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Wide-Band Signal Processing Using the Two-Beam Surface Acoustic Wave Acoustooptic Time Integrating Correlator

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***Abstract*— A new acoustooptic architecture for performing real-time correlation of high-frequency wide-band signals has been developed. It uses a surface-acoustic-wave (SAW) delay line, and features the optical inteference of two coherent light beams which have been Bragg-diffracted by SAW's propagating in the line. The signal multiplication, and subsequent time integration of the product formed, is performed by a photodiode array which detects the diffracted light. This architecture has achieved time-bandwidths products exceeding 10^6 (34 MHz$\times$30 ms), and has several attributes which make it particularly well suited for use as a spread-spectrum signal processor. These include linearity of operation, large dynamic range, a large time aperture over which the correlation can be observed, and the ability to determine the center frequency and bandwidth of the signals. A correlator with this architecture has been used to detect a number of wide-band spread-spectrum signals. Its suitability for use as a signal processor in several spread-spectrum systems is considered.**

I. Introduction

SEVERAL present-day radio systems used for communications, navigation, and radar make use of spread-spectrum techniques to obtain greater range, multiple user access, better range resolution, lower probability of intercept, and improved antijam capability [1, ch.1]. The transmitted signal in these systems is spread over a frequency band that is much larger than is necessary for the information being sent. In most systems of interest, this bandwidth spreading is accomplished by using a special signal in addition to the information being sent to modulate the radio carrier. At the receiver, a correlation demodulation process "collapses" the excess bandwidth, obtaining the improved performance in the system. The wide-band signals employed in these systems present special problems in real-time signal detection and reception, problems which a large time–bandwidth product correlator may help solve.

Current digital processors are too limited in bandwidth to successfully perform real-time correlation of many of these spread-spectrum signals. The "traditional" analog correlator (which consists of a mixer-multiplier followed by an operational amplifier integrator) can process long duration wide-band signals. However, its output represents the correlation of the signals at only one relative time delay between them. It is very desirable to have a correlator which can simultaneously perform the correlation of the signals over a range of relative time delays, thereby sim-

Manuscript received February 15, 1980.
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plifying signal acquisition and identification. Real-time analog convolvers which use ultrasonic delay lines [2], [3] can obtain correlation of wide-band signals over a range of relative time delays. However, these devices are limited to signal durations of a few tens of microseconds because the correlation integration is performed over a spatial variable in the convolver, which limits performance of these devices.

Several schemes for combining ultrasonic delay lines with some form of time integration have resulted in correlators which can process long-duration wide-band signals over a range of relative time delays. Two general types of delay line time-integrating correlators have been developed: those which use an acoustoelectric interaction [4] and those which are based on an acoustooptic interaction [5]–[7]. The acoustoelectric time-integrating correlator is a new device and has not attained the integration time and dynamic range of several of the acoustooptic time-integrating correlators [8]. The compactness and ruggedness of the acoustoelectric devices makes them very attractive, however.

The acoustooptic (AO) architectures for time integration correlation have used coherent and incoherent light, and have used bulk acoustic wave and SAW delay lines. The architectures which use an interferometric approach with coherent light seem to have some performance advantages. While very similar processing architectures can be implemented using either bulk wave or surface wave delay lines, SAW delay lines have certain advantages in terms of manipulation of the signals, especially with regard to the AO interaction. A new AO architecture for performing time-integration correlation (which we call a two-beam SAW AO time-integrating correlator) is the subject of this report. It features the optical interference between two coherent light beams which have been Bragg-diffracted by SAW's and achieves large instantaneous bandwidth and very long integration times. Time–bandwidth products exceeding one million have been achieved while maintaining large useful dynamic range, and the correlation is obtained over a substantial range of relative signal delays. This new correlator has been used to detect a number of wide-band spread-spectrum signals, and an analysis of its operation indicates that it may be particularly well suited for use as a spread-spectrum signal processor.

II. Correlator Analysis and Performance

Fig. 1 illustrates the operation of the two-beam SAW AO correlator. A SAW delay line is used in a setup similar to those previously reported [3], [9], [10] for obtaining convolution and correlation with AO devices. The transducers on the delay line are fabricated so that they are tilted from the perpendicular to the z-axis of the delay line (y-cut z-propagating lithium niobate) by $\theta_{\beta\eta}$, the Bragg angle [11] in the delay-line material at the correlator design center frequency ω_0. The relative tilt between the transducers is $2\theta_{\beta\eta}$. The correlator input signals, $A(t)\cos\omega_A t$ and $B(t)\cos\omega_B t$, generate counterpropagating SAW's which interact with two sheet beams of laser light that are

