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Technical Report 1385
December 1990

**Wide-band RF Signal
Processing with
Optoelectronic Devices
and Fiber-Optic Delay
Lines**

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ADMINISTRATIVE INFORMATION

This document consists of two previously released papers: "Finite Impulse Response Filter for Wide Bandwidth Signal Processing," prepared for the Annual ONT IR/IED Program Review, June 1990, and "Integrated Optical Differential Amplifier With Large Dynamic Range for Gigahertz Bandwidth Electrical Signals," prepared for IEEE Photonics Technology Letters, October 1990.

The work reported in these documents was done under the Integrated Ocean Surveillance block program, Independent Exploratory Development (IED) project ZE80, "Variable Weight Tap for a Fiber-Optic Transversal Filter." The work was done in FY 90 by M. H. Berry and D. M. Gookin of the Processing Research and Development Branch, NOSC Code 761.

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SUMMARY

OBJECTIVE

Develop a new technology for processing wide bandwidth signals. This capability is important to the Navy for use in communications systems using high data rates, secure communications, RF signal processing, beamforming, and for handling complex signal environments in electromagnetic warfare.

RESULTS

Developed and demonstrated a fiber-optic finite impulse response (FIR) filter for processing 1-GHz bandwidth signals. The FIR filter was reconfigurable and could be adapted in realtime. This fiber-optic filter design with variable weights significantly increases the range of filtering operations possible for processing wide bandwidth signals. This type of FIR filter can be used to encode/decode wide bandwidth radar signals, to process spread spectrum signals, or to notch out agile jammers.

Developed and demonstrated a differential amplifier for gigahertz bandwidth electrical signals using fiber optics and integrated optical devices. This device could be used as an integrated module in radar and communications systems.

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CONTENTS

INTRODUCTION	1
FIBER-OPTIC FINITE IMPULSE RESPONSE FILTERS	1
FIR FILTER WITH VARIABLE WEIGHTS	2
NULL SHIFTING AND BEAM FORMING	5
POSITIVE AND NEGATIVE WEIGHTS	10
DIFFERENTIAL AMPLIFIER	14
SUMMARY	17
REFERENCES	19

FIGURES

1. Integrated optical coupler (IOC)	3
2. Fiber-optic FIR filter	4
3. Fiber-optic FIR filter system	5
4. Examples of corrected response of the transversal filter	6
5. Frequency response using a raised cosine weighting function	7
6. FIR filter and RF direction finder	8
7. Three non-zero tap weights	10
8. Pulse compression using a 4-bit Barker code	11
9. Schematic of first optical system to implement positive and negative weights	12
10. Fiber-optic transversal filter implemented with positive and negative weights without duplicate weight sets	13
11. Modulation characteristics of IOC	13
12. Pulse compression using a 7-bit Barker code	14
13. Optoelectronic differential amplifier schematic	15
14. Two-tone distortion	16
15. Common mode rejection ratio measured for two different time differences between the two input paths	17

TABLES

1. Weight assignments	12
2. Wide-band filter characteristics	18

INTRODUCTION

Independent Exploratory Development project (ZE 80) was directed toward developing a new technology capable of processing wide bandwidth signals. This capability is important to the Navy for use in communications systems using high data rates, secure communications, RF signal processing, beam forming, and handling complex signal environments in electromagnetic warfare (reference 1). These signal processing problems are currently being dealt with by digital and analog electronic systems. Digital electronic systems are flexible and can be quickly changed to react to changing signal processing needs. However, digital electronic filters are only used for signals with bandwidths of less than 100 MHz. The signal bandwidth that a digital electronic filter can process is limited by the state-of-the-art in analog-to-digital converters and by the speed of math coprocessors. The fastest 4-bit analog-to-digital converter demonstrated operates at 1 Gsample/s (reference 2). When the signal of interest has a bandwidth exceeding the capability of digital electronic filters, analog electronic devices are used. Analog electronics can handle wide signal bandwidths. The disadvantage of using analog electronic filters is that they are static. After their components (microstrips, waveguides, etc.) are built, analog filters are confined to their original configuration.

The limitations imposed by the state-of-the-art in electronic filters have restricted the Navy's capability to process wide bandwidth signals. There is a need for a flexible, realtime, wide bandwidth, signal processing system. For example, fixed, analog matched filters for active radar systems are subject to jamming and deception. A flexible, wide-band matched filter is necessary to selectively switch between filter configurations to prevent interference. Variable wide-band filter technology could also be applied to wide-band radar. Impulse radar systems are currently being planned that will require signal processing at a sampling rate greater than 5 Gsamples/s in realtime. This is not possible with the anticipated developments in electronic technology. Another need is in the area of electromagnetic warfare where an active adaptive wide-band filter could be used to notch out electromagnetic interference from agile jammers. Fiber optics is currently used in secure communication systems because it is not susceptible to electromagnetic interference. The complexity of the encryption schemes and the signal bandwidth used in encryption are limited by the sampling rate. A flexible, high-sampling-rate optical encryption system will greatly increase link security.

The signal processing capability of filters designed to handle large bandwidth signals is limited by the filter sampling rate and the dynamic range of the filter response. Transversal filters with sampling rates greater than 100 MHz were built previously using fiber-optic delay lines (references 3, 4, 5). The usefulness of these systems was limited by the inflexibility of their tap-weighting schemes. Since previous fiber-optic filters had fixed tap weights, they had no advantage over analog electronic filters. The dynamic range of fiber-optic filters was significantly lower than that of the analog electronic filters.

FIBER-OPTIC FINITE IMPULSE RESPONSE FILTERS

We developed and tested a fiber-optic finite impulse response (FIR) filter for processing 1-GHz bandwidth signals (references 6, 7). The technology we developed could also be applied to other wide-band signal processing applications.

A generalized FIR filter consists of analog delay lines. Each delay line taps off a sample of the signal at a different time interval. The signal sample is weighted and recombined with the other signal samples. FIR filters are commonly used for matched filtering and pulse compression. We will show how they can also be used for null shifting and beam forming.

Fiber-optic delay lines can carry wide bandwidth (>200 GHz) signals with low loss. Problems with previous fiber-optic FIR filters were primarily associated with the tap weight implementation. Previous

optical weighting schemes were limited to fixed tap weight values and fixed tap locations. In other words, there was no way to adjust the incremental delay line lengths after the filter was built, and the tap weights were also permanent. The tap weights on previous optical transversal filters were always constrained to positive values. These fiber-optic transversal filters had no advantage over analog electronic filters.

FIR FILTER WITH VARIABLE WEIGHTS

In our design, we have dramatically increased the usefulness of optical transversal filters by incorporating variable, positive and negative, tap weights. We used fiber-optic delay lines because of their wide signal bandwidth capability. The tapped delay lines were weighted using integrated optical two-by-two Mach-Zehnder directional couplers (Integrated Optical Couplers (IOC)). By using various properties of the IOCs, we were able to provide variable weighting and to create the first technique for optically implementing negative weights.

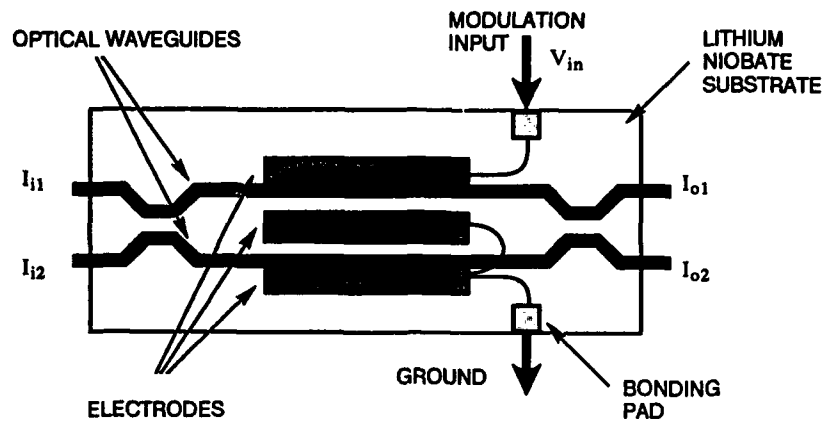
An example of an IOC is shown in figures 1a and 1b. The IOCs used in our experiments had two input fibers, I_{11} and I_{12} , and two output fibers, I_{01} and I_{02} . One fiber-optic input, I_{11} , was used to bring the signal of interest into the IOC. The IOC divided the signal between the fiber-optic outputs of the two waveguides, I_{01} and I_{02} . The electronic bias voltage controlled the ratio of the signals out of I_{01} and I_{02} . Only one optical output of the IOC tap weights, I_{01} , was used to weight the tapped delay lines. The DC bias and RF components of V_{in} were applied to the IOC through a bias-tee. The DC bias voltage could be used to attenuate, or weight, the signal in I_{11} . The IOCs could also be used as optical modulators. An RF voltage applied to the IOC modulated the intensity of the optical signal out of I_{01} . The IOCs used in our experiments could modulate signals at up to 3 GHz. Other IOCs are available off-the-shelf with modulation bandwidths up to 18 GHz.* When an IOC was used as an optical modulator, the DC bias voltage determined the modulation linearity.

