \alpha_b \kappa_{j+1}}. \tag{2}
\]

The last rung does not need couplers since no power is needed for further sensors. Thus \( \kappa_{N+1} = 1 \), and (2) gives

\[
\kappa_j = \frac{1 - \alpha_b}{\alpha_b^{N-j-1} - \alpha_b}. \tag{3}
\]

As is evident from this expression, the coupling ratios can be significantly affected by the loss due to the fiber, splice, and coupler.

Substituting this result back into (1) we find that the total power transmitted through every rung of the ladder array is

\[
P_{\text{return}} = \left( \frac{1 - \alpha_b}{\alpha_b^N - \alpha_b} \right)^2 \alpha_j^2 P_{\text{in}}. \tag{4}
\]

This is the peak power per pulse at the output of the ladder (which is also the input to the compensating interferometer) resulting from the geometry shown in Fig. 2, if the couplers are set according to (3). In the limit of \( \alpha_v = \alpha_b = 1 \), i.e., no excess loss, the coupling ratios and power returned reduce to

\[
\kappa_j = \frac{1}{N - j + 2} \tag{3'}
\]

and

\[
P_{\text{return}} = \frac{P_{\text{in}}}{(N + 1)^2}. \tag{4'}
\]

Given that the input pulses to the compensator are equal in power, the optimum coupling ratios for the directional couplers in that interferometer should be determined to give the maximum sensitivity. Assuming shot noise limited detection, the modulation depth of the interfered signal should be maximized. It is originally assumed that the directional couplers have coupling ratios \( \kappa_{c1} \) and \( \kappa_{c2} \) which may be, but are not necessarily, the same. If optical power \( P_c \) enters a compensator which has power transmissions in the long and short arms of \( \alpha_i^2 \) and \( \alpha_j^2 \), respectively, the detected signal power in either output port is proportional to

\[
S_x = 4P_c^2 \alpha_i^2 \alpha_j^2 \kappa_{c1} \kappa_{c2} (1 - \kappa_{c1}) (1 - \kappa_{c2}) \Delta \phi_{\text{rms}}^2 R^2. \tag{5}
\]

It has been assumed that \( \Delta \phi_{\text{rms}} \), the sensed phase modulation impressed on the optical carrier, is small and the interferometer is biased to the point on the response curve where it exhibits maximum sensitivity. \( R = e \eta / (hv) \) is the responsivity of the detector, where \( e \) is the charge of an electron, \( \eta \) is the quantum efficiency of the detector, \( h \) is Planck's constant, and \( \nu \) is the frequency of the light.

The average light level at the detector is determined by the output port which is monitored. For the port which is detected in Fig. 2, the shot noise term is given by

\[
S_n = [(1 - \kappa_{c1}) (1 - \kappa_{c2}) \alpha_i^2 + \kappa_{c1} \kappa_{c2} \alpha_j^2] 2eRB \eta P_c. \tag{6}
\]

To optimize the signal to noise ratio per hertz for this port, we then maximize

\[
\text{SNR} = 4 \alpha_i^2 \alpha_j^2 \left[ \frac{\kappa_{c1} \kappa_{c2} (1 - \kappa_{c1}) (1 - \kappa_{c2})}{(1 - \kappa_{c1}) (1 - \kappa_{c2}) \alpha_i^2 + \kappa_{c1} \kappa_{c2} \alpha_j^2} \right] \frac{\eta}{(hv)} \frac{P_c}{2B} \Delta \phi_{\text{rms}}^2. \tag{7}
\]
The optimum result for the above equation can be found to be

\[ K_{c1} = K_{c2} = \frac{1}{1 + (\alpha^2_1/\alpha^2_2)^{1/3}}. \]  

When there is no loss, or the relative loss in the two arms of the compensator are equal, the optimum solution is at 50-percent coupling. To optimize the signal-to-noise ratio per hertz for the other port, the solution is \( K_{c2} = (1 - K_{c1}) \), where \( K_{c1} \) is given by (8).

By combining the expression for the peak optical power per pulse in (4) with the power through the compensating interferometer, we get

\[ P_{\text{out}} = \alpha_1^2 \left( \frac{1 - \alpha_b}{\alpha_b - N - \alpha_b} \right) \left[ \alpha_1^2 K_c^2 + \alpha_2^2 (1 - K_c)^2 \right] P_{\text{in}} \]

\[ = 2 \gamma P_{\text{in}} \]  

(9)

for the peak power in each interfering pulse through the array, where it is assumed that we have used the port which results in \( K_{c1} = K_{c2} = K_c \). 

IV. SIGNAL AND NOISE SPECTRA AND SIGNAL-TO-NOISE RATIO

In this section we present general expressions for the signal and noise spectra which are important in the sensor sensitivity calculations. We assume that the only source of noise is the quantum white noise generated in the laser. The acoustic signal detected by the sensor array is narrow-band centered about zero frequency.

In the case of electronic demultiplexing, our model for the signal and noise spectra calculations is shown in Fig. 3(a). It consists of the following sequence of elements:

(a) A laser source whose output is CW light with a dc optical power of \( P_{\text{in}} \) and quantum white noise.

(b) An optical switch, with a basic pulse shape of \( p_1(t) \) and a switching period of \( \tau \), which in essence samples the dc light and the white noise of the laser output. It is assumed that this switch is wide-band, however, the switching speed determines the basic pulse shape \( p_1(t) \).

(c) A fiber sensor array which receives the optical pulse trains from the output of the optical switch and samples the acoustic signal both in time and space. In addition, the array performs a multiplexing operation by placing the output signals from all of the sensors in nonoverlapping time slots. The spatial sampling period is set by the physical separation of adjacent sensors. For any particular sensor in space, the time sampling period is \( \tau \). We assume that in the sensing process only the acoustic signal is sampled and no further noise is added.

(d) A photodetector which detects the intensities of the pulses coming from the array. The bandwidth of the detector is determined by its transfer function \( H_1(f) \).

(e) An electronic switch with a basic pulse shape of \( p_2(t) \) and a switching period of \( \tau \), synchronized with the optical switch in (b) with the proper delay. This second switch actually performs a demultiplexing operation by sampling the signals from a given single sensor. Further, non-signal-bearing pulses are also discarded. Again it is assumed that the switch is wide-band, however, the switching speed determines the basic pulse shape \( p_2(t) \).

(f) A narrow-band (low-pass) filter, with transfer function \( H_2(f) \), centered around the signal frequency to recover the continuous acoustic signal from its sampled pulse train.

A subject of importance in systems involving sampling and switching operations is the aliasing effects. Since the noise is broad-band (white) and the sampling frequencies are finite, the noise aliasing effects are present and should not be ignored. Assuming that the Nyquist sampling rate mentioned above is maintained for the acoustic signal, the signal is not aliased.

Using the above model we have calculated both the signal and noise spectra at the outputs of all the elements described earlier. At the output of the laser the squared optical intensity spectrum at dc is

\[ S_{\text{in}}(f) = P_{\text{in}}^2 \delta(f) \]  

(10a)

where \( \delta(f) \) is the dirac delta function. The noise power spectrum is

\[ S_{\text{n}}(f) = P_{\text{in}} h \nu. \]  

(10b)

Thus, the square of the optical dc signal-to-noise ratio per hertz \( \text{SNR}_1 \) is given by

\[ \text{SNR}_1 = \frac{S_{\text{in}}(f)}{S_{\text{n}}(f)} = \frac{P_{\text{in}}^2}{P_{\text{in}} h \nu} \]  

(10c)

where \( h \nu \) is the photon energy and, in this case, \( P_{\text{in}} \) is also the average power.

