Ranging and Data Transmission Using Digital Encoded FM-«Chirp» Surface Acoustic Wave Filters

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Abstract—A digital encoded multislope chirp modem/demodem unit has been implemented using two 3-port surface acoustic wave filters. Each filter operating at a 30-MHz center frequency provides either positive or negative slopes as digital 1's or 0's with a time–bandwidth product of 280, an unweighted bandwidth of 5.6 MHz, and a time dispersion of 50 μs. The modems were used to calibrate and compare the operational performance of the conventional multitone CW ranging system with a multislope chirp ranging system. Range measurements and range-rate observations were made to a synchronous satellite with both systems using a ground communication satellite terminal. Both ranging techniques provided accuracies well within the predictable satellite range of 20 000 nmi; the multitone system provided a theoretical range resolution of 1 m, and the chirp system 0.4 m. Data transmission was also accomplished, using 12-bit binary code at a 1.25-Mbit rate.

The significant advantage to be noted with the chirp system is the ability to obtain continuous range data from the satellite repeater simultaneously while other modes of information are being transmitted, and to combine the ranging and data-link transmission on a time-order basis using digital encoded chirp sequences. The chirp system was found the more desirable technique, since it provides a more direct range measurement, with minimal calibration requirements, and provided greater processing gain with relative ease and reliability. The data transmission at low data rates provided little deterioration in theoretical compressive gain. However, at the higher data rate a greater loss was encountered due to power sharing of the overlapping chirp coded carriers in the limiting satellite.

I. Introduction

A SERIES of ranging and data-transmission experiments have been accomplished using a ground communication satellite terminal and a synchronous satellite repeater. The objectives of the experiments were to compare the relative performance of the conventional multitone continuous-wave (CW) ranging technique with the frequency modulated (FM) multislope chirp implementation [1]–[3] using surface acoustic wave filters. Additional objectives included the observation of the ability to use the multislope chirp waveform to provide range corrections due to Doppler shifts in the link, or due to satellite motion, and the spread spectrum data transmission of digital encoded chirp at low data rates (4 kbit) with single carrier operation and high data rates (1.25 Mbit) with multiple carrier operation via a limiting satellite transponder.

The experiments were conducted at the Hughes Commercial Satellite Ground Station at El Segundo, Calif., in December 1971 and June 1972. The terminal used a 16-ft diameter antenna with a gain of 46.5 dB and a system noise temperature of 100 K. The receiver used a liquid helium cooled parametric amplifier. The transmitter used a high-power linear klystron amplifier with a 15-kW power output capability. The satellite was located at synchronous altitude. The satellite transponