Above the crystal, the potential distribution is given by

$$\phi' = V_a \exp(j\beta x - \beta z).$$

If it is assumed that the crystal is weakly piezoelectric as in the case of quartz, the potential inside the crystal is given by

$$\phi = V_a \exp(j\beta x + \beta z).$$

The condition to be fulfilled at the surface by the normal components of the electric induction in empty space D_2' and in the crystal D_2 is

$$D_2' - D_2 = q.$$

Always neglecting the piezoelectric effect, on a Y-cut quartz with the X direction oriented along the x axis, this may be written

$$\epsilon_0 \beta V_a + \epsilon_{zz} \beta V_a = q_0.$$

This expression allows the definition of the unit surface capacitance as

$$\Gamma = \frac{q_0}{V_a} = \epsilon_0 \beta [1 + \epsilon_{zz}/\epsilon_0].$$

In a more general case, with different crystal cuts, or with a crystal having a much higher piezoelectric coupling than quartz, this formula holds by using ϵ_{zz}/ϵ_0 instead of a more involved expression ϵ' taking into account other components of the dielectric tensor or some of the piezoelectric constant [4]

$$\epsilon' = \frac{1}{\epsilon_0} \sqrt{\epsilon_{zz}\epsilon_{xx} - \epsilon_{zx}^2} + \text{piezoelectric terms.}$$

ACKNOWLEDGMENT

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Application of Acoustic Surface-Wave Technology to Spread Spectrum Communications

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Invited Paper

Abstract—**Spread spectrum transmission is being proposed for an increasing number of digital communication, navigation, and radar systems. One of the reasons is the simplicity and availability of surface-wave devices (SWD) for performing the necessary signal generation and processing. The properties of spread spectrum signals, the operation of SWD's, and their advantages and limitations when used in communication systems are discussed. Spread spectrum terminology and basic concepts are defined in terms common to both systems engineers and device designers.**

I. INTRODUCTION

SPREAD spectrum transmission is a form of signal processing which trades transmission bandwidth for enhanced detectability and interference rejection in various digital communication, navigation, and radar systems [1]. In this paper, we consider the advantages and limitations of the application of acoustic surface-wave technology to spread spectrum systems. As a continuing theme to this discussion, we shall consider techniques for transmitting a digital data signal occupying a bandwidth considerably larger than required for the specified data rate. Three principal reasons for artificially enlarging the bandwidth of an information signal will be discussed in Section II. First, spread spectrum techniques permit a communication link to exhibit an attenuation against average power limited interfering signals that are not correlated with the particular waveform used to spread the spectrum. Such interfering signals might be deliberate jamming, random natural events, or even other users of the same spectrum. Second, signal-to-noise improvement even against receiver noise may be obtained by certain systems which make use of several codes; i.e., a given message may be communicated with a given reliability with less energy

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consumed than would be possible with an uncoded signal. Finally, enhanced time resolution may be obtained with the increased bandwidth, as may be desired for range measurements and some forms of analog transmission.

A serious consequence of spectrum spreading is the complexity of the signal processing required to extract the useful information. One common modulation technique is to use coded sequences to transmit each bit of information. For other systems, a linear FM waveform has significant advantages [2]. In any case, some form of correlation or matched filtering [3]–[5] is required to synchronize the transmitter and receiver to extract the original information. The processing time and equipment expense of doing this filtering with digital techniques has limited spread spectrum applications to areas where system cost is a secondary consideration, such as secure communications and satellite data links [6], [7].

Surface-wave devices (SWD) have the potential to revolutionize spread spectrum systems. The necessary matched filtering can be performed at high rates with a simple microelectronic device of small size. For example, a signal waveform 50 chips long with 5-MHz chip rate would require a digital processor capable of handling 250 analog multiplications and additions per microsecond to perform real-time matched filtering. Equivalent signal processing can be performed by an SWD with dimensions of approximately 1.5 by 0.3 by 0.1 in. A signal with the same number of chips and a 50-MHz chip rate would require an even more complex digital processor, while the corresponding SWD matched filter is even smaller than the 5-MHz device.

It is the purpose of this paper to discuss spread spectrum communications, SWD's, and their combination into communication systems in terms meaningful to both systems engineers and device designers. To keep the discussion within bounds, all specific examples will consider phase-shift keyed (PSK) signals, although most of the techniques described apply equally to other waveforms. Section II describes advantages to expanding the bandwidth of a digital data signal: increased protection against certain classes of interference, increased communication efficiency, and improved time resolution. The significant properties and limitations of SWD's are covered in Section III, followed in Section IV by descriptions of applications to spread spectrum systems particularly suited to SWD's.

Several other papers in this special issue deal with more specialized aspects of SWD's in spread spectrum systems. Grant *et al.* [2] describe in detail the requirements for air-traffic-control systems as well as several illustrative systems using SWD's. Staples and Claiborne [26] review the status of programmable filters.

II. Spread Spectrum Communications

Many possible waveforms can be used to spread the spectrum of a digital signal. A technique in common usage for many years involves PSK modulation of an RF carrier such that the transmitted waveform consists of a series of equal-length segments of carrier signal which differ only in phase. If the phase is 0 or 180°, the waveform is biphase PSK. For the sake of having a specific signal to discuss, this waveform will be assumed unless otherwise specified. Other waveforms of interest are linear and nonlinear FM, pulse-position modulation (PPM), frequency hopping, and various combinations.