Our filter was built using the design shown in figure 2. An IOC (XMOD) was used to externally modulate the output of a carrier wave (cw) 1.3- μ m laser diode (LD). The RF modulation frequency and power were controlled by a Hewlett Packard network analyzer. The modulated light was tapped into seven optical fiber delay lines via a 3-dB coupler and 1×4 trees. The fiber-optic delay lines were cut so the differences in length were as near as possible to 10 centimeters. The relationship between the delay length increment, ΔL , and the effective sampling frequency, f_s , is

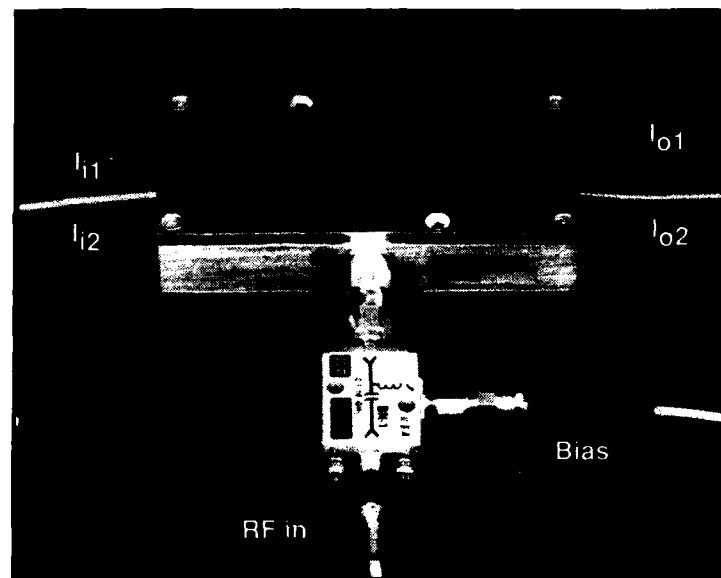
$$f_s = \frac{c}{n_g \cdot \Delta L} \quad (1)$$

For the fiber group velocity, n_g , a ΔL of 10 cm corresponded to an effective sampling rate of 2.02 Gsamples/s. The signal in each tapped delay line of the filter was weighted by an IOC. Only one input and one output of each IOC was used for this application. The intensity of the light passed by each IOC was controlled by a DC bias voltage. The appropriately weighted and delayed intensity signals were collected by an asymmetric star coupler and 3-dB coupler and directed to an InGaAs avalanche photodetector (DET). The attenuations of the IOCs were normalized with respect to each other. The signal at the detector was the incoherent sum of the optical intensities from each tap.

*Marconi Research Centre, Chelmsford, U.K.



(a) THE SUBSTRATE IS LiNbO_3 WITH TITANIUM-DIFFUSED WAVEGUIDES.



(b) PACKAGED FORM OF IOC SHOWING BUNDLED FIBERS AND BIAS-TEE.

Figure 1. Integrated optical coupler (IOC).

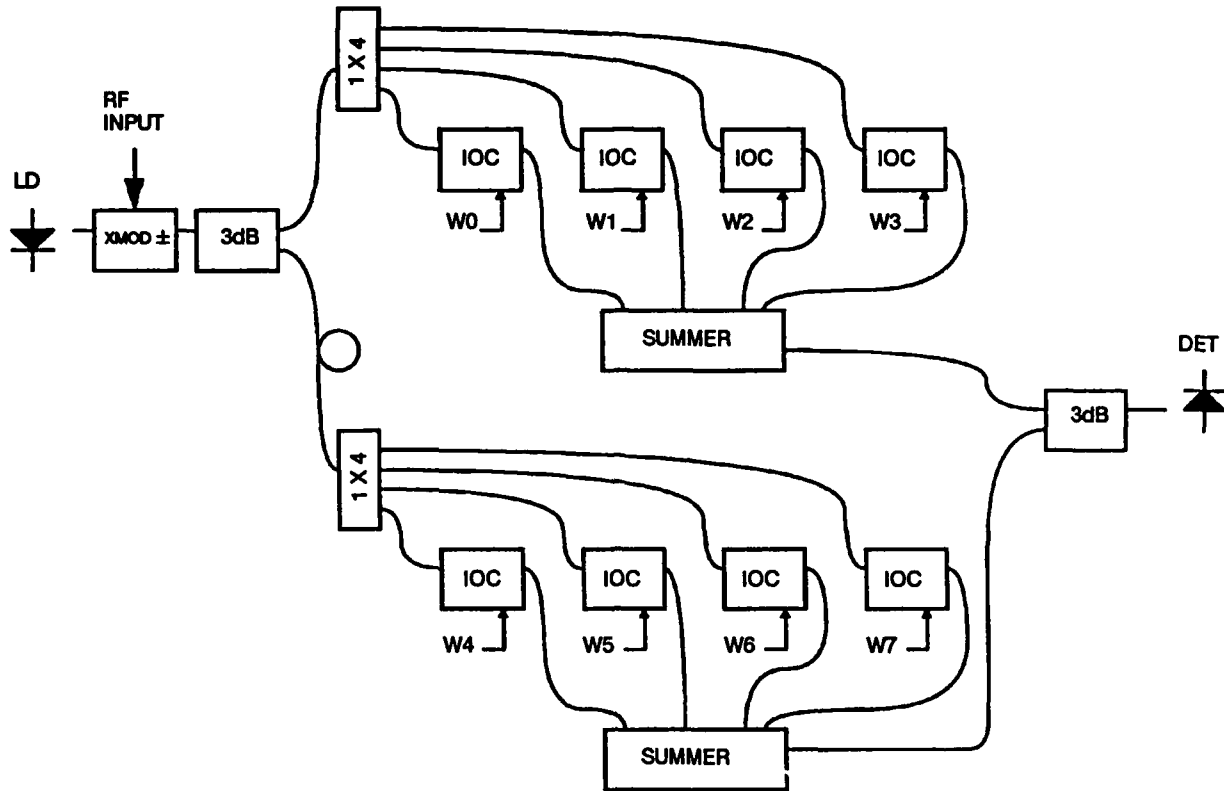


Figure 2. Fiber-optic FIR filter. (Laser diode (LD), external modulator (XMOD), integrated optical coupler (IOC), 1×4 -star coupler, 3 dB- 1×2 coupler, asymmetric star coupler (SUMMER), photodetector (DET))

The response function of this filter is given by

$$Y(t) = \sum_{k=0}^{N-1} W_k \cdot X(t - k\tau_s) . \quad (2)$$

X is the electrical input signal power, and Y is the filtered output signal power. W_k is the value of the k^{th} tap weight, N is the number of taps, and τ_s is the sampling period.

Unlike other optical systems where signal processing is done by modifying the amplitude of the light signal, this system used only the intensity of the light. The processing was done incoherently at the optical carrier frequency, but was coherent at the RF modulation frequency. Because the external modulator was operated in the small signal regime, the relationship between the electrical RF input power and the intensity of the light in the filter was linear, and the conversion of the light intensity at the filter output to RF electrical power was linear. Therefore, a 10-dB change in electrical RF signal power into the filter resulted in a 10-dB change in the electrical power out of the filter. In previous optical signal systems, the dynamic range was reduced by a factor of two (i.e., 80 dB to 40 dB) because the conversion from the optical signal to the RF electrical output followed a square-law relationship. Since our system has a linear input to output relationship, we have dramatically improved the dynamic range over previous systems.

Figure 3 shows the fiber-optic FIR filter system.