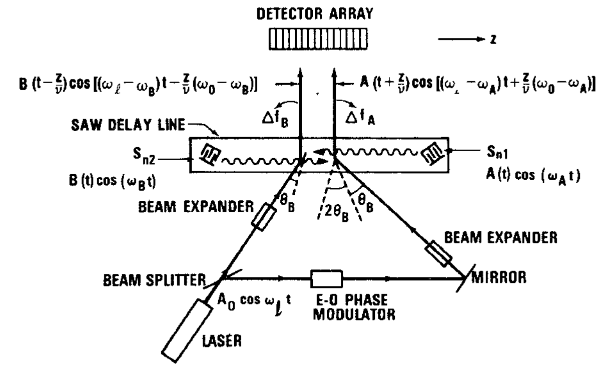

Fig. 1. Functional block diagram of the two beam SAW AO time integrating correlator.

projected into the top side and along the top surface of the delay line. One novel feature of this device is that these sheet beams are derived from a single laser. The light output of the laser is split into two beams which are then shaped, phase equalized, and projected into the delay line with an angle $4\theta_\beta$ between them. Here θ_β is the Bragg angle in air [11] for the frequency ω_0. Because of the strong angular dependence of the AO Bragg intersection, the right sheet beam shown in Fig. 1 interacts primarily with the SAW launched by the right transducer. Similarly, the left beam interacts with the SAW from the left transducer. Cross terms can be held to -40 dB [12]. The light that is Bragg-diffracted by the SAW's is imaged onto an integrating square-law detector diode array. The spatial amplitude, frequency, and phase of this light can be (for equal, uniform-intensity sheet beams) described by the equation

$$L(t,Z)=A\left(t+\frac{Z}{v_a}\right)\cdot\cos\left[\omega_l t+\frac{\omega_l Z}{c/\sin\theta_i}-\omega_A\left(t+\frac{Z}{v_a}\right)\right]\cdot+B\left(t-\frac{Z}{v_a}\right)\cos\left[\omega_l t-\frac{\omega_l Z}{c/\sin\theta_i}-\omega_B\left(t-\frac{Z}{v_a}\right)\right]. \tag{1}$$

Here ω_l is the light frequency, t is time, Z is the distance along both the delay line and the diode array (the center position is $Z=0$), v_a is the acoustic propagation velocity, c is the free-space light velocity, and θ_i is the angle of incidence of the light beams. This angle, θ_i, equals $2\theta_\beta$. The sine of the Bragg angle θ_β equals the light wavelength divided by twice the acoustic wavelength (at the design acoustic frequency ω_0); therefore, where ω_0 is the design center frequency

$$\frac{\omega_l}{c/\sin\theta_i}=\frac{\omega_0}{v_a}.$$

Note that (1) consists of two terms, one from each Bragg-diffracted light beam. Since both sheet beams originate from a single coherent source, the spatial phase terms (in

Z/v_a) in the diffracted light describes the constructive and destructive interference of the light diffracted onto the diode array by the two SAW's. The diodes form the square of the sum, yielding an output current

$$\begin{aligned} I(t,Z) = {} & A^2\left(t+\frac{Z}{v_a}\right) \\ & \cdot \cos^2\left(\omega_l t+\frac{\omega_0 Z}{v_a}-\omega_A\left[t+\frac{Z}{v_a}\right]\right)+B^2\left(t-\frac{Z}{v_a}\right) \\ & \cdot \cos^2\left(\omega_l t-\frac{\omega_0 Z}{v_a}-\omega_B\left[t-\frac{Z}{v_a}\right]\right) \\ & +2A\left(t+\frac{Z}{v_a}\right)B\left(t-\frac{Z}{v_a}\right) \\ & \cdot \cos\left(\omega_l t+\frac{\omega_0 Z}{v_a}-\omega_A\left[t+\frac{Z}{v_a}\right]\right) \\ & \cdot \cos\left(\omega_l t-\frac{\omega_0 Z}{v_a}-\omega_B\left[t-\frac{Z}{v_a}\right]\right). \end{aligned} \tag{2}$$

The cross-product term in (2) may be manipulated using the appropriate trigonometric identity to yield a difference term

$$\begin{aligned} D(t,Z) = {} & A\left(t+\frac{Z}{v_a}\right)B\left(t-\frac{Z}{v_a}\right) \\ & \cdot \cos\left([\omega_B-\omega_A]t+\frac{Z}{v_a}[2\omega_0-\omega_A-\omega_B]\right) \end{aligned} \tag{3}$$

and a sum term

$$\begin{aligned} S(t,Z) = {} & A\left(t+\frac{Z}{v_a}\right)B\left(t-\frac{Z}{v_a}\right) \\ & \cdot \cos\left(2\omega_l t-[\omega_A+\omega_B]t+\frac{Z}{v_a}[\omega_B-\omega_A]\right). \end{aligned} \tag{4}$$

The array time-integrates the detected signal (diode current) for a period T, yielding an output voltage for the difference term (3) of

$$\begin{aligned} V(T,Z) = {} & \int_T A\left(t+\frac{Z}{v_a}\right)B\left(t-\frac{Z}{v_a}\right) \\ & \cdot \cos\left([\omega_B-\omega_A]t+\frac{Z}{v_a}[2\omega_0-\omega_A-\omega_B]\right)dt \end{aligned} \tag{5}$$

which is recognized as the correlation of $A(t)\cos(\omega_0-\omega_A)t$ and $B(t)\cos(\omega_0-\omega_B)t$ in a compressed time frame.