A few definitions are in order. Fig. 1 shows a typical PSK signal of duration T_B. This signal is made up of N_C equal-length chips which may have any prescribed phase with respect to a reference carrier (0 or 180° for biphase as shown). The phases are assigned according to some code sequence that in the general case defines a particular symbol. A collection of such codes can define an alphabet of symbols, in which case successful communication of a symbol results in the transmittal of several bits of information. In the simplest binary antipodal case, there is one symbol and the negative (or complement) of that symbol, resulting in 1 bit of information being communicated in the time T_B by N_C chips of code. In the nonspread spectrum case, $N_C=1$, that is, the symbol is uncoded.

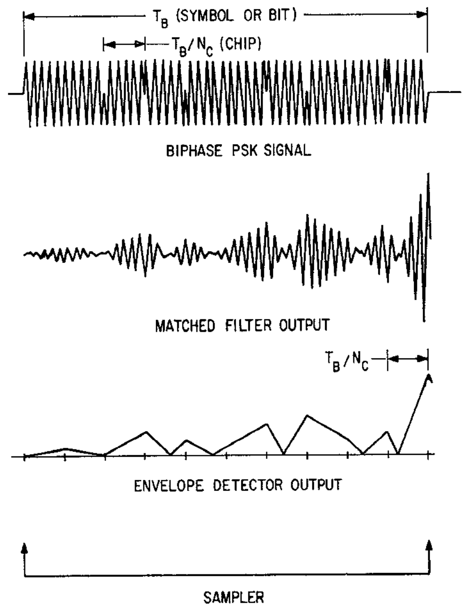

Fig. 1. Timing diagram for binary spread spectrum system.

A brief discussion of both correlation and matched filter techniques for the detection and decoding of a binary biphase signal will be used to illustrate the protection that spread spectrum offers against average power limited interference as well as the role of SWD's in implementing matched filter processing.

A. Interference Protection in Digital Communications

1) Signal Processing Techniques: The signal waveform of Fig. 1, or its complement, is to be used to transmit a data symbol at some bit rate $f_B=1/T_B$ on a carrier f_0. That is, the transmitted waveform is given by

$$V_{\pm}(t) = \cos\left[2\pi f_0 t \pm \alpha_i(t)\right] \tag{1}$$

where $\alpha_i(t)=\pm\pi/2$ for each chip, depending on the code sequence being used, and the sign before α is determined by whether the symbol or its complement is desired.

The bandwidth is determined by the duration of the chip T_B/N_C rather than the symbol duration. Thus, the bandwidth has been increased by a factor of N_C resulting in a spread spectrum signal.

A conventional technique for detecting and decoding the data stream involves use of a correlation receiver in which locally generated replicas of the symbol and its complement are mixed with the arriving signal. If the locally generated signals are given by

$$W_{\pm}(t) = \cos\left(2\pi f_0 t \pm \alpha_i + \gamma\right) \tag{2}$$

where γ is an arbitrary phase term, then the output of the

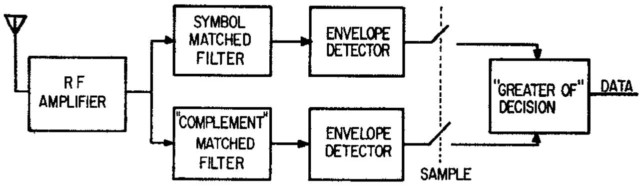

Fig. 2. Matched filter receiver, binary spread spectrum modulation.

mixers after filtering are given by

$$X = \cos\gamma, \qquad \text{reference same as signal}$$
$$= \cos(2\alpha_i - \gamma), \qquad \text{reference complement to signal.} \tag{3}$$

Provided that the local replicas are phase coherent with the incoming signal ($\gamma = 0$), then the output is $+1$ at the mixer which has the same reference as the received signal and -1 otherwise. The system complexity required to achieve this coherence (i.e., synchronization) is the principal disadvantage of this technique.

An alternate approach to data detection for this type of signaling format is the use of a matched filter receiver. The complement to the symbol is now a different symbol rather than its own negative. The receiver contains a pair of matched filters, one for each symbol, as shown in Fig. 2. Here, SWD's are especially useful as they readily permit implementation of matched filters for complex signal structures such as those given in (1).

To illustrate the advantages of SWD's for matched filter application, it will be helpful to briefly review the elements of matched filter theory. Let $h(t)$ be the impulse response of a linear filter. For an arbitrary input signal $s_i(t)$, linear system theory gives the output signal $s_o(t)$ as

$$s_o(t) = \int_0^t s_i(\tau) h(t - \tau)\, d\tau. \tag{4}$$

For matched filter applications, there is a specific relationship between the input signal and the impulse response of the filter. They are matched in the sense that

$$h(t) = s_i(T - t) \tag{5}$$

i.e., the impulse response is the reverse time image of the signal with an arbitrary fixed time offset T.