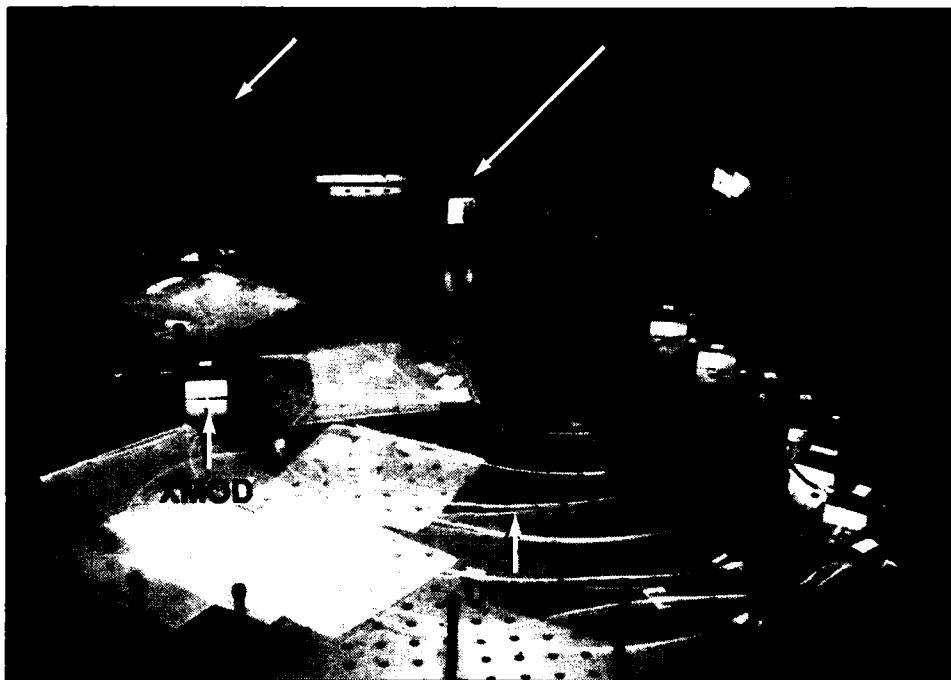


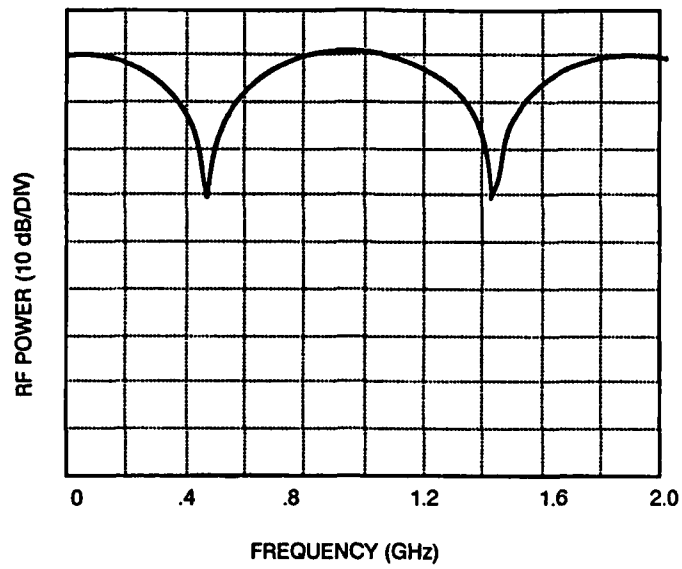
Figure 3. Fiber-optic FIR filter system. The light from the laser diode (LD) is externally modulated by the IOC (XMOD). The signals are delayed by the fiber (DELAY), weighted by the IOCs, and collected before the photodetector (DET).

A 0- to 2-GHz electrical signal was used to characterize the filter response. The signal bandwidths of all the passive optical components were much greater than the 2-GHz bandwidth used. We used the storage capability of the network analyzer to correct for the rolloff in the modulator/detector response. As a result, we could examine the actual response of the filter. Examples of the corrected response of the transversal filter are shown in figures 4a and 4b and figure 5. The weights were chosen for figures 4a and 4b such that the frequency response was similar but the null was shifted 60 MHz. The tap weights were set to a raised cosine weighting function to achieve the frequency response shown in figure 5.

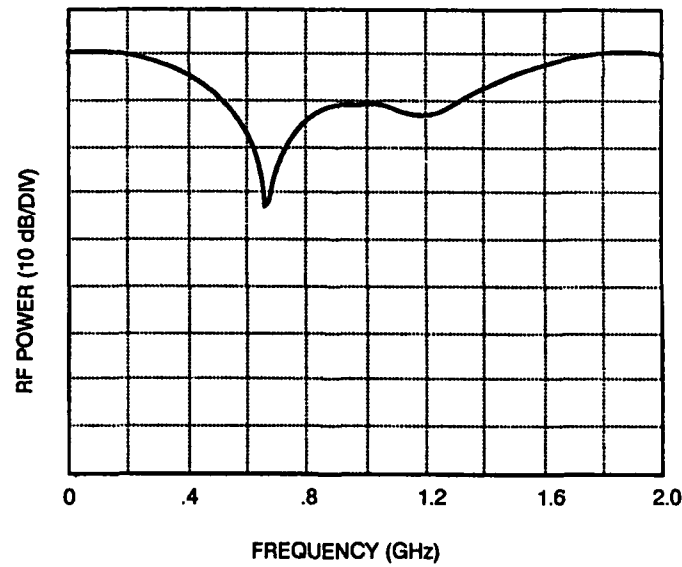
The depth of the null in the filter response was determined by the dynamic range of the filter. Our filter dynamic range was greater than 70 dB for a 10-kHz IF bandwidth. This corresponds to 110 dB normalized to a 1-Hz noise bandwidth. Since the laser diode was modulated externally, the dynamic range was limited primarily by the detector noise. Although the signal bandwidth of a transversal filter depends on the sampling period and the bandwidth of the RF input and output devices, the bandwidth of a *fixed tap weight* fiber-optic transversal filter is critically limited by the fiber cut accuracy (reference 8). By using variable weights, we were able to modify the filter response to correct for errors in the fiber lengths.

NULL SHIFTING AND BEAM FORMING

Null shifting in the frequency domain is directly related to radar beam forming. In the past, several efforts have been made to take advantage of the characteristics of fiber optics to solve some phased array radar problems. For example, fiber-optic delay lines were used in phased array radar to



(a) FILTER FREQUENCY RESPONSE TO AN ARBITRARY WEIGHTING FUNCTION.



(b) THE FIRST NULL IN THE FREQUENCY RESPONSE WAS SHIFTED BY CHANGING THE WEIGHTING FUNCTION.

Figure 4. Examples of corrected response of the transversal filter.

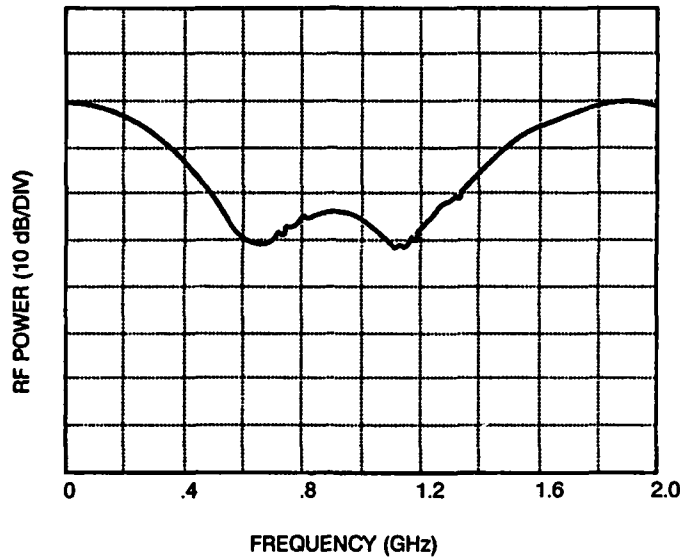


Figure 5. Frequency response using a raised cosine weighting function.

do direction finding (references 8, 9). Usually the angular response of the antenna array was controlled by the delay line lengths. Recently, some groups have developed new techniques for varying the delays (references 10, 11). Although changing the delay lines could change the angle of peak response of the direction finder, they could not change the shape of the response or the relative location of nulls.

The performance of the fiber-optic FIR filter we built and tested is directly related to the performance of a radar direction finder or beam former. Frequency null shifting in the filter corresponds to changing the nulls in the angular response of a direction finder. A diagram of a fiber-optic FIR filter and a diagram of a fiber-optic direction finder are shown in figures 6a and 6b. Both systems consist of weighted fiber-optic delay lines. The difference in the delay line lengths, Δl , determines the sampling frequency, f_s , of the filter, or the angle of arrival for peak response, θ_0 , for the direction finder. The relationship between these quantities is given in

$$\Delta l = \frac{c}{f_s \cdot n_g} = \frac{d \cdot \sin(\theta_0)}{n_g} \quad (3)$$

where n_g is the group index of the optical fiber, d is the separation between antenna elements, and τ_s is the sampling period, $1/f_s$. The major difference between the construction of the two systems is that the FIR filter has only one signal input, whereas the direction finder has a separate input from each element of the antenna array.

The variable weights in this FIR filter made it possible to shift the frequency of the nulls in the filter response. In most fiber-optic and analog systems, null shifting is achieved by changing the delay between taps or by changing the phase of the tap weights. However, null shifting can also be done with real weights as was shown in figures 4a and 4b.