The sum term, (4), varies rapidly with time (as about $\cos[2\omega_l t]$) and the time integral of this term tends to zero for integration times $T \gg 1/2\omega_l$. The integration of the A^2 and B^2 terms yields a "pedestal" output level on which the correlation term rides.

Fig. 2 shows the output of the diode array when both delay-line inputs consist of the same binary code (a linear pseudonoise code) on a phase-shift-key (PSK) modulated carrier (a "direct-sequence" spread-spectrum signal) [1, ch. 2]. The carrier frequency was 187 MHz (the design center frequency), the minimum chip duration was 0.25 μs (a 4-Mbit code rate), the code length was 2^{16-1} bits, and the integration time was 30 ms. Evaluating (5) for a PSK signal yields

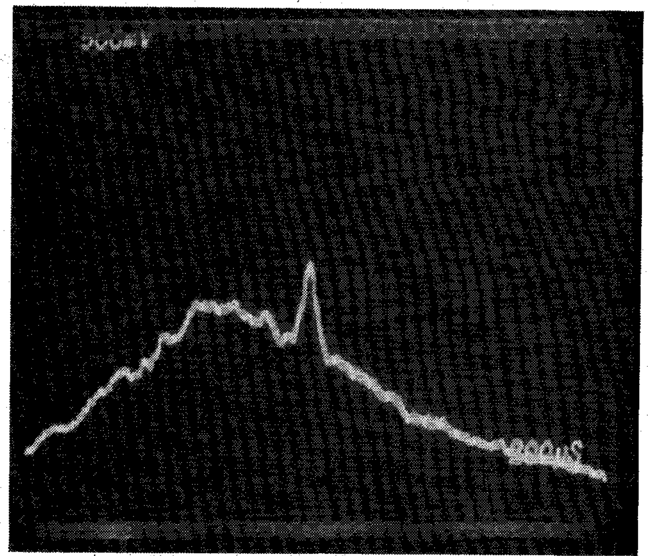

Fig. 2. Correlator output for a 4-Mbit direct sequence input signal at 187 MHz.

$$V(T,Z)=\cos\left(\frac{2Z}{v_a}[\omega_0-\omega_A]\right)\int_T A\left(t+\frac{Z}{v_a}\right)A\left(t-\frac{Z}{v_a}\right)dt. \tag{6}$$

When $\omega_0=\omega_A$, the output contains only the real part of the autocorrelation of the modulation (the direct sequence code) compressed in time by a factor of 2. The autocorrelation of a direct sequence code is a triangular function. This is shown in Fig. 2, riding on a pedestal which is due to the square terms. The amplitude variation of the pedestal is caused by the approximately Gaussian intensity profile of the sheet beams of light used in the device.

When $\omega_A \neq \omega_0$, the output voltage from the array has an additional variation due to the $\cos((2Z/v_a)[\omega_0-\omega_A])$ term in (6). This spatial "fringe pattern" results when the light beams diffracted by the SAW's are not parallel, which is the case when the input frequency is not equal to ω_0 (Fig. 3). By examining the spacing and sense of the fringes one can determine the input-signal frequency relative to ω_0, and find both the real and imaginary parts of the correlation. The maximum frequency difference that can be determined is limited by the spacing of the detector diodes in the array (which "samples" the fringe pattern). The array used in the processor has 1024 diodes across a 2.54-cm aperture, limiting fringe resolution to the equivalent of $\pm$34-MHz frequency deviation (Nyquist frequency). The delay-line bandwidth is 50 MHz.