Substitution of (5) into (4) yields

$$s_o(t) = \int_0^t s_i(\tau) s_i(T - t + \tau)\, d\tau. \tag{6}$$

When allowance is made for the time-limited nature of real signals, (6) is a time-shifted replica of the autocorrelation function of the signal $s_i(t)$.

The practical limitation to using this technique has been the implementation of the integral in (6). Digital techniques require a tremendous number of multiplication and addition steps at each sample point. The SWD is inherently an analog correlation device, thereby removing computation restrictions in the frequency and time ranges where it can operate (see Section III).

2) *Signal-to-Noise Improvement:* We assume that it is desired to detect the presence or absence of a signal whose exact waveform is known *a priori*. Further, assume that the signal has been corrupted by additive white noise and that one is constrained to use only a linear filter for improving the signal-to-noise ratio (SNR), a quantity which determines the reliability of the "presence–absence" decision. Many references [3]–[5], [8] show that the matched filter defined by (5) produces the maximum value for the peak signal-to-rms-noise ratio. This ratio is

$$\mathrm{SNR}_{\max} = \frac{2E}{N_0} \tag{7}$$

where E is the total energy contained in the $s_i(t)$ signal and N_0 is the noise spectral density of the additive white noise at the input to the matched filter. The maximum value of the peak signal-to-rms-noise ratio thus depends only on the signal energy and the white-noise spectral density, independent of signal waveform.

For the receiver shown in Fig. 2, one would select different (rather than complementary) pseudorandom sequences to represent the two possible transmitted symbols. If these sequences are judiciously chosen, the performance of the receiver shown in Fig. 2 is given by [8, p. 298]

$$P_e = \tfrac{1}{2} \exp(-\rho/2) \tag{8}$$

where P_e is the probability of bit error and ρ is the SNR at the output of the matched filter whose corresponding symbol has been received at the time when a decision is made. This SNR is a maximum and given by

$$\rho = \frac{E}{N_0} = \tfrac{1}{2}\,\mathrm{SNR}_{\max} \tag{9}$$

only at time $t = T_B$, i.e., 1-bit duration after reception of the symbol. This implies that the samplers of Fig. 2 are synchronized to the reception time of the symbols, that is, the system has bit synchronization. Assuming such bit synchronization, the quantity which determines system performance is the SNR ratio E/N_0.

Now consider the effect of an average power limited noise-like jamming signal. This noise may be nonthermal noise, intentional jamming, or even a statistical representation of other users occupying the same channel. The noise spectral density in the receiver is given by

$$N_0 = N_{0R} + N_{0J} \tag{10}$$

where N_{0R} and N_{0J} are noise spectral density contributions due to receiver thermal noise and jammer, respectively. If the jammer is average power limited to J W and adjusts its spectral occupancy to match the signal bandwidth, then

$$N_{0J} = \frac{J}{(N_C/T_B)} = \frac{JT_B}{N_C}\,. \tag{11}$$

Since $N_{0J}' = JT_B$ would define the jammer effect if the data signal occupied a bandwidth equal to its data rate, (11) shows that the effect of the bandwidth expansion has been to reduce the effect of the jammer by the factor N_C, which is also the bandwidth expansion factor.

B. Communications Efficiency: M-ary Transmission

It is generally known in communications that the most efficient signals for binary communications are antipodal [8, ch. 7]: one signal is the negative of the other. Coherent PSK is an example of antipodal signaling since a phase shift of 180° is equivalent to changing the algebraic sign. Such signals are most efficient in the sense that bit error probability is minimized for a given bit energy-to-noise spectral density ratio.

If one permits the signal alphabet to become more complex than binary, a given number of information bits can be transmitted at a specified error rate with less total energy than required for the optimum antipodal binary signal. In exchange for a savings in required bit energy, one must pay in the form of increased bandwidth and equipment complexity. A thorough development of this concept is provided by Viterbi [9], [10]. Briefly, one constructs an alphabet of $M=2^k$ symbols. Transmission of one such symbol conveys

$$I = \log_2 M = k \tag{12}$$

bits of binary data. Such schemes are called M-ary coded transmissions. Each of the symbols consists of a sequence of n elementary signals or chips. Typically, $n \gg k$ in order to permit alphabets to be constructed where each member is orthogonal (or nearly orthogonal) to all other members. Alphabets which have been found to have this property include those called orthogonal, biorthogonal, and transorthogonal codes [10].

Optimal processing at the receiver requires the incoming signal to be correlated with all M possible waveforms. The receiver determines the most probable waveform to be that one having the highest value of correlation as measured by the M receiver correlators. This decision minimizes the probability of error.

Surface-wave technology is important to M-ary communications because it permits a simple implementation of the parallel correlator function in the optimum receiver. Each SWD is matched to one particular alphabet symbol. All M SWD's receive the incoming waveform simultaneously. The matched filter processing action of the SWD's is exactly that required by the optimum receiver. At the instant of peak correlation, the outputs of all M SWD's are compared. The SWD having the greatest output voltage corresponds to the most probable signal waveform.

Justification for this added complexity comes from the ability to maintain a given bit error probability with less received energy per bit. Upwards of 3- to 6-dB energy savings can be realized if one goes to sufficiently large alphabets, say $M=32$ to $M=1024$ symbols. Realization of this energy savings requires an increase in bandwidth in keeping with Shannon's theory [11]. Typically, the bandwidth is expanded by a factor proportional to

$$\frac{2^k}{k} \tag{13}$$

which increases exponentially with k.