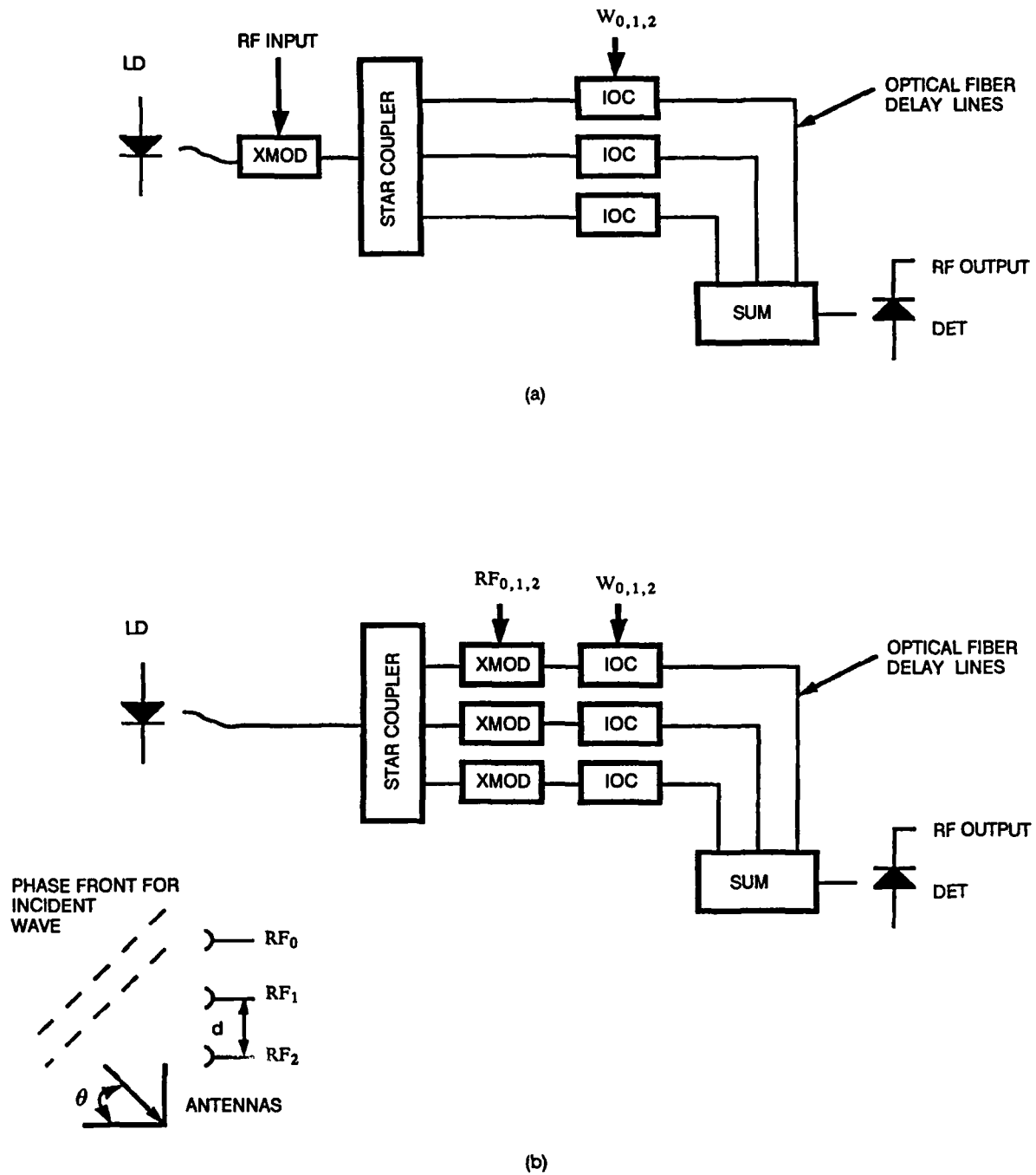


Figure 6. FIR filter and RF direction finder.

The transfer function of an N tap transversal filter (in the frequency domain) is given by the expression

$$H(f) = \omega_0 + \omega_1 e^{-i2\pi f \tau_s} + \omega_2 e^{-i2\pi 2f \tau_s} + \dots + \omega_{N-1} e^{-i2\pi (N-1)f \tau_s} \quad (4)$$

where ω_i are dummy weights. In general, the weights can take on real or imaginary values where imaginary values represent phase shifts. In a fiber-optic FIR filter, the weights are all real. As an example, we will describe how the weights are chosen to set one null for a three-tap filter. The calculations for larger filters are similar. When the number of filter taps, N , is odd, then $(N-1)/2$ nulls can be independently controlled. If N is even, the one null is always fixed at $f_s/2$ and the number of independently controlled nulls is $(N-2)/2$. To find the weight values that will give real nulls at any arbitrary frequency, f_n , we need to solve equation 4 for $H(f_n) = 0$. Since we do not want any solutions that require complex weights, we separate equation 4 into two different equations:

$$0 = \omega_0 + \omega_1 \cos(x) + \omega_2 \cos(2x) \quad (5a)$$

$$0 = \omega_1 \sin(x) + \omega_2 \sin(2x) \quad (5b)$$

In these equations $x \equiv 2\pi f_n \tau$. Only the relative values of the weights are important, so we can set one of the weights to an arbitrary constant to scale the solution. We choose to set $\omega_2 = 1$. Then the solutions are

$$\begin{aligned} \omega_0 &= 1 \\ \omega_1 &= -\frac{\sin(2x)}{\sin(x)} \\ \omega_2 &= 1 \end{aligned} \quad (6)$$

The absolute value of the weights cannot exceed 1; hence, we must normalize the weights separately for each null frequency we select. It is important that the normalizations result in allowing as much light as possible to go through the filter because then the dynamic range of the filter will be maximized. The normalizations are calculated using $MAX = \text{maximum} \{ |\omega_0|, |\omega_1|, |\omega_2| \}$. The new weights are

$$\begin{aligned} W_0 &= \frac{\omega_0}{MAX} \\ W_1 &= \frac{\omega_1}{MAX} \\ W_2 &= \frac{\omega_2}{MAX} \end{aligned} \quad (7)$$

These solutions are plotted versus desired null frequency in figure 7. We find that with only three filter taps, the null can be shifted over the entire frequency bandwidth using positive and negative tap weights. With only positive tap weights, we can shift the null from one-fourth the sampling frequency to three-fourths the sampling frequency. We checked the theoretical results experimentally using only three taps of our FIR filter. We set the first and third tap to $W_0 = W_2 = 1$, and we increased the middle tap weight from $W_1 = 0$ to $W_1 = 1$. The experimental data are shown in the frequency range from 500 to 667 MHz in figure 7.

We have demonstrated how a null can be placed at any arbitrary frequency by controlling the weights of a three-tap filter. Similarly, two nulls can be placed at two different arbitrary frequencies with a five-tap filter, etc. Because the weights can be adjusted very rapidly, nulls in the direction finder response will shift in realtime. Filter null shifting is directly analogous to null steering in a radar direction finder. It is not possible to change the angle of peak response of a radar direction finder with fixed delays.

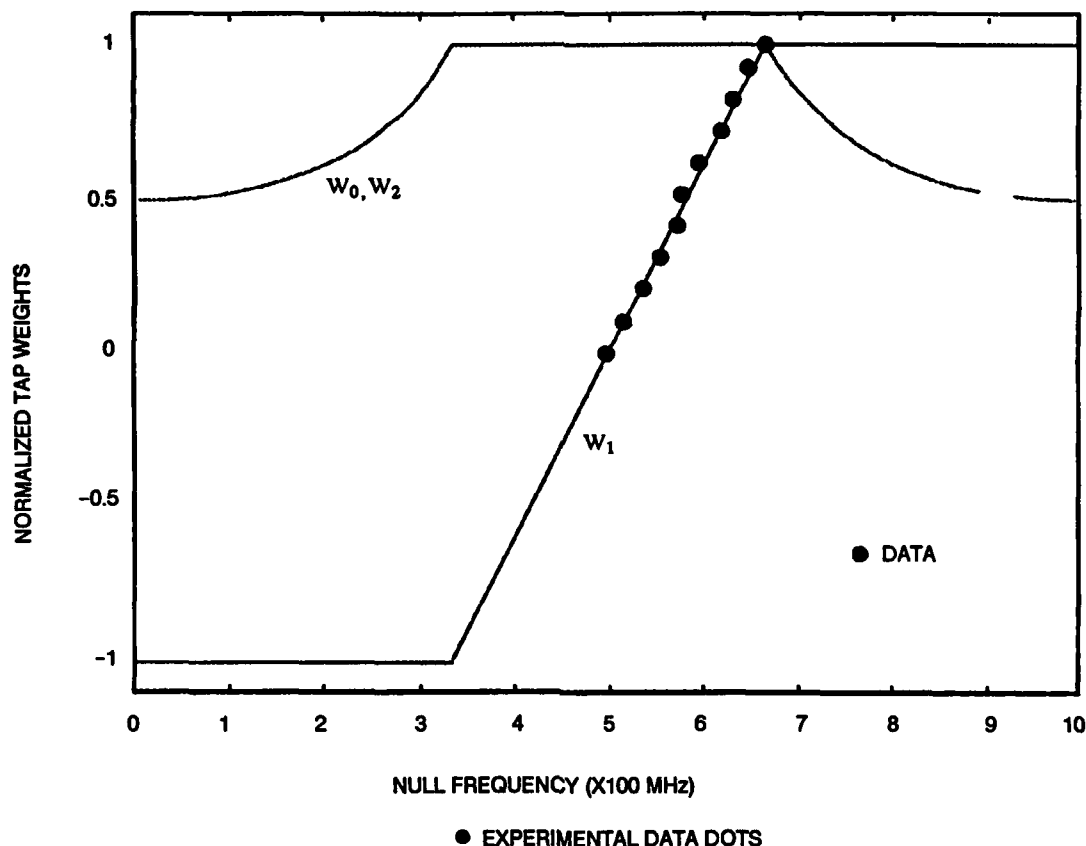


Figure 7. Three non-zero tap weights (W_0 , W_1 , and W_2) will shift the first null of the filter frequency response across the entire bandwidth.