The pedestal due to the square terms may be removed with the following scheme. Two integrations of the signals are performed, with the phase of one of the delay-line inputs changed by 180° (by using a double balanced mixer) for the second integration. This changes the sense of the

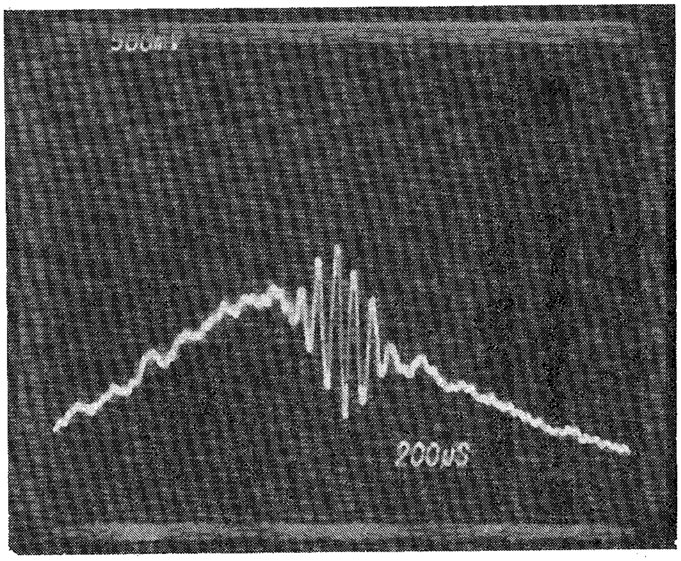

Fig. 3. Correlator output for a 1.5-Mbit direct sequence input signal at 190 MHz.

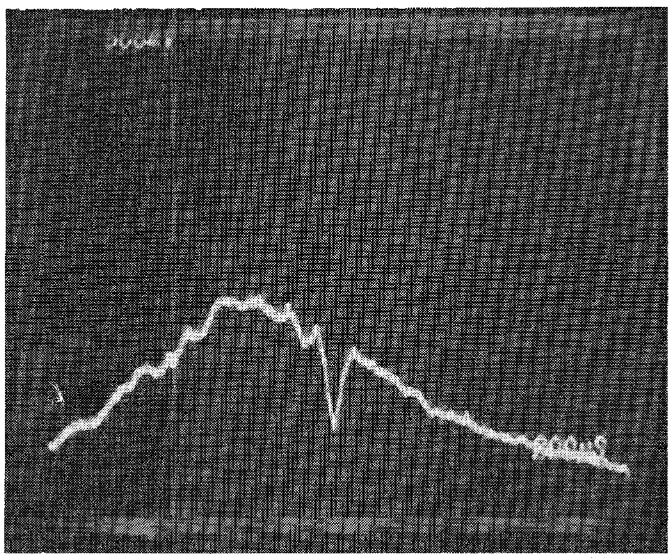

Fig. 4. Correlator output for a 4-Mbit direct sequence input signal at 187 MHz, with 180° phase shift on one input.

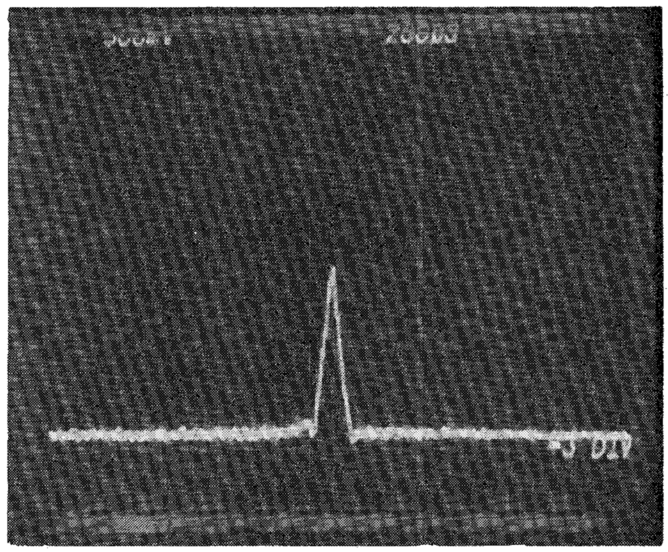

Fig. 5. Result of subtracting the correlator output shown in Fig. 4 from the correlator output shown in Fig. 2.

correlation output as shown in Fig. 4. The output from the second integration is subtracted from the first (using digital postprocessing), doubling the amplitude of the correlation output and removing the pedestal (Fig. 5).

Evaluating (5) for a PSK signal when one of the inputs is delayed by a time t_0 yields

$$V(T, Z)=\cos\left(\omega_A t_0+\frac{2Z}{v_a}[\omega_0-\omega_A]\right)$$

$$\cdot\int_T A\left(t+\frac{Z}{v_a}\right)A\left(t+t_0-\frac{Z}{v_a}\right)dt \qquad (7)$$

which indicates that the position of the correlation peak on the diode array is shifted by an amount $\Delta Z=t_0 v_a/2$. This displacement in location of the modulation correlation peak on the array is a direct measure of the time difference of arrival (TDOA) of the two signals. The 2.54-cm aperture of the correlator (7 μs of delay) limits TDOA measurements to about ± 7 μs at present (twice the time delay aperture).