Limited experiments have been reported by Darby *et al.* [12] for the case of $k=1$, using a 13-chip Barker code and its inverse as the waveforms. A more ambitious scheme using four FM signals ($k=2$) in an air-traffic-control system is described in [2].

C. Ranging

In many system applications it is desired not only to communicate data between two terminals, but also to estimate the distance between the terminals. Air-traffic-control [2] and collision-avoidance systems are examples. This can be accomplished either by active transponder ranging or by passive hyperbolic techniques [7], [13], [14]. In either case, for data communication systems, the quantity of interest is the time of arrival of pulses at a receiver. For a noncoherent system, the time of arrival of a pulse is estimated by the time at which the envelope of the pulse (or the envelope of the processed pulse) exceeds some threshold. The rms error in such a measurement is given by [15]

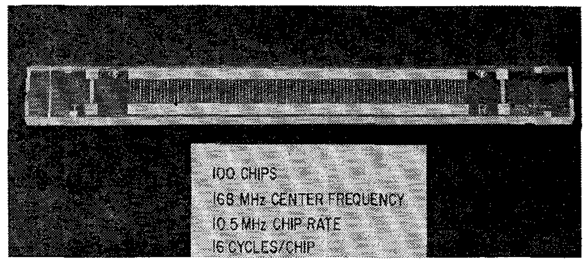

Fig. 3. 100-chip device on ST-cut quartz.

$$\sigma_t = \frac{1}{\beta(2E/N_0)^{1/2}} \tag{14}$$

where β is the effective bandwidth of the predetection pulse, E is the energy in the pulse, and N_0 is the noise spectral density in the receiver, which is a function of receiver design. Equation (14) thus implies that accuracy can be improved by increasing β or E. For a peak-power-limited transmitter, the received energy can be increased only by increasing the transmitted pulse duration τ. For conventional (nonspread spectrum) pulses, the pulse duration is inherently tied to the bandwidth of the signal; increasing the pulse duration decreases the bandwidth, and vice versa. Thus, in (14) one effect tends to offset the other. Clearly, what is desired is a signal structure in which the pulse bandwidth is not tied to the pulse duration so that the two parameters may be selected more or less independently. This is precisely the situation with spread spectrum pulses where the time–bandwidth product for the pulse waveform exceeds unity. It is interesting to note that the use of spread spectrum pulses for measuring range precedes any human attempts at communication; bats have utilized frequency-swept signals in their sonic navigation system since before prehistoric times [16].

III. Coded Surface-Wave Devices

A. Design and Operation of Coded Devices

The matched filter operation described in the previous section is one of a class of filters, called transversal filters, which may be described as tapped delay lines. The application of SWD's to transversal filtering was reviewed very effectively by Squire *et al.* [17] and others [18]. While a great many techniques have been proposed for generating and detecting surface waves [19], the interdigital transducer [20] has proven to be the most practical, as indicated by its dominant role in more recent surveys of surface-wave technology [21]–[23], especially in application to spread spectrum and communications systems [24].

A typical coded SWD (as in Fig. 3) consists of n-pair interdigital transducers at each end of a multiply tapped center transducer. The interdigitated geometry of the transducers is shown more clearly in Fig. 4. A voltage applied between the pads results in a strong field between alternate fingers of the pattern. Since the substrate is piezoelectric, this field produces a periodically varying stress in the material. The resultant wave propagates on the surface away from the electrodes in both directions, in a manner similar to radiation from an endfire antenna. Conversely, as the wave passes under other electrodes, it produces a voltage which may be detected by

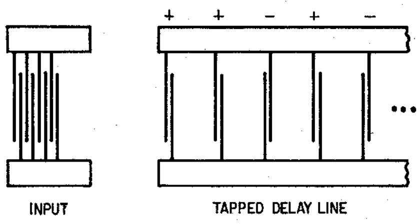

Fig. 4. Tapped delay line finger placement.

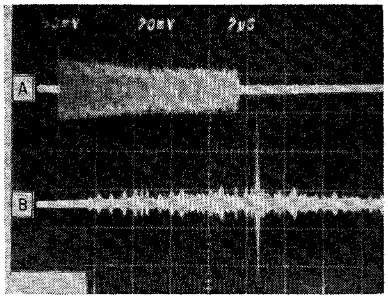

Fig. 5. 100-chip device waveforms. (a) Impulse response. (b) Correlation response.

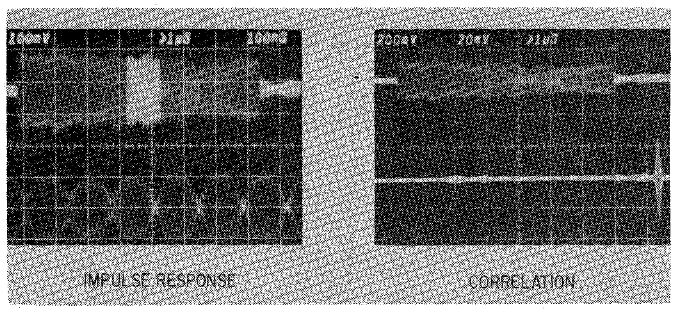

Fig. 6. 64-chip Frank polyphase code on *ST*-cut quartz. Center frequency: 120 MHz. Chip rate: 7.5 MHz.