At the output of the first optical switch the squared optical intensity spectrum is

\[ S_{\text{out}}(f) = \left[ f_{\text{sp}}^2 \sum_{k=-\infty}^{\infty} |P_1(k f_{\text{sp}})|^2 \delta(f - k f_{\text{sp}}) \right] P_{\text{in}}^2 \]  

(11a)

where \( f_{\text{sp}} = 1/\tau \) is the sampling frequency of each sensor and \( P_1(f) \) denotes the Fourier transform of the pulse shape \( p_1(t) \). The noise power spectrum is

\[ S_{\text{out}}(f) = f_{\text{sp}}^2 \sum_{k=-\infty}^{\infty} |P_1(k f_{\text{sp}})|^2 = \left[ f_{\text{sp}}^2 \sum_{k=-\infty}^{\infty} |P_1(k f_{\text{sp}})|^2 \right] P_{\text{in}} h \nu \]  

(11b)

which is again white as in (10b), but with a different intensity. Note that in (11b) the noise is aliased after the sampling operation. The square of the optical dc signal-to-noise spectral density ratio \( \text{SNR}_2 \) is given by

\[ \text{SNR}_2 = \frac{P_1(0)^2}{\sum_{k=-\infty}^{\infty} |P_1(k f_{\text{sp}})|^2} \frac{P_{\text{in}}}{h \nu} = \left( \frac{T_{\text{nsp}}}{\tau} \right) \frac{P_{\text{in}}}{h \nu} \]  

(11c)
where $T_{ac1} \equiv \int_{-\infty}^{\infty} p_1(t) dt^2 \int_{-\infty}^{\infty} p_1^2(t) dt$ is the autocorrelation width of the pulse shape $p_1(t)$. In the special case when $p_1(t)$ is a rectangle function of width $T_{p1}$, its autocorrelation width is also $T_{p1}$. As a result, the right hand side of (11c) becomes the product of the duty cycle per sensor and the input power, which in turn is the average power per sensor. This is a result consistent with intuition.

The acoustic signal sampling and interference effects by the fiber sensor array result in the addition of sidebands to the spectrum of the pulse train intensity in (11a) which is the input to the array. Furthermore, the output power reaching the detector drops due to the array structure (i.e., the use of directional couplers) and excess loss, discussed in the previous section. It can be shown that at the input of the second filter the spectrum of the electrical signal is

$$S_{en}(f) = (2\gamma P_{in} R)^2 \cdot \left[ f_{sp}^4 \sum_{k=-\infty}^{\infty} \left| P_1(kf_{sp}) \right|^2 \right] \cdot H_1(f - k'f_{sp} + kf_{sp})^2 \delta(f - k'f_{sp})$$

(12a)

where $\gamma = P_{out}/(2P_{in})$ (see (9)) and $P_2(f)$ denotes the Fourier transform of the pulse shape $p_2(t)$. The electrical noise power spectrum is

$$S_{ne}(f) = \left[ f_{sp}^4 \sum_{k=-\infty}^{\infty} \sum_{k'=\infty}^{\infty} \left| P_1(kf_{sp}) \right|^2 \right] \cdot H_1(f - k'f_{sp} + kf_{sp})^2 (2\gamma P_{in})(2eR).$$

(12b)

The single-sided dc electrical signal-to-noise spectral density ratio $\text{SNR}_3$ is given by

$$\text{SNR}_3 = \frac{\eta}{2hv} \left[ f_{sp}^4 \sum_{k=-\infty}^{\infty} \sum_{k'=\infty}^{\infty} \left| P_1(kf_{sp}) \right|^2 \right] \cdot \frac{P_2(-kf_{sp}) H_1(kf_{sp}) 2\gamma P_{in}}{2\gamma P_{in}}$$

(12c)