In (7), the location of the fringe pattern under the correlation envelope is offset by a phase term $\omega_A t_0$. When $\omega_A=\omega_0$, there are no fringes and only the real part of the correlation is available, which results in zero output when $\omega_A t_0=\pi/2$ and odd multiples of $\pi/2$. In the more usual case, $\omega_A\neq\omega_0$, and the $\omega_A t_0$ phase term (at $(2n+1/2)\pi$) changes the fringe pattern from cosine to sine, indicating the change in correlation from pure real to pure imaginary.

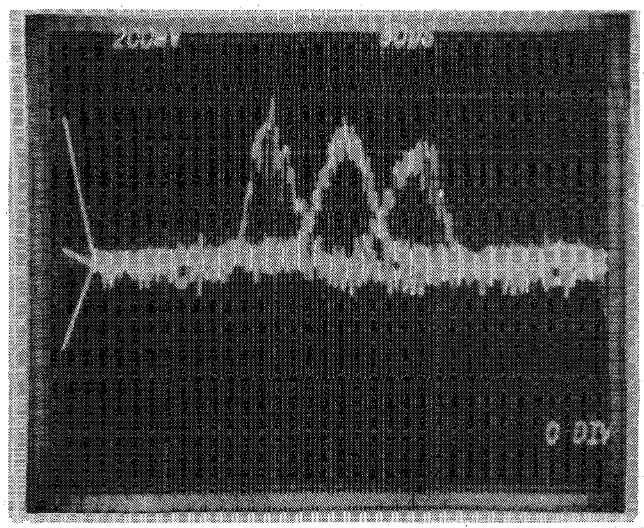

Fig. 6. Time difference of arrival of −260 ns on left, 0 in center, and +260 ns on right.

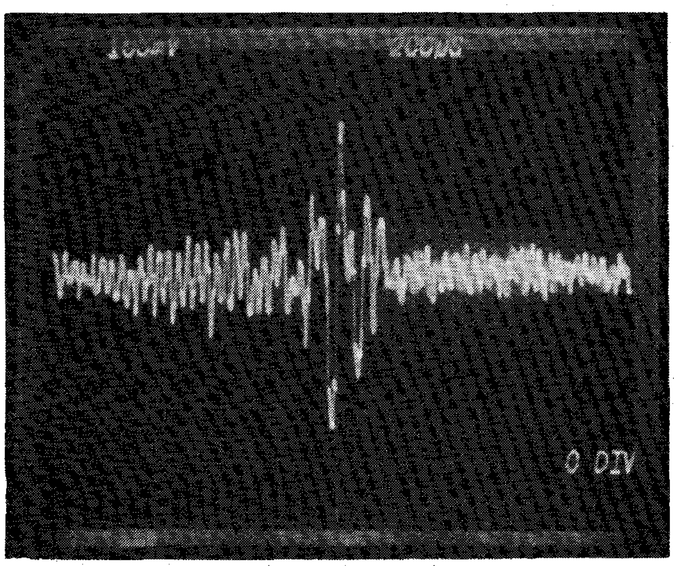

Fig. 7. Correlator output for an input of three simultaneous signals: a CW signal, an FM chirp, and a direct sequence signal.

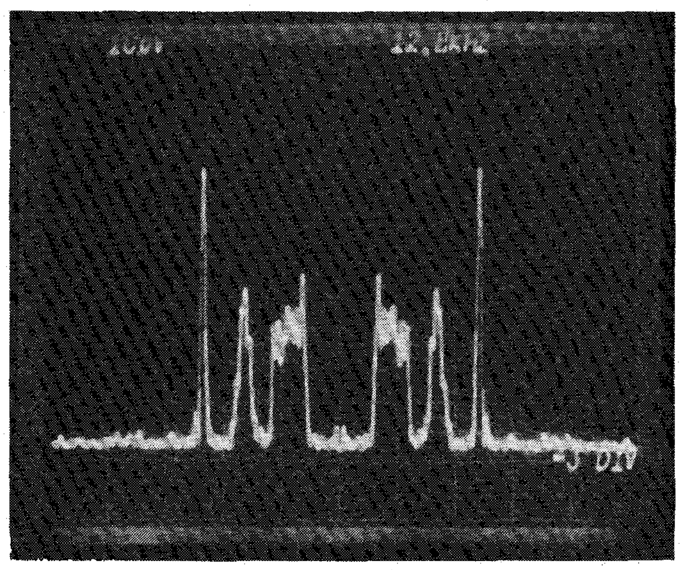

Fig. 8. Power spectrum of the correlator output shown in the preceding Fig. 7.

Fig. 6 shows the shift in position of the correlation peak of a direct-sequence signal obtained by delaying one of the inputs to the correlator. The right peak corresponds to 260-ns delay in the right signal, the central peak is for no delay, and the left peak results from 260-ns delay in the left signal. The ability of the correlator to resolve TDOA is limited by the diode spacing in the detector array. Using an array with 1024 diodes in 2.54 cm yields a TDOA resolution limit of about 14 ns.