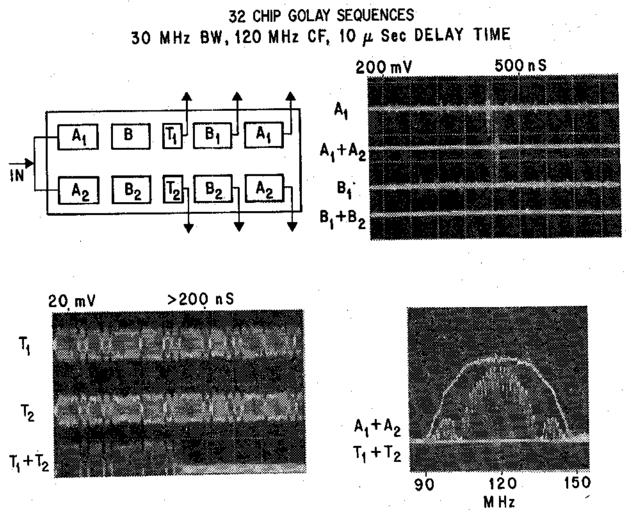

Fig. 7. 32-chip Golay complementary and orthogonal codes.

external circuitry, such as shown in Fig. 5, curve A, for the device illustrated in Fig. 3.

In a spread spectrum communication system, the generated signal is normally clipped, gated, and transmitted from a class C amplifier. In the receiver, the signal will be applied (usually after some signal conditioning) to a second device with the time inversed code. When the signal matches the coded device, a strong response results. At other times, only noise or the relatively weak sidelobe response results, as illustrated in Fig. 5(b). The operation of the system then depends on detecting the existence and timing of the main correlation peak.

The accessibility of the wave to tapping is key to very flexible code generation and transversal filtering. To first order, there is a one-to-one relation between the location of a tap on the substrate and the signal generated by the transducer (i.e., its impulse response) [25]. By proper design of the electrode placement and overlap, it is possible to generate signals of prescribed amplitude, phase, and frequency versus time, including all the standard weighted and unweighted biphase and polyphase codes. Most work to date has been done with codes of fixed center frequency, equal chip length, and fixed chip sequence, but even these restrictions are not necessary. For example, Grant *et al.* [2] describe the usefulness of programming chip sequence and Staples and Claiborne [26] review the current status of techniques for providing this capability.

The design of a device consists of two steps: specify the desired impulse response, then choose substrate material and finger geometry to realize this impulse response in a practical device [25]. Specification of the desired impulse response (code sequence, frequency, etc.) is primarily the responsibility of the communication system designer, who must be aware that some practical limitations exist due to fabrication techniques and choice of material. As will be discussed in the next section, these limitations affect the total time length, carrier frequency, fractional bandwidth, and allowable temperature variation.

Realization of a specified impulse response starts with a first-order design based only on the required impulse response and results in approximate numbers for overall device size, beamwidths, finger placement, impedance, insertion loss, etc. As more and more chips are added to the code sequence at higher frequencies, smaller and smaller levels of distortion become significant while the distortive effects become stronger. The early surface-wave equivalent circuit models [27] can be used to analyze the effect of electrical loading of the taps, but must be extended to include the effects of acoustic reflections at each electrode edge [28], [29]. While the details of the design can be complicated, it is possible to realize in practice almost any desired impulse response within the size and frequency limits which can be fabricated. For example, results have recently been published on a 127-chip biphase device [30], [31] and a 1000:1 FM pulse compressor [32].

Some representative codes should be mentioned before considering fabrication. The 100-chip code shown in Fig. 5 is a conventional biphase code taken from a set of such codes generated by random selection of the phase sequence. Only those codes with desirable auto- and cross-correlation properties are retained in the set. If the criteria are not too stringent, this technique results in a nearly unlimited supply of code sequences with fairly consistent correlation responses. A more deterministic approach to code selection is illustrated by the very limited number of Frank codes [33], noted for their extremely low nonperiodic sidelobes. The codes are all polyphase, with n phases and n^2 chips. An example, with 64 chips and 8 phases, is shown in Fig. 6. Still another approach is to use Golay complementary sequences [34]. Two codes are generated, as shown in Fig. 7, which have sidelobes of opposite

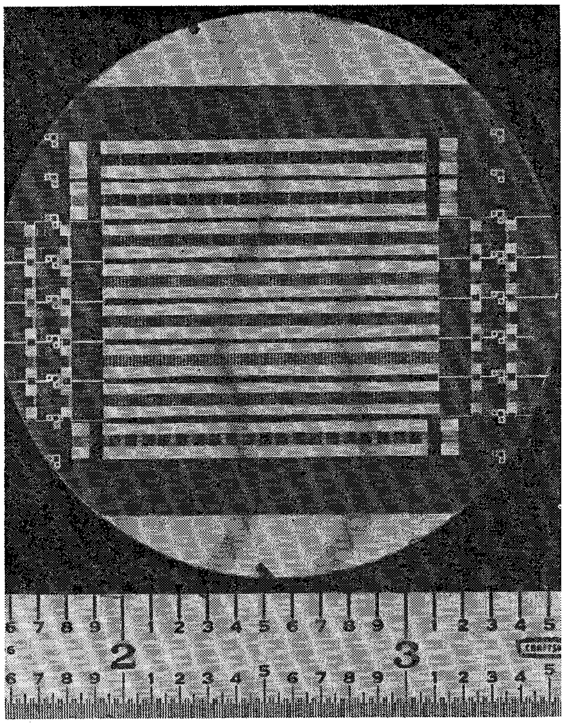

Fig. 8. PSK devices on 2-in slice of ST-cut quartz.