If the pulse shape $p_2(t)$ is such that the pulse shape convolved with the filter impulse response $p_1(t) * h_1(t)$ remains unchanged under their multiplication, then (12c) reduces to

$$\text{SNR}_3 = \left( \frac{\eta}{2hv} \right) \sum_{k=-\infty}^{\infty} \left| P_1(kf_{sp}) H_1(kf_{sp}) \right|^2 2\gamma P_{in}$$

$$= \left( \frac{\eta}{hv} \right) \left( \frac{T_{ac1}}{\tau} \right) \gamma P_{in}$$

(13)

where now $T_{ac1}$ is the autocorrelation width of $p_1(t) * h_1(t)$. In this case, for given pulse shape $p_1(t)$ and $h_1(t)$, the expression in (13) can be evaluated by calculating the autocorrelation width $T_{ac1}$.

To maximize the signal-to-noise ratio for a given input power, the ratio $T_{ac1}/\tau$ should be made as large as possible. However, to minimize crosstalk between sensors $T_{ac1}$ should be less than the time delay through the sensors $\Delta T$. Optimally the bandwidth of the switch should be sufficiently large to have $T_{ac1}/\Delta T \approx 1$.

If an optical switch is used for demultiplexing in place of the electronic switch, as shown in Fig. 3(b), only the order of the detection and switching operations is interchanged. In that case, the final signal to noise ratio given by (12c) and (13) is independent of the bandwidth of the detector $H_1(f)$.

The shot noise limited sensitivity (incorporating the effects of shot noise aliasing) is given by

$$\frac{\Delta \phi_{rms}}{\sqrt{B}} = \frac{1}{\sqrt{\text{SNR}_3}} = \left( \frac{1}{\sqrt{\eta/\hbar \nu}} \right) \left( \frac{T_{ac1}}{\tau} \right) \gamma P_{in}$$

(14)

When the repetition rate is set at its maximum rate $\tau = (N + 1) \Delta T$. In this case,

$$\frac{\Delta \phi_{rms}}{\sqrt{B}} = \frac{\hbar \nu (N + 1) \Delta T}{\eta T_{ac1} P_{in}}$$

(15)

Fig. 4 shows the shot noise limited sensitivity for an array containing $N$ sensors. It has been assumed that $\eta = 0.5$ and $\nu = 3.66 \times 10^{14}$ Hz. The laser is assumed to have an output power of 10 mW, pulsed using a Bragg cell having a conversion efficiency of 50 percent. The solid line $a$ shows the result for a lossless array with $T_{ac1}/\Delta T = 1$. For such an ideal system 13 sensors with shot noise limited minimum detectable phase modulation of approximately 1 $\mu$rad/$\sqrt{\text{Hz}}$ may be multiplexed. If the system is lossless but the input pulse shape is significantly different than a rectangle of width $\Delta T$ then the number of sensors which can be multiplexed diminishes. For example, if $T_{ac1}/\Delta T = 0.5$, as shown by line $b$ in Fig. 4, then only 10 sensors may be multiplexed while maintaining 1 $\mu$rad/$\sqrt{\text{Hz}}$ sensitivity.

The dotted lines in Fig. 4 show the sensitivity for a particular nonideal system. We assume that there is a single splice on each rung of the ladder, on the return bus between couplers, and on the short arm of the compensating interferometer. Each splice has 0.2-dB loss. Where a delay line is placed (i.e., on the input bus between rungs
If electronic switching is used after the detector, it is essential that the detected signal be dc coupled through the final sampling. AC coupling the detector output before the switch will result in crosstalk from other sensors. In addition, assuming that the nonsignal bearing pulses are overlapped for optimal duty cycle of the system, source phase-induced intensity noise will be generated in the resulting pulse, and ac coupling will result in leakage of phase noise into the sensor pulses.