Beam forming and direction finding can be performed on high bandwidth radar signals using fiber-optic delay lines. By using variable weights, we have circumvented the usual limitations imposed on direction finders with fixed delays. The angle of peak response of the direction finder is determined by the delay lines, whereas the null angles depend on the weights. A change in null frequency in the transversal filter response is related to the change in the angle of the direction finder null. Variable weights can be used to sweep a null in the angular response of a direction finder through 180° . The direction finder delay lines can be used to change the angle of peak response. The variable delay techniques developed by other groups (references 10, 11) can be used in concert with our variable weight system to provide complete control over the direction finder response.

POSITIVE AND NEGATIVE WEIGHTS

The tapped delay lines in the filter implementation described only had positive weights. This filter could do low pass filtering, limited null steering, and related tasks. More extensive filtering operations such as pulse compression (see figure 8), phase encoding, signal time differentiation, unlimited null steering, matched filtering, and waveform generation require both positive and negative weights (reference 12). Therefore, a technique for incorporating negative weights into the filter was developed. The function of negative weights is to subtract a delayed sample, X_k , of the RF signal from the rest of the signal samples. This is equivalent to inverting the phase of X_k with respect to the original RF input signal.

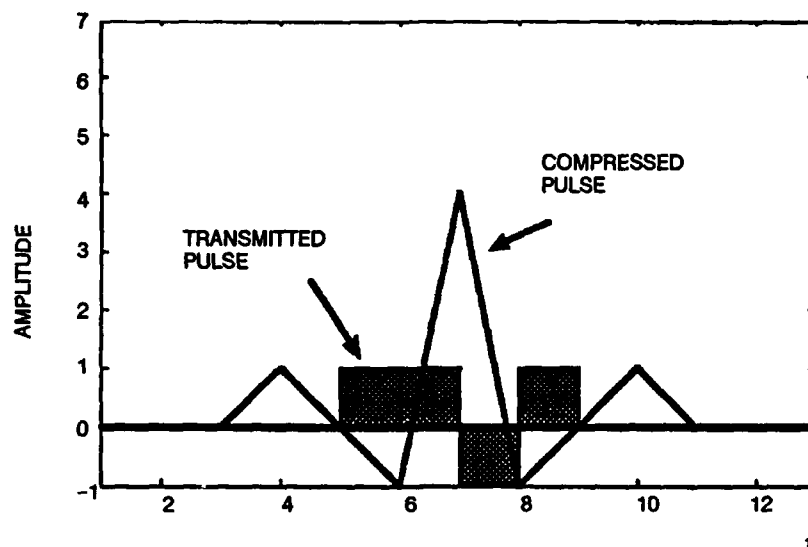


Figure 8. Pulse compression using a 4-bit Barker code. This length code gives a 12.0-dB peak-to-sidelobe ratio.

A general theoretical transversal filter can modify the frequency and phase response of a signal continuously. However, all electronic filters, analog or digital, have some constraints. The phase response of a typical digital filter can be modified within the limits imposed by the digitization (4 bits yields $11^\circ/\text{bit}$).

Previous optical filters did not have negative weights. Negative weights are implemented in electronic filters by changing the sign of the voltage of the appropriate signal sample and adding the voltages ($-X_k + X_0$). This type of negative weight implementation cannot be done optically since it is not possible to subtract light intensities.

We have implemented negative weighting in our fiber-optic FIR filter by inverting the phase of the RF-modulated optical signals in the appropriate tapped delay lines. The optical output of an IOC can be in-phase or out-of-phase with respect to the modulating RF signal. The modulated optical intensity out of I_{01} is 180° out-of-phase with respect to the output of I_{02} . This property of the IOC was used to produce two out-of-phase signals out of XMOD. The optical output of I_{01} was sent into one filter, and the output of I_{02} was sent into another identical filter. A schematic of the optical FIR filter that used this positive and negative weight implementation is shown in figure 9.

As an example, the filter in figure 9 can be used to perform the pulse compression operation illustrated in figure 8. The most popular type of radar signal phase coding is binary; that is, the pulse amplitude is either +1 or -1. This maximizes the power out of the transmitter. The weighting function in figure 8 is the Barker code for four elements: $W_{0-3} = +1, +1, -1, +1$. This weighting function can be implemented using the filter in figure 9 by setting the weights on the tapped delay lines as shown in table 1.

To have a completely flexible weighting capability where every weight could be either positive or negative, two identical filters must be built. One filter would be used for each sign of weight. This is a drawback of this implementation because it doubles the number of IOCs required. If the same tapped

Table 1. Weight assignments.

In-Phase	Out-of-Phase	Equivalent Weight
$\omega_0^+ = 1$	$\omega_0^- = 0$	$\omega_0 = +1$
$\omega_1^+ = 1$	$\omega_1^- = 0$	$\omega_1 = +1$
$\omega_2^+ = 0$	$\omega_2^- = 1$	$\omega_2 = -1$
$\omega_3^+ = 1$	$\omega_3^- = 0$	$\omega_3 = +1$

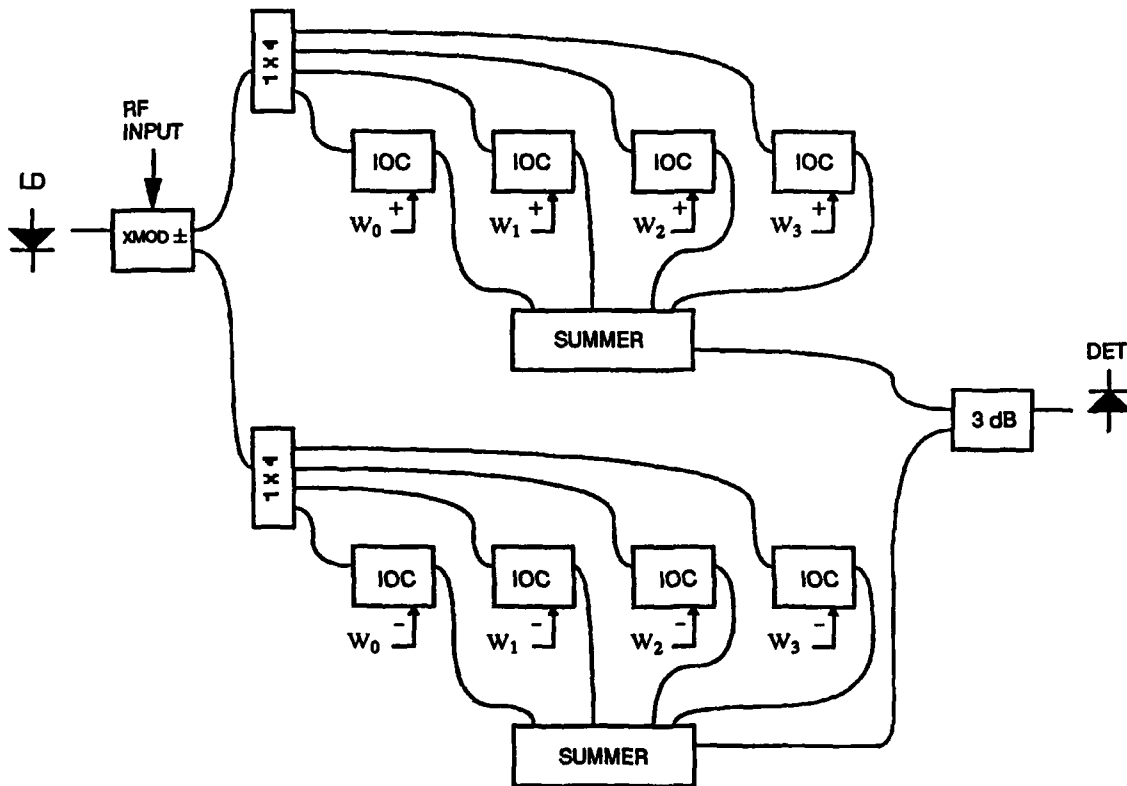


Figure 9. Schematic of first optical system to implement positive and negative weights.