phase. When added, the sidelobes disappear, as shown in trace A_1+A_2. Orthogonal pairs also exist such that their cross correlation is zero [35] (trace B_1+B_2), allowing two sets of data to be transmitted on the two channels required. A variety of other codes have been summarized in the literature on radar signal design [36], as well as some special topics oriented to codes for spread spectrum applications [37], [38]. One last signal of interest is linear FM because of its good correlation properties and immunity to Doppler and phase errors. While waveform variability is somewhat limited, device design has been considerably developed for application to high resolution radar systems [39].

These examples are chosen more to demonstrate the variety of coding techniques that may be implemented with surface waves than to recommend particular codes. Each spread spectrum system will have its own requirements to be satisfied as determined by the number of distinct sequences required, allowable sidelobes, jamming environment, duty cycle, error rates, etc.

B. *Performance Limitations*

The performance of SWD's has been improving at a rapid rate for several years as device designers become more familiar with the photolithographic techniques developed for integrated circuits. Progress is expected to continue at a rapid rate because the microelectronics industry is itself on the threshold of a revolution in circuit lithography as electron-beam and ion-etching techniques become practical. In perspective, it should be remembered that we are discussing applications in communication systems. Since large numbers of users are usually involved, the choice of technique becomes price sensitive as well as technology sensitive. With this in mind, let us look first at what can be done within the constraints of current mass production techniques.

The most highly automated production lines in the world for photolithography are those used for making integrated circuits. Fig. 8 shows a set of PSK devices built on a 2-in

TABLE I

TYPICAL LITHOGRAPHY LIMITATIONS

Technique	Typical Equipment	Full Field Capabilities		
		Field Size (in)	Linewidth (μm)	Line Pairs
Large-scale photoplotting	Gerber photoplotter	30 by 40	50	10 000
Contact printing	Conventional production equipment, 3-in mask	3.5 (diagonal)	4	12 000
High-resolution projection	Mann projection printer	0.25 by 0.25	1	3100
Electron beam	Production oriented	0.300 0.050	2.5 0.5	1500 1250
Electron beam	Scanning electron microscope	0.004	0.1	500

quartz slice using such equipment. Both quartz and lithium niobate are readily available in this size. The codes shown are 10 μs on ST quartz. Longer codes (up to 13 μs) can be accommodated by deleting one of the end transducers and scraping more material at the top and bottom of the slice. While it is desirable to operate below 200 MHz, acceptable results can still be obtained at 300 MHz on ST quartz or 400 MHz on 41.5° Z-cut X-propagating lithium niobate [40]. For those production lines using 3-in slices, an additional 8–10 μs is available, but at slightly lower resolution.

Size restrictions are removed by going to custom fabrication, although increases above about 4 in involve some premium from materials suppliers and increasing difficulty in handling and fabrication. For example, a single code pattern longer than 3.5 in requires special equipment to generate and use. Maximum possible code length is a somewhat nebulous quantity usually not of practical interest in multiple-user communication systems because of the expense involved. For those interested, quartz may be obtained longer than 10 in and lithium niobate in lengths approaching 10 in. The corresponding delay times can be increased even more by such techniques as the wrap-around delay line [41], [42] with separate devices being driven from each delay line tap or by printing parts of the code on each side of the substrate [43].

Resolution restrictions can be eased by use of narrow field of view projection printing. Resolution of 1 μm (0.7–1.0-GHz fundamental) is paid for by a maximum field of 0.25 in from a given reticle. Larger patterns require joining several segments with placement errors of a fraction of a linewidth.

Further improvement in resolution can be obtained by using electron-beam techniques [44]–[46] for producing photomasks and/or devices. The very best resolution (0.14 μm) has been obtained using scanning electron microscopes [47] but over extremely limited fields of view. Machines designed especially for mask generation are capable of larger fields at somewhat worse resolution (e.g., 0.5 μm over 50 mil, for 2500 lines, and similar numbers of lines for larger fields) [48].

Several pattern generation and reproduction machines are compared in Table I in terms of field of view, resolution, and maximum number of line pairs (i.e., cycles of carrier signal) which can be produced without realignment.