For convenience we assume that the signal of interest is detected by sensor 1. Following the notation of the previous section, at the output of the detector we have the following frequency spectrum for all of the pulses from all N sensors:

\[
Y_{\text{tot}}(f) = f_{sp} \sum_{n=1}^{N+1} \sum_{m=-\infty}^{\infty} e^{j2\pi(n-1)mf_{sp} \Delta T} P_1(mf_{sp}) \\
\cdot [\delta(f - mf_{sp}) + X_{\phi_i}(f - mf_{sp})]
\]

where we assume a ladder array of N rungs, and no change in the pulse when it passes through the filter, i.e., \(H_1(f) = 1\). The spectra of the signal and/or noise in pulse n is given by \(X_{\phi_i} \cdot Y_{\text{tot}}(f) \) is the spectrum of all of the pulses out of the array including the dc component and the signal spectra \(X_{\phi_i}\) that the pulses carry. If we ac couple the signal eliminating all frequencies within a bandwidth \(\Delta B/2\) around zero frequency, where \(\Delta B/2 \ll f_{sp}\), we then have

\[
Y_{\text{ac}}(f) = Y_{\text{tot}}(f) - f_{sp} \sum_{n=1}^{N+1} P_1(0) \delta(f) \\
+ X_{\phi_i}(f) \Pi\left(\frac{f}{\Delta B}\right)
\]

where \(\Pi(f/\Delta B)\) is the rectangle function, unity from \(-\Delta B/2\) to \(\Delta B/2\) and zero elsewhere.

Passing the filtered spectrum through the electronic switch, we get

\[
Y_{\text{ac}}(f) = f_{sp} \sum_{m=-\infty}^{\infty} P_2(mf) \delta(f - mf_{sp})
\]

where \(*\) denotes convolution. The term in the first set of brackets corresponds to the desired signal, which we would obtain with dc coupling. The second term contains the effects of ac coupling.

The second term of (18) can be rewritten as

\[
\Delta Y(f) * f_{sp} \sum_{m=-\infty}^{\infty} P_2(mf) \delta(f - mf_{sp})
\]

\[
= f_{sp} P_1(0) \sum_{n=1}^{N+1} \sum_{m=-\infty}^{\infty} P_2(mf) [\delta(f - mf_{sp})] \\
+ X_{\phi_i}(f - mf_{sp}) \Pi\left(\frac{f - mf_{sp}}{\Delta B}\right)
\]
This result can be rewritten as

\[ Y_{\text{LPF}}(f) = f_{\text{LPF}} \int_{-\infty}^{\infty} P_1(t) P_2^*(f) \, df \left[ \delta(f) + X_{\phi_1}(f) \right] \]

\[ - f_{\text{LPF}}^2 P_1(0) P_2(0) \left( N + 1 \right) \delta(f) \]

\[ + \sum_{n=1}^{N+1} X_{\phi_n}(f) \right] \Pi \left( \frac{f}{\Delta B} \right). \]  

(20)

The first term is the signal term, and is decreased by \((N + TT;)AT\). The second term contains only the dc component, and carries no sensor signal. The third term contains the crosstalk from other sensors as well as noise from the phase-noise carrying pulse. The magnitude of the crosstalk term from any other given sensor at frequencies below \(\Delta B/2\) is given by the same expression as the decrease in the signal. Thus, we can define the crosstalk ratio at a given frequency within the \(\Delta B/2\) band as

\[ \text{crosstalk}_{\text{ac coupling}} = \frac{\left[ T_{\text{cc}} / (N + 1) \Delta T \right]}{1 - \left[ T_{\text{cc}} / (N + 1) \Delta T \right]} \]  

(22)

For a large array \((N \gg 1)\), the effect of ac coupling becomes negligible. However, for a sensitive sensor array we have shown above that a reasonable number of sensors is only on the order of ten. For such an array dc coupling is important for the elimination of this particular source of crosstalk between sensors if the sensor signal is within the bandwidth \(\Delta B/2\). However, if the acoustic signals have no components at frequencies below \(\Delta B/2\) then there is no effect. (Note: there are still other sources of crosstalk which will be discussed in the experimental section.) In addition, the use of dc coupling prevents the leakage of source phase-induced intensity noise at frequencies below \(\Delta B/2\) from the nonsignal bearing pulses into the pulses of interest.