delay lines were always going to have one sign of weight, then some of the duplication can be eliminated. However, for the most versatile system, it is desirable to be able to implement positive or negative weighting at every delay increment. An alternative fiber-optic transversal filter design which eliminates the need for duplicate weight sets is shown in figure 10. According to this implementation, the phase of the optical signal at the selected output, I_{01} , is controlled by a DC bias voltage. This is shown schematically in figure 11. Under ordinary conditions, the DC bias is set to the linear region of the IOC response curve. The IOCs used in our experiments had a cosine squared response. IOCs with different modulation responses could also be used. The electronic control voltage consists of a DC

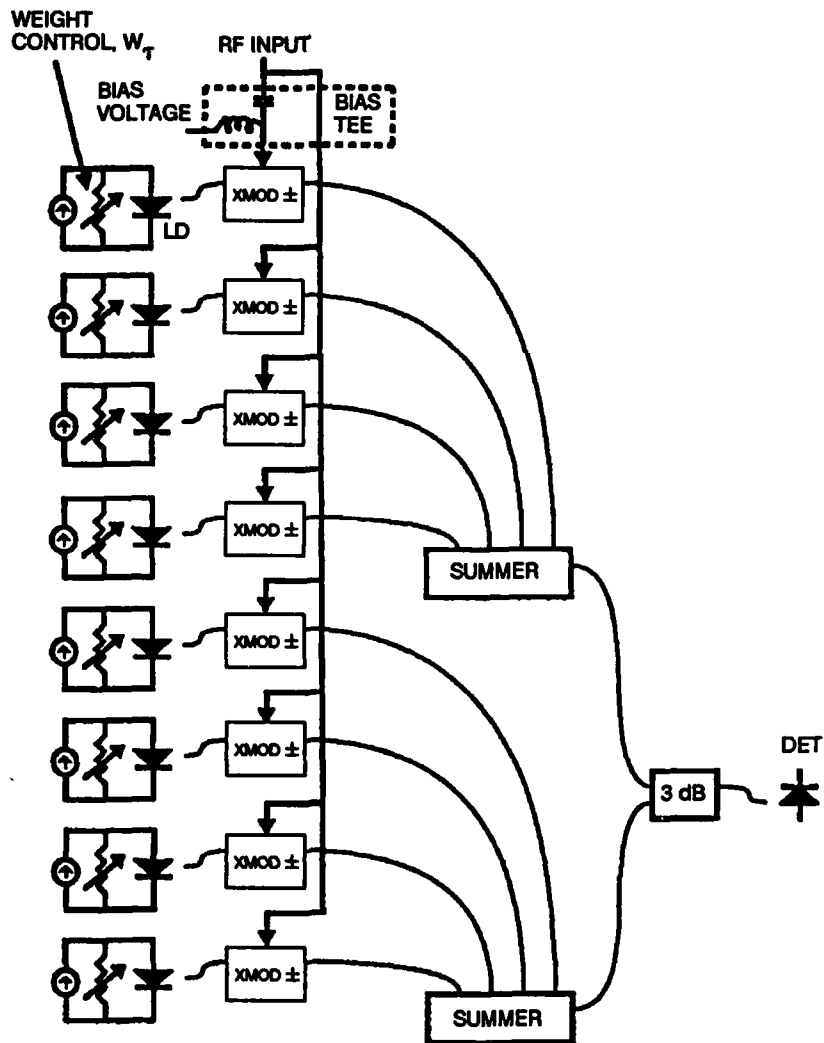


Figure 10. Fiber-optic transversal filter implemented with positive and negative weights without duplicate weight sets.

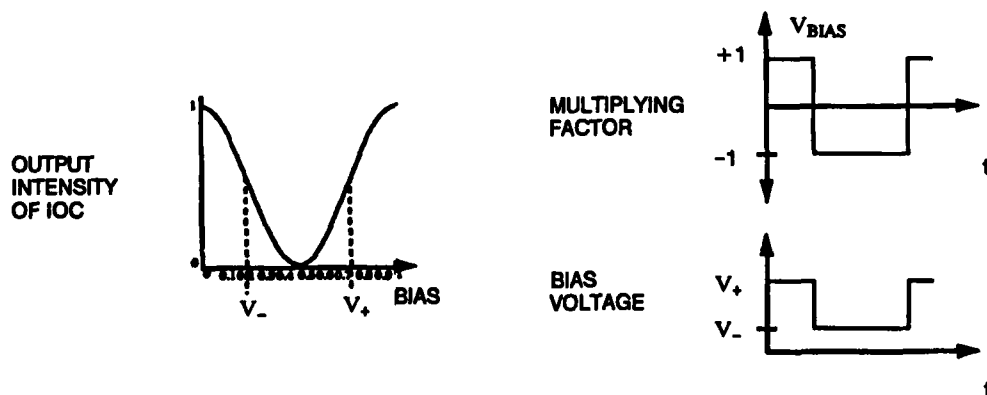


Figure 11. Modulation characteristics of IOC.

bias component and an RF signal component. The bias voltage can be adjusted so that the operating point of the IOC modulator is on the positive slope or the negative slope of the response curve as shown in figure 11. Consequently, the optical signal at I_{O1} can be in-phase or out-of-phase with respect to the electrical RF input signal. For the pulse compression example from figure 8, the bias voltages should be set to V_+ , V_+ , V_- , V_+ . Since this positive and negative weight implementation frees more IOCs for use as tap weights, a seven-element Barker code could be used instead of the four-element code. The use of a longer code results in a corresponding increase in signal-to-sidelobe ratio as shown in figure 12. The laser diodes should be adjusted to equal intensities to correspond to unity weights. For a phase encoding scheme which uses nonunity weights, the amplitude weighting is controlled by setting the laser diode intensities appropriately.

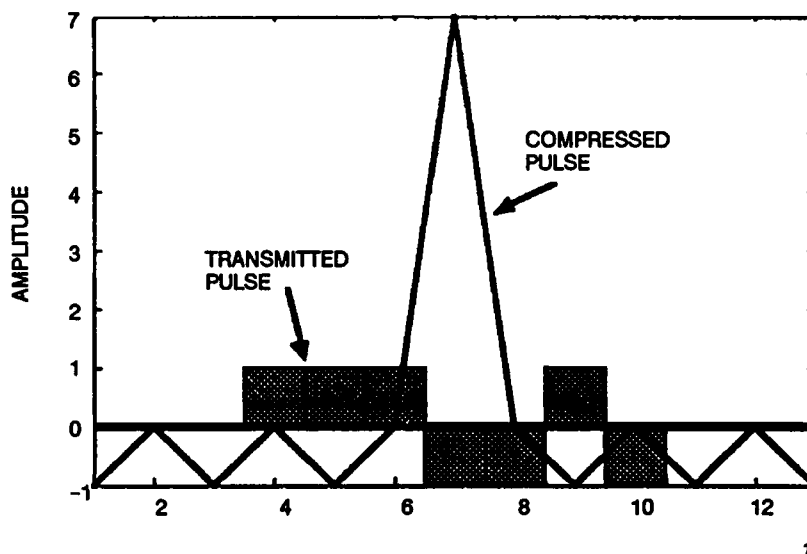


Figure 12. Pulse compression using a 7-bit Barker code. This length code gives a 16.9-dB peak-to-sidelobe ratio.

DIFFERENTIAL AMPLIFIER

The phase inversion technique we developed to implement positive and negative weights in our FIR filter can be used for other applications. We have invented a differential amplifier which works on signals with bandwidths greater than 1 GHz (references 13, 14).

A schematic of the optoelectronic differential amplifier is shown in figure 13. The light from a 1.3- μm cw laser diode was conveyed, via single mode fibers and a 3-dB coupler, through two IOCs, and recombined at an InGaAs detector. This is an AC coupled system. The IOCs are the most critical components of the differential amplifier. For this application, the IOCs were used as modulators. The electrical signals of interest $g(t)$ and $h(t)$, were applied to IOC-1 and IOC-2, respectively. The optical modulators were biased with voltages V_1 and V_2 to the linear portion of their response curve. Only one output port of each IOC was used. The modulated optical intensity out of port A of these IOCs is in-phase with respect to the RF input signal, and the intensity out of port B is 180° out-of-phase with respect to the RF input. The output ports of each IOC were chosen so that the intensity modulation corresponding to $g(t)$ was in-phase, and the intensity modulation from $h(t)$ was out-of-phase, with respect to the RF inputs. The summation of the two signals with an 180° phase shift was equivalent to subtracting $h(t)$ from $g(t)$.