For frequencies above 300 MHz and code sequences longer than 1 μs, intrinsic surface-wave attenuation can become important. Small losses can usually be ignored and larger losses can be compensated by weighting the taps. Weighting is less

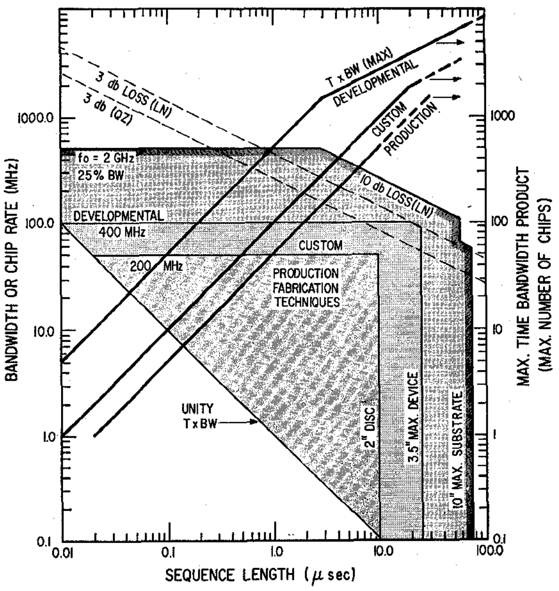

Fig. 9. Current fabrication capabilities for coded devices.

desirable since the same device cannot be used for both code generation and correlation, thereby making fabrication tolerances more stringent. A reasonable loss allowance for practical devices is 3 dB. Assuming that surface preparation has been properly done [49], then attenuation increases quadratically with frequency, with 1 dB/μs at 0.6 and 1.05 GHz for quartz and lithium niobate, respectively [50]–[53].

These restrictions can be summarized graphically as shown in Fig. 9. A 25-percent bandwidth has been assumed for practical reasons (insertion loss, spurious responses from device and electronics, etc.). Devices within the center shaded region are readily obtainable in large quantities. Devices in the lightly shaded region are also readily available but require more custom work. The outer shaded region contains those devices which are technically feasible and potentially producible in quantity but still in the development stage. Extrapolations outside this region entail considerable technical as well as economic risk. The top and right-hand boundary determine the maximum bandwidth and time lengths which may be obtained using the given level of technology. The product is the maximum chip length obtainable and is shown in the figure for each of the three regions.

IV. Surface-Wave Solutions to Spread Spectrum Problems

A. Introduction

A comparison of the needs of spread spectrum systems (Section II) with the capabilities of SWD's (Section III) shows the advantages and limitations of SWD's for this application. The decisive advantage is the capability to do matched filtering with a single simple device at high chip rates. Alternative digital techniques are expensive both in processing time and equipment. The main SWD limitations are due to finite substrate size, resulting in limited chip-sequence lengths and a minimum practical chip frequency. The match between requirements and capabilities has been best for certain types of synchronization, addressing functions, and ranging. These functions can be implemented in a variety of ways, of which several will be briefly described in Section IV-B.

Certain functions which can be performed by spread spectrum communication systems are particularly compatible with SWD's. It should be remembered that each of these functions is being carried out in a noisy environment with limited transmitted or received power. The noise may be natural, deliberate jamming, or incidental to shared occupancy of the channel with other users, but it must be circumvented.

In its simplest form, a communications system exists to transmit data. The presence of a correlation peak can be considered the successful transmittal of a symbol of information. The meaning of this symbol depends upon the organization of the system. Limitations of SWD's require that this symbol be communicated with a relatively short duration waveform.

One particularly useful type of information to send is synchronization information. Conventional digital spread spectrum systems spend a great deal of time acquiring synchronization due to the difficulty of performing correlation processing with digital circuits. Since correlation is a natural function for SWD's, they may be used to initialize and activate the sequence generator of a conventional digital system.

By allowing a variety of codes or adding higher level coding, this same technique can be used to identify or address individual users of the communication channel.

The narrow correlation peak can be used to advantage to make timing measurements. Range between users is the most important parameter based on time, but analog information can also be transmitted by PPM. The accurate timing required would not be possible with narrow-band systems, and the length of the waveforms required is usually within the capability of SWD's.

B. System Examples

The wide variety of applications for SWD's in spread spectrum systems will be illustrated by a few examples for specific systems. Each system can perform one or more of the functions previously described. The dual waveform and on–off-keyed systems describe the transmission of 1 bit of information and hence are basic to understanding the other systems. Each of the other systems involves an extension of the waveform to improve system performance for some particular application.

1) Dual Waveform Binary Communication System: This is the approach discussed in Section II and shown in Fig. 2. Binary data is used to select one of two possible spread spectrum waveforms for transmission. These waveforms may be biphase or polyphase shift keyed, frequency hopped, or any combination. Maximum likelihood detection at the receiver results in a processing gain against certain types of jamming signals, as described in Section II-A.

2) On–Off-Keyed Binary Communication System: This approach is a special case of the dual waveform binary communication system just discussed with one of the two waveforms being no signal at all. Maximum likelihood detection reduces to the decision as to whether signal plus noise is present or noise only. Accurate control of the threshold level relative to the noise level is a more difficult implementation problem than the comparable problem of making greater-of decisions between two matched filter outputs in the case of dual waveform approach.

3) Pulse-Position Modulation: The narrow correlation peak response naturally suggests its use in encoding analog or digital data. Analog data would be encoded by continuously modulating the separation interval between consecutive pulses with the (sampled) analog signal in the transmitter.