VI. EXPERIMENTAL SYSTEM AND RESULTS

An experimental sensor array was constructed, as shown in Fig. 5. The geometry of Fig. 2 was used for the array. A continuous wave 830-nm single-mode laser diode (Hitachi HLP 1400) was gated by an acousto-optic cell with a 35-ns risetime. The Bragg cell was used, instead of direct modulation of the input current to the laser, in order to avoid modulation of the laser spectrum [14]. An optical isolator was placed between the laser and the Bragg cell to suppress optical feedback into the laser.

The 100-ns wide pulses were routed through fiber-optic Mach–Zehnder interferometers, each with a path difference \(\Delta T\) of about 230 ns. For good sensitivity of the sensor array, matching of the path imbalance \(\Delta T\) of the sensor interferometers with that of the compensating interferometer is critical. In order to ensure that phase-induced intensity noise does not arise to a measurable level on the signal pulse it is necessary to match the imbalance to within a small fraction of the coherence length of the laser. The accuracy with which fiber lengths can be matched constitutes a practical lower limit to the coherence length of the optical source which may be used in these arrays. Measurement of the path differences may be accomplished by amplitude modulating a laser diode and determining the characteristic frequency of the filtering of each Mach–Zehnder individually. By utilizing the high order frequency filtering characteristics of the unbalanced interferometers, relative time delays can be compared to within 1 mm out of 100 m, or 10 ppm. The details of this technique have been published previously [15]. In order to equalize the path differences, small lengths of fiber from one arm of an interferometer were removed by using capillary tubes to hold the fibers for splicing, and grinding down and resplicing the capillary tubes when length adjustment was required. The capillary tubes were polished at an angle to minimize reflections back into the laser, which would affect the laser spectrum. The differential path through each sensor was about 48 m, and was matched to the compensator delay to within 2.5 mm. The accuracy of path length matching was limited by our patience and facility with splices, rather than by the measurement technique.

The fiber system was constructed using tunable directional couplers and Corning single-mode sensor (high NA) fiber. Each interferometer contained a phase modulator, consisting of a fiber coiled around a piezoelectric cylinder, and a polarization controller. The phase modulators in the sensors were used to simulate acoustic signals. The phase modulator in the compensating interferometer was used to generate relatively high frequency modulation (10.4 kHz) for the synthetic-heterodyne demodulation technique [16] we employed to avoid signal-fading caused by optical phase drift due to environmental changes.

Signal fading may be caused by polarization drift as well as phase drift. For a system constructed of low-birefringence single-mode fiber, such as ours, manual adjustment of a single polarization controller in each sensor
is sufficient to overcome signal-fading due to the polarization drift in that sensor. The polarization controller in the compensating interferometer was used for preliminary experiments to ascertain that the optical switch was not leaking measurably. The polarization controllers are a laboratory tool, and would not be used in a practical system. Among the possible practical solutions are the use of polarizations-preserving fiber or polarization masking.

Fig. 6 shows two photographs of pulse trains from the two sensor array. The top photo shows two interfering pulses, from the two sensors and two extra pulses that passed through the longest and shortest optical paths through the system \((\tau = 4\Delta T)\). Time-varying signals were applied to the sensors, resulting in blurred signal pulses. The noninterfered pulses look clean and noiseless. The bottom photo shows the effect of overlapping the first and last pulses from consecutive input pulses, when \(\tau = 3\Delta T\): the single pulse which results is very noisy. This noisy pulse is discarded, and the system operates at a detector duty cycle of \(2/3\). An important consideration for sensor systems is the manner in which signal fading due to environmentally induced optical phase drift may be overcome. Several methods have been developed: active feedback [18], and heterodyning or synthetically heterodyning [16], [19], [20] the signal by placing a frequency shifter or phase modulator in one arm of the interferometer. If the interferometer has two matched paths, this would require the sensor to have an active element in its remote location. However, by using unbalanced interferometers any of the above methods may be applied at the signal processing location without compromising the passivity of the sensor array in the environment.