One feature of the AO signal processors under development at the Harry Diamond Laboratories is the high degree of linearity of the AO interaction in these devices, suggesting their suitability for multisignal processing. This is true of the two beam surface wave AO time integrating correlator, as can be seen in Figs. 7 and 8. Fig. 7 is the output of the device for an input consisting of three simultaneous signals: a CW signal at 179 MHz, a 1-Mbit direct-sequence signal at 181 MHz, and a 2-MHz chirp centered at 184 MHz. A digital fast Fourier transform is done on this output, and the power spectral density is shown in Fig. 8. The form, center frequency, and bandwidth of the input signals can be estimated from this power spectrum. The individual input signal correlations can then be reconstructed and TDOA information determined.

The dynamic range of the correlator is demonstrated by Figs. 9 and 10. Fig. 9 shows the correlator output when one input was a "clean" direct-sequence signal at a reference power level 0 dB, and the other input was the direct-

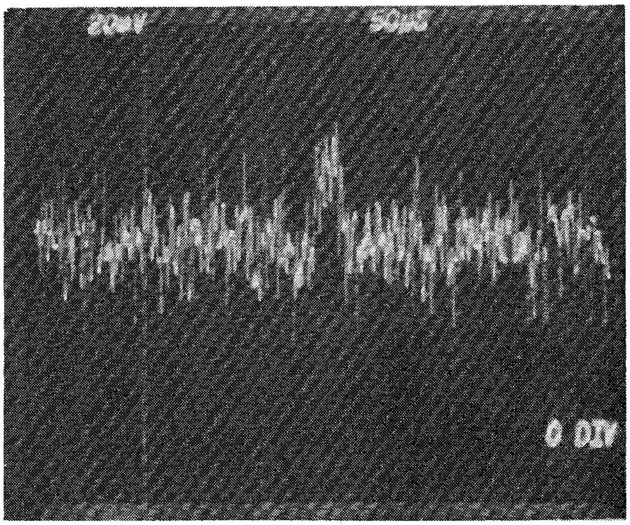

Fig. 9. Correlator output with reference input at 0 dB and signal input at -43 dB (direct sequence signal).

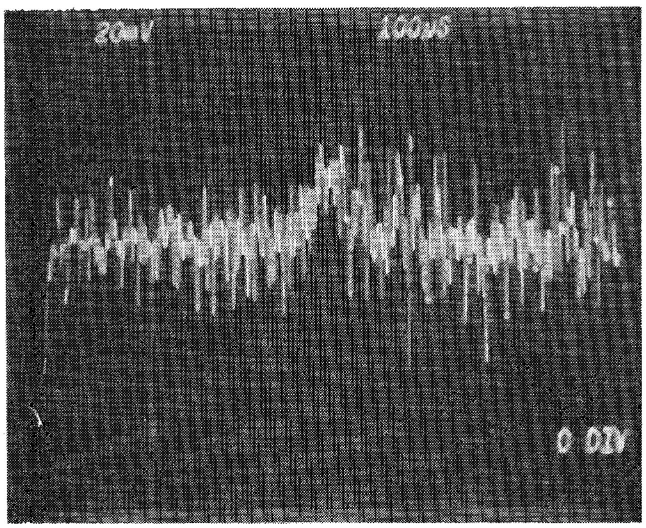

Fig. 10. Correlator output with both inputs at -24 dB (direct sequence signal).

sequence signal at -43 dB. Fig. 10 shows the output for bilinear operation when both inputs were the direct-sequence signal at -24 dB. The integration time for both measurements was 30 ms. The maximum correlation output amplitude is limited to about 3 V by the diode-array saturation level. The array readout process and the analog to digital (A to D) conversion used in the postprocessing cause an output noise level which varies from about 16 mV rms for low light levels (corresponding to bilinear operation with both inputs at -24 dB) to 24 mV rms at high light levels (as when one input is at 0 dB). This variation in noise level with light level explains the difference in dynamic range of about 5 dB for the two cases considered. The ultimate dynamic range which might be achieved with reasonably expected improvements in readout circuitry and in the A to D conversion is about 70 dB [13] (35 dB for bilinear operation) for the detector array presently used.

III. System Applications

The large processing gain, linearity, and TDOA capability of the two-beam SAW AO time integrating correlator should make it generally useful as a spread-spectrum signal processor. Three specific applications which will be discussed are 1) use as a synchronization detector in a spread-spectrum communication system, 2) use as a high-gain matched filter in a low probability of intercept (LPI) radar system, and 3) use as an intercept receiver in an electronic warfare system.