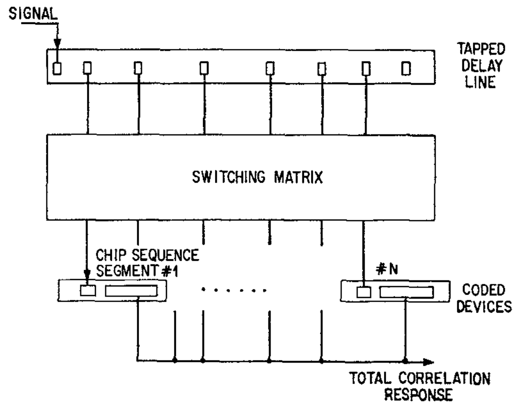

Fig. 10. Correlation of partly programmable code sequences.

The receiver would recover the analog samples by measuring the time between threshold crossings. Examples of useful analog information are sensor signals and round-trip travel time (range). Time-bandwidth-product processing gain is provided in this approach.

Digital data can be encoded in a number of ways. Straight binary PPM would have the interval between pulses constrained to one of two possible values. The receiver would investigate threshold crossing intervals in making binary decisions. M-ary coded PPM would have $M=2^k$ distinct possibilities for the interval between consecutive pulses. Upon measuring the pulse separation interval, the receiver would then have decoded $\log_2 M=k$ bit of information each time it received a new pulse. This technique is analogous to the approach discussed in Section II-B on communication efficiency in that each alphabet symbol conveys more than 1 bit of information.

Higher level coding is possible with PPM where groups of spread spectrum pulses are PPM encoded in unique patterns. The receiver is more complex in that it must decide if it has seen a pattern of individual pulses rather than one pulse The ability to compound the coding scheme through higher level PPM is virtually unlimited [54].

4) *M-ary Coded Transmission:* This is a generalized technique for communications involving the transmission of one of M distinct symbols. This approach was discussed in Section II-B. One implementation was previously described for PPM. An alternative method uses separate codes for each symbol with separate surface-wave correlators for detection.

5) *Partly Programmable Codes:* Chip-sequence length and programming limitations can be partly circumvented with the system shown in Fig. 10. The chip sequence consists of several equal-length sequences in series. The signal to be correlated is applied to a delay line with taps at equal spaces corresponding to the short sequence length. Each tap is connected to an SWD with the chip sequence corresponding to that portion of the total sequence. The outputs of all SWD's are summed to produce the total correlation signal. If each of the n taps is connected to any one of the n SWD's (but only one) by means of a switching matrix, then there are $n!$ distinct sequences available to choose from. Thus, n "easy" devices plus one "custom" device can be used to generate signals equivalent to $n!$ custom devices.

6) *SWD-Acquisition-Aided Conventional Spread Spectrum Communication System:* The shift-register version of spread spectrum communications can be modified to include an SWD in the receiver which is used to permit fast acquisition. Each transmitted message is preceded by a chip sequence which is matched to the SWD being addressed. The amplitude of the SWD response is an indication of the extent of agreement between the incoming signal and the SWD coded address. When this agreement reaches a threshold value, the decision is made that a legitimate preamble has been received and the receiver's code sequence generator is suitably initialized and activated.

V. Summary

SWD's are becoming increasingly important in the development of communications systems based on spread spectrum concepts. Both PSK coding and linear FM waveforms are being used to increase the bandwidth of digital data in order to decrease the error rate and power requirements. Grant *et al.* [2] describe in detail how spread spectrum is applied to a particular operational problem—air-traffic control. We have been concerned here with the more general questions of why use spread spectrum, what are the advantages of SWD's, and some brief general descriptions of representative systems. A short review of the properties of spread spectrum signals showed the importance of matched filtering to successful operation of such systems. Advantages in improved SNR, communications efficiency, and time (or range) measurement pointed up potential application areas.

A broad look at SWD's showed that they are ideally suited to matched filtering operations. Fabrication processes were examined to determine current and potential limitations on chip rates, sequence lengths, and operating frequency. These results were summarized in Fig. 9 in terms of increasing sophistication and cost required for fabrication. In general terms, chip rates below 2 MHz are of debatable practicality and from 50 to 500 MHz involve increasing cost and risk using current technology. Total sequence lengths up to 10 μs are easily obtained below 200-MHz center frequency and can be extended to 25 μs with relatively little effort and to 75 μs in cases where substrate cost is not important. As the frequency is increased into the gigahertz range, acoustic attenuation becomes dominant for longer signals. With proper choice of parameters, it is relatively easy to achieve chip sequences of 1000, and the techniques required to reach 10 000 are clearly defined.

When spread spectrum system requirements were examined, it was seen that the best match to SWD capabilities occurred for the functions of preamble synchronization, including short data bursts; M-ary (or multiple code) transmission, for efficiency or discrete addressing; and timing measurements, for range or PPM. The examples of Section IV illustrated one or more of these functions in systems ranging from simple data channels to more sophisticated concepts employing higher level coding, multiple coded SWD's, and combination with conventional digital systems.

Analog matched filtering is not limited to SWD's. Charge transfer devices (CTD's) have the same capability as SWD's of sampling the signal at defined locations which allows for matched filtering. However, time delays up to 1 s and chip rates of less than 5 MHz make CTD's complementary to SWD's [55]–[57]. The systems of the future may well combine both devices to obtain much longer sequences or a wider range of rates.

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