Our experiments investigated the sensitivity of the sensor array with two sensors operating, and measured the level of the crosstalk between sensors. The set of signal processing electronics we employed was designed to use the synthetic heterodyne demodulation technique of Dandridge et al. [16]. The noise floor of the electronics corresponded to approximately \(10 \mu\text{rad}/\sqrt{\text{Hz}}\) over a range of signal frequencies out to 2 kHz.
The minimum detectable phase shift in the sensor was ascertained by measuring the signal to noise ratio \( S/N \) displayed on the spectrum analyzer for a known, small phase modulation amplitude from the sensor. To calibrate the phase modulation amplitude induced on the optical signal, the voltage corresponding to 3.83 rad (which nulls the first Bessel function sideband) was measured at each "signal" frequency. Linearity of the phase modulator to applied voltage was checked. To correct for errors in the spectrum analyzer's measurement of the noise level \([21]\), 2.5 dB was subtracted from the measured \( S/N \).

The results of the sensitivity measurement for the dual-sensor array are shown in Fig. 7. The array is operated at its highest possible duty cycle, where \( \tau = 3\Delta T \). In this photograph the pulses from the second sensor (i.e., the sensor farthest from the laser) are being sampled. All other pulses are discarded. The noise spikes at multiples of 60 Hz which are noticeable at low frequencies were caused by the line voltage. A signal corresponding to 10 mrad (rms) at 800 Hz was applied to the phase modulator in the second sensor for comparison with the noise floor. The noise spikes at multiples of 60 Hz corresponds to the electronic noise floor of 10 \( \mu \)rad/\( \sqrt{\text{Hz}} \) mentioned above. The sensitivity was the same when the first sensor was sampled.

Fig. 7 also includes a signal peak at 1.3 kHz corresponding to a phase modulation of 100 mrad (rms) applied to the first sensor while the second sensor was monitored. Comparing the signals seen from each sensor, it was found that the crosstalk level was at most 1 percent of the direct signal. Further investigation led to the conclusion that the source of the crosstalk was the electronic gating used in the experiment. It was found that by placing an optical gate before the detector (in addition to the electronic gate after the detector) the crosstalk level could be reduced to 0.2 percent.

VII. Summary

In conclusion, we have described and analyzed arrays of all-passive remote sensors utilizing time-domain addressing which do not require highly coherent sources. The arrays are free from signal-fading, source phase-induced intensity noise, and downlead sensitivity, and capable of high detector duty cycle. A high duty cycle array was analyzed with respect to optimum construction and output. It was shown that the sampling of shot noise affects the noise spectrum and thus the final signal to noise ratio. Using our analysis, we determined the maximum number of sensors for a given sensitivity in an array fed from a single laser diode. The use of dc coupling of the detector to the electronic gate was shown to be important for eliminating a source of crosstalk and noise.

An experimental all-fiber-optic sensor array was discussed which exhibits sensitivity of approximately 10 \( \text{rad}/\sqrt{\text{Hz}} \) at signal frequencies out to 2 kHz. The measured sensitivity of our system was limited by the noise level of the electronics. Crosstalk between sensors in the array was investigated, and could be measured to be at least 56 dB below the signal level when the system was operated at its maximum repetition rate with both optical and electronic switches at the output. The use of better switches could improve crosstalk even further.

Acknowledgment

The authors would like to thank R. H. Wentworth for helpful discussions and computational assistance, and M. R. Layton at Litton Systems for helpful discussions and for providing several of the components used in the experimental system.

References


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