A. Synchronism Detector

A major problem associated with spread-spectrum communication systems is that of detection and synchronization [13]. For example, in the simplest "push-to-talk" spread-spectrum systems only the code sequence employed by the transmitters (e.g., man-portable mobile units) may be accurately known. Exact carrier frequency and code rate are limited by the accuracy of transmitter reference oscillators (typically a few parts in 10^6 for simple crystal oscillators) and by Doppler effects. Relative code phase is completely uncertain when a simple "cold-start" transmitter is used. With these system characteristics it can prove difficult to determine that a transmission has occurred, let alone decode the information sent.

The large processing gain and the time aperture of the two-beam correlator make it very attractive as a detection and synchronization processor. Many spread-spectrum systems employ a short, information-free code sequence (a preamble) at the start of a transmission. An 8-Mbit direct-sequence system might employ a 1-ms word repeated 100 times before starting data transmission, for example. A two-beam correlator with one input consisting of this preamble as a reference signal would have a processing gain of about 42 dB, and would search a relative code phase "window" equal to twice the time-delay aperture of the correlator. The reference code phase would be shifted by this doubled time delay before each integration until a correlation peak occurs, rather as a discrete version of the "sliding correlator." The 14-μs time window of the two-beam correlator would require an excessive number of searches (1-ms integrations) to find the example signal. However, SAW AO devices with 15-cm delay lines (40-μs apertures in $LiNbO_3$) have been built, and could conceivably be used in a two-beam time-integrating correlator. Such devices could search through all relative phases of the example preamble in 13 integrations (each of 1-ms duration). This 13-ms acquisition time compares very favorably with the 1.4 S required by a conventional sliding correlator having similar processing gain [1, ch.6]. The TDOA feature would allow code phase determination to within a fraction of a bit, thus allowing the receiver code reference to be synchronized for data recovery.

In more sophisticated spread-spectrum communication systems where universal timing is employed, phase code uncertainties still exist because of clock errors (which can be made arbitrarily small) and range variation. In such systems, the integrating correlator can be used as a high-gain processor for accurately determining the range between the transmitter and the receiver. However, system frequency errors and Doppler effects place definite limits on the processing gain which can be attained with the time-integrating correlator. The gain limit can be estimated by examining correlator response to two single-frequency CW signals, $\cos \omega_A t$ and $\cos \omega_B t$. From (5) one obtains

$$V(T,Z)=2T\cos\left(\frac{Z}{v_a}\left[2\omega_0-\omega_A-\omega_B+\frac{\{\omega_B-\omega_A\}T}{2}\right]\right)\cdot\frac{\sin\left([\omega_B-\omega_A]\frac{T}{2}\right)}{[\omega_B-\omega_A]T/2}. \quad (8)$$

When the frequency difference, $\omega_B-\omega_A$, is equal to $2\pi/T$, the output drops to zero because of the sine function. For a 1-ms integration time, this corresponds to a total frequency difference of 6 kHz. This suggests that for modulated-carrier systems there is a maximum-frequency deviation which cannot be exceeded if correlation detection with long integration times (large processing gain) is to be used.

B. LPI Radar Systems

LPI radar requires transmitted signals with both low power spectral density and low absolute power. Spread-spectrum signals and signals like FM chirps spread the signal power over large bandwidths. The processing gain of the matched filters which can be used to receive such signals greatly reduces the transmitted power required to detect a target at a given range. The very large processing gains that can be achieved with the two-beam correlator (perhaps 60 dB) could allow a substantial reduction in the required transmitted power, as current technology limits matched filters to time-bandwidth products of about 10^4. Most correlation receivers have been extremely limited in the relative time variation between reference signal and received signal over which correlation can be achieved, making them unsuitable for radar signal processing. However, the time delay aperture of the two-beam correlator gives a large range window over which the return signal can be observed, and the range location of the window can be easily changed by delaying the correlator reference input.

The real limitation on the use of the correlator as a radar signal processor is the limitation on usable processing gain due to Doppler frequency shift. Equation (8) suggests that the correlator output degrades significantly when the frequency difference between the reference signal and the received signal exceeds π divided by the integrating time. A 10-GHz radar would experience Doppler shifts of about 9 kHz for radial target velocities of 500 km/h, and use of standard superheterodyne receiver techniques would give this same frequency shift at the correlator input. This would limit the correlator integration time to about 330 μs, yielding a time-bandwidth product (30-MHz processor bandwidth) of about 10^4. To make use of the increased processing gain available from the two-beam correlator would require using many correlators in parallel, each for a particular band of Doppler frequencies. Alternatively, a spread-spectrum signal which has a correlation less sensitive to Doppler shift might be used [14]. An elegant solution which extracts useful information from the Doppler is a two-dimensional frequency-scanning AO time-integrating correlator that has been recently reported [16].