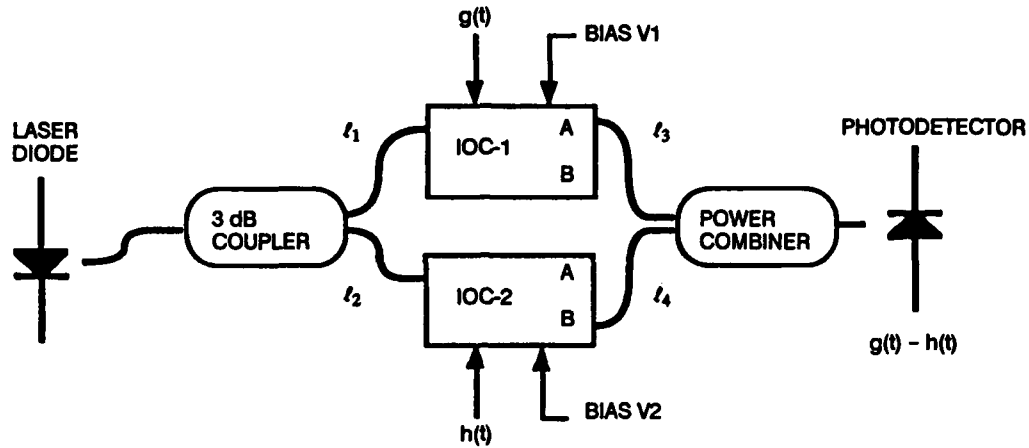


Figure 13. Optoelectronic differential amplifier schematic.

Differential amplifier operation requires that the two electrical signals be subtracted coherently at the electronic signal frequency. For this to occur, the two intensity-modulated signals must be summed incoherently at the optical carrier frequency while maintaining coherence at the electronic signal frequency. By summing the optical signals incoherently, we eliminated optical interference effects. In our experiment, we ensured that the signals summed incoherently by introducing a path difference longer than the laser coherence length (ℓ_c) into the system. This was done by satisfying

$$|\ell_1 - \ell_2| \gg \ell_c . \quad (5)$$

The lengths of the optical fibers out of the IOC's, ℓ_3 and ℓ_4 , must be equal for the signals to be subtracted simultaneously. The tolerance on ℓ_3 and ℓ_4 depends on the highest frequency component in the electrical signal, f_H .

$$|\ell_3 - \ell_4| \ll \frac{c}{f_H} . \quad (6)$$

Other approaches could be used to ensure that the optical signals are summed incoherently. For example, two different laser sources could be used, one injecting light into IOC-1 and one for IOC-2.

This differential amplifier has the signal properties of two analog optical communications links combined at the detector. The common mode rejection ratio (CMRR) can approach the dynamic range of the system. Our system had (in a 1-Hz bandwidth) a linear dynamic range of 134 dB, a minimum detectable signal of -157 dBm, spurious free dynamic range of 98 dB/Hz^{2/3}, and third-order intercept of -11 dBm (see figure 14). These values can be improved by choosing appropriate components.

The differential amplifier gain/loss depends on the loss and gain in the system. Loss is due to optical device insertion and coupling losses. Gain or loss depends on the $V\pi$ of the IOC: the smaller the $V\pi$, the larger the gain. Gain can be increased by increasing the intensity of the laser or by introducing optical amplifiers. Gains as high as 11 dB (without optical amplifiers) have been reported for

optical communications links (reference 15). Since the loss/gain of the two optical paths is not generally the same, it is advantageous to use the two-laser approach so the intensity incident on the detector from each path can be equalized. The transfer function of the IOCs does not vary significantly from device to device because it is a physical characteristic of the modulator, unaffected by manufacturing conditions. Therefore, it is possible to make the CMRR of the differential amplifier approach the dynamic range of the system by equalizing the gain of the two paths. The common mode extinction for very wide bandwidth signals will be less because the IOC frequency response over the multi-gigahertz bandwidth is device dependent.

For measurement purposes, we defined the CMRR as the ratio of the electrical power out of the differential amplifier for the case where an input signal, $g(t)$, was applied and $h(t) = 0$ to the case where the input signals, $g(t)$ and $h(t)$, were equal. The CMRR was measured at several frequencies from 100 MHz to 1.5 GHz. The results are shown in figure 15. We found that the CMRR depended strongly on the difference between the two signal path lengths. The relationship between the CMRR, the measurement frequency, f , and the time difference between the two paths, τ , is given by

$$CMRR < -20 \log[2 \sin(\pi f \tau)] . \quad (7)$$

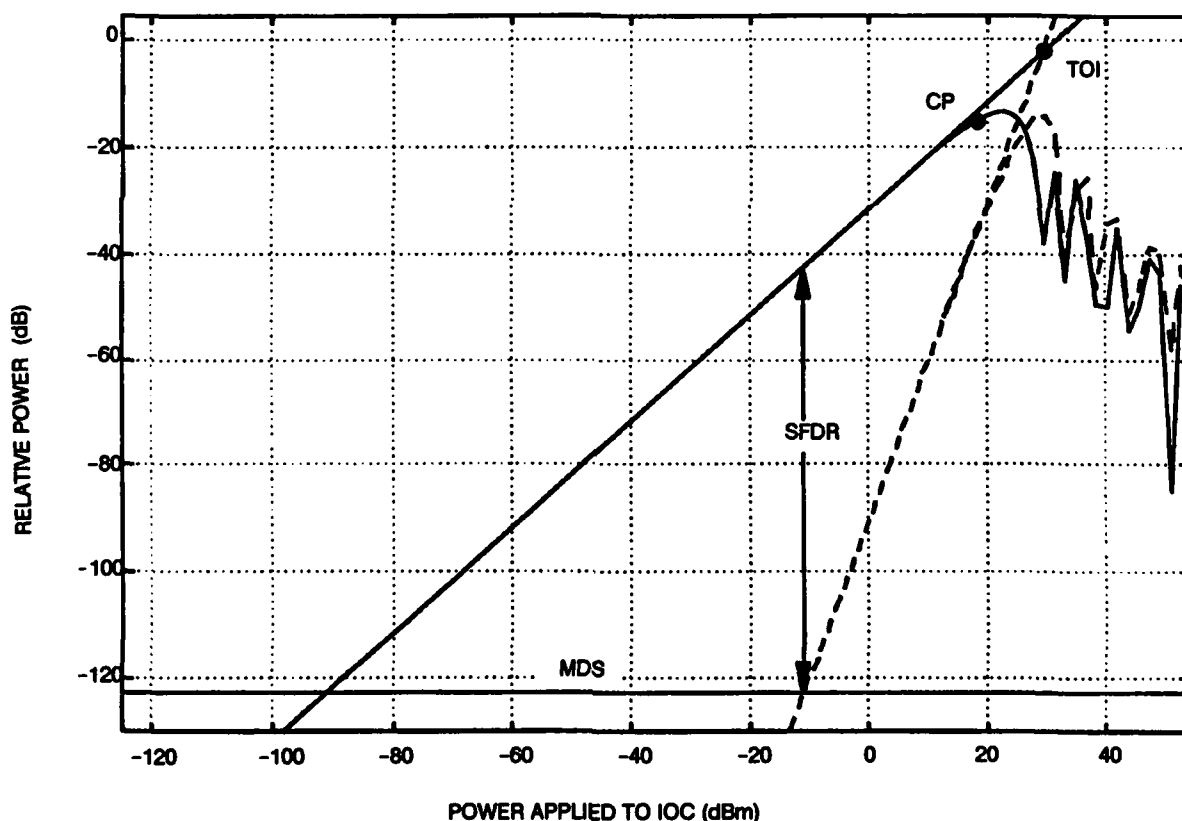


Figure 14. Two-tone distortion. (—fundamental power, - - -third order intermodulation power. CP (1-dB compression point) = 16.8 dBm, TOI (third order intercept) = 29.5 dBm, MDS (minimum detectable signal) = -122.8 dBm, SFDR (spurious free dynamic range) = 80.6 dB. Noise bandwidth = 3000 Hz. IOC input RF to detector RF output loss = 31.7 dB, $V_{\pi} = 12$ volts, load impedance = 50 ohms.)

Data A in figure 15 is shown with a theoretical curve for $\tau = 58$ ps. An effort was made to reduce the time discrepancy. Data B is shown with the theoretical curve for $\tau = 4$ ps. Our data shows CMRR >30 dB at 1 GHz. According to this theory, CMRRs greater than 50 dB could be obtained for 1-GHz signal bandwidths if the path differences are less than 0.5 picosecond. This could be done easily by putting the two IOCs together in an integrated optic differential amplifier module. For very small path differences, the CMRR is limited by the spurious free dynamic range of the differential amplifier. We predict that the largest CMRR we could obtain at 1 GHz with our system would be 70 dB. This is for -10-dBm input signals and output path length differences less than 50 femtoseconds.

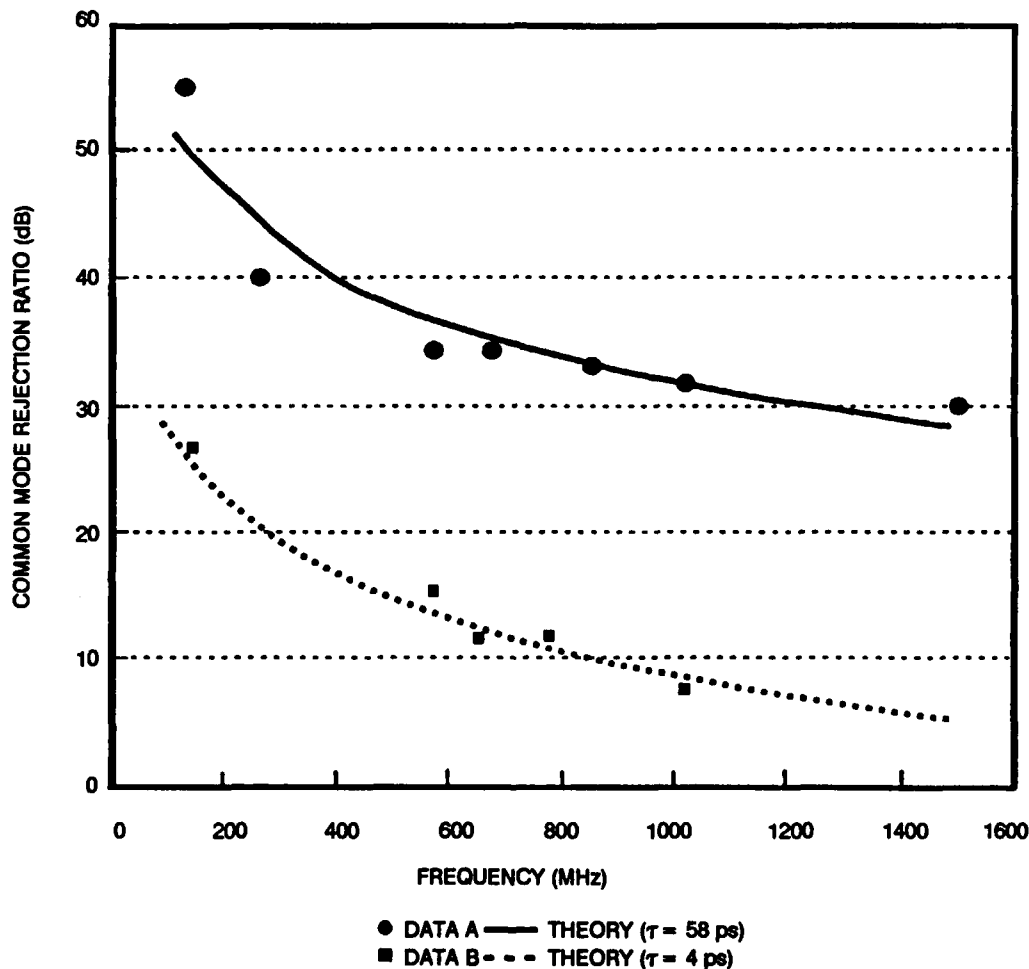


Figure 15. Common mode rejection ratio measured for two different time differences between the two input paths.

SUMMARY

We have developed a wide-band signal processing system. The fiber-optic FIR filter we built works in realtime, can handle wider bandwidth signals than digital electronic filters, and has more versatility than analog electronic filters or previous optical filters. Table 2 lists the important characteristics for a wide bandwidth filter and a corresponding list of our accomplishments. The weights can be controlled electronically and can be reset rapidly to react to changing signal environments. This fiber-optic transversal filter is suitable for use in an adaptive signal processing system or neural network.

Table 2. Wide-band filter characteristics.

Properties	Achievements	Future Potential
Increased signal bandwidth	1 GHz	10 GHz
High sampling rate	2 Gsample/s	10 Gsample/s
Large dynamic range	110 dB (1 Hz BW)	140 dB (1 Hz BW)
Capability to change weights		
Amplitude	yes (40 dB)	yes (>30 dB)
Phase	yes (0°, 180°)	—

By incorporating variable, positive and negative weights into the filter, we were able to demonstrate a variety of signal processing techniques on wide bandwidth signals. The fiber-optic transversal filter will do pulse compression, phase encoding, signal time differentiation, unlimited null steering, matched filtering, and waveform generation. These filter operations can be used to generate, compress, and encode radar pulses at the transmitter, and to do beam forming and matched filtering at the receiver. The speed and high sampling rate of the fiber-optic transversal filter are particularly suited to impulse radar applications because of the short pulse length and high sampling rates required. The filter is reconfigurable so the encoding schemes can be changed from pulse to pulse to prevent interception and deception. Conversely, since the filter works in realtime, it can be used to deceive enemy aircraft effectively.

In addition to radar applications, the filter can be used to identify and categorize signatures in electronic warfare. The filter can be incorporated into an adaptive noise canceler to follow and null out agile jammers. Fiber-optic carriers are useful in electronic warfare scenarios because they are not susceptible to electromagnetic interference. Fiber-optic links are used for remote antenna positioning. The fiber-optic FIR filter is useful for processing signals already in optical form.

The signal processing technology we have developed for the fiber-optic FIR filter also applies to other devices. We have also developed a crossbar switch for communications networks that uses fiber optics and IOCs to do time-division signal multiplexing (reference 16). We have demonstrated a new optoelectronic wideband differential amplifier. Unlike previous optoelectronic systems, the signal subtraction was done in the optical regime. This design does not require careful matching and tuning as in electronic wideband differential amplifiers or balanced circuit configurations. This differential amplifier implementation takes advantage of recent developments in low-noise, large-dynamic-range, analog fiber-optic communications systems. Custom device development could improve the differential amplifier performance, but it is not necessary. This differential amplifier module could be a useful component in radar and communication systems.

We have significantly increased the variety of defense strategies available to the Navy by demonstrating a gigahertz bandwidth signal processing system with a large dynamic range. The techniques for using fiber-optic interconnects and delay lines and IOCs for weighting and switching wide bandwidth signal developed in this project can be used in many Navy applications. There are applications where fiber-optic FIR filters can be substituted for existing equipment. For example, the filter can be substituted for acousto-optic Bragg-cell matched filters in existing radar systems. More importantly, however, the variable weight wide bandwidth capability of the filter opens up new possibilities in other areas such as in phased array radar beam forming and secure communications.

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REPORT DOCUMENTATION PAGE

Form Approved
OMB No. 0704-0188

Public reporting burden for this collection of information is estimated to average 1 hour per response, including the time for reviewing instructions, searching existing data sources, gathering and maintaining the data needed, and completing and reviewing the collection of information. Send comments regarding this burden estimate or any other aspect of this collection of information, including suggestions for reducing this burden, to Washington Headquarters Services, Directorate for Information Operations and Reports, 1215 Jefferson Davis Highway, Suite 1204, Arlington, VA 22202-4302, and to the Office of Management and Budget, Paperwork Reduction Project (0704-0188), Washington, DC 20503.

1. AGENCY USE ONLY (Leave blank)		2. REPORT DATE December 1990		3. REPORT TYPE AND DATES COVERED	
4. TITLE AND SUBTITLE WIDE-BAND RF SIGNAL PROCESSING WITH OPTOELECTRONIC DEVICES AND FIBER-OPTIC DELAY LINES				5. FUNDING NUMBERS PE: 0602936N WU: DN309059	
6. AUTHOR(S) M. H. Berry, D. M. Gookin					
7. PERFORMING ORGANIZATION NAME(S) AND ADDRESS(ES) Naval Ocean Systems Center San Diego, CA 92152-5000				8. PERFORMING ORGANIZATION REPORT NUMBER NOSC TR 1385	
9. SPONSORING/MONITORING AGENCY NAME(S) AND ADDRESS(ES) Office of Chief of Naval Research Independent Exploratory Development Programs Arlington, VA 22217				10. SPONSORING/MONITORING AGENCY REPORT NUMBER	
11. SUPPLEMENTARY NOTES					
12a. DISTRIBUTION/AVAILABILITY STATEMENT Approved for public release; distribution is unlimited.				12b. DISTRIBUTION CODE	
13. ABSTRACT (Maximum 200 words) This report discusses a new technology for processing wide-band, >1 GHz, rf signals with a large dynamic range. Fiber-optic delay lines and optoelectronic devices were used to build and test a finite impulse response (FIR) filter and a wide-band differential amplifier. Other applications, including radar beamforming and pulse compression are described. These devices work in realtime.					
14. SUBJECT TERMS wide bandwidth signal processing fiber-optic finite impulse response filters					15. NUMBER OF PAGES 26
					16. PRICE CODE
17. SECURITY CLASSIFICATION OF REPORT UNCLASSIFIED	18. SECURITY CLASSIFICATION OF THIS PAGE UNCLASSIFIED	19. SECURITY CLASSIFICATION OF ABSTRACT UNCLASSIFIED	20. LIMITATION OF ABSTRACT SAME AS REPORT		

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