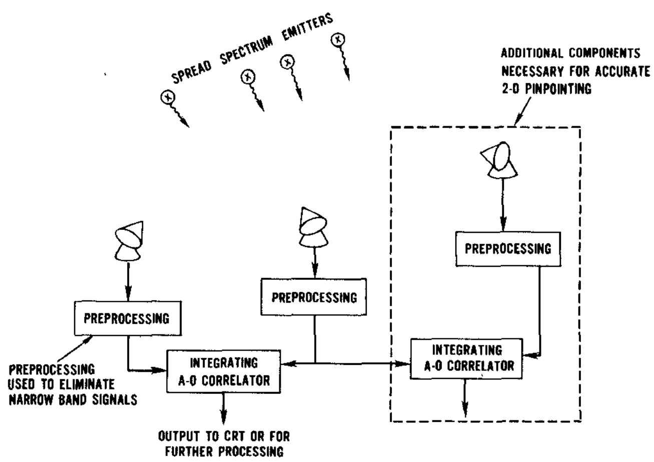

Fig. 11. Spread-spectrum intercept system.

C. Spread-Spectrum Interception

Fig. 11 is a block diagram of a system for using the two-beam correlator as a spread-spectrum interceptor. The inputs to the correlator are the signals and noise from two separate antenna systems. An analysis by Torrieri [17] suggests that if the noise from the two antennas is independent, then the signal-to-noise ratio (S/N) required to establish a given probability of signal detection (P_D) and probability of false alarms (P_F) is (allowing for certain approximations)

$$S/N\simeq\frac{1}{\sqrt{TB}}(\beta_c-\xi). \quad (9)$$

Here T is the correlator integration time which is equal to or less than the signal duration, S is the signal power, B is the signal and noise bandwidth, N is the noise power, and β_c and ξ are scalars derived from the P_D and P_F. Torrieri has also analyzed the wide-band radiometer as a signal interceptor [16] and estimates the required S/N for given P_D and P_F to be

$$S/N=\sqrt{\frac{2}{TB}}\,(B-\xi). \quad (10)$$

Although the approximations made in the analysis are such as to make the small difference in sensitivity insignificant, it is significant that the correlator sensitivity is probably as great as that of a radiometer, because the correlator has many other advantages as an interceptor. In particular, the correlator can distinguish between a spread-spectrum signal and a CW signal which are received simultaneously, while the radiometer can only determine if some signal in addition to noise is being received. The correlator also can find the frequency and bandwidth of received signals, and the TDOA measurements can be used for direction finding.

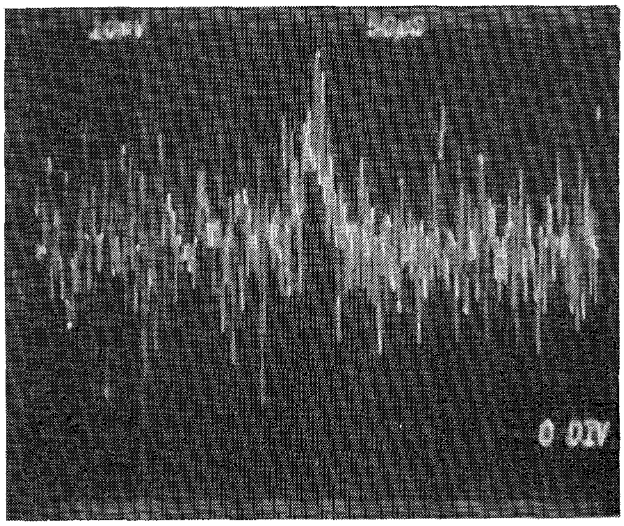

Fig. 12. Correlator output for both inputs at -24-dB S/N (direct sequence signal).

Fig. 12 shows the output of the two-beam SAW AO time-integrating correlator when one input is a direct-sequence spread-spectrum signal 20 dB below the power level of noise from a band-limited "white" noise source, and the other input is the spread-spectrum signal 20 dB below the noise from a second noise source. While the processing gain should have been about $(5\times10^5)^{1/2}$, the minimum signal that could be detected was limited by the readout and digitizing noise previously mentioned. A test was performed with input S/N of -5 and -10 dB, yielding measured values of P_D and P_F at the -10-dB S/N in line with those calculated using (9) and Torrieri's formulas for calculating β_c and ξ [17]. In one series of tests, over 170 000 trials were made without an error. It is felt that this was achieved in spite of the system noise because of the strong limit observed on the peak to average value of this noise.

IV. Conclusion

In summary, a new architecture for performing time-integration correlation has been demonstrated. This new AO device has exceptional ability for processing spread-spectrum signals. It can simultaneously determine the center frequency, bandwidth, and time difference of arrival of intercepted signals. Its very large time–bandwidth product allows very high processing gain when used as a correlation receiver, and the linearity of the AO interaction allows operation in dense signal environments.

Acknowledgment

The authors wish to thank Dr. N. Karayianis for his help in developing the mathematical description and analysis of the two-beam SAW AO time-integrating correlator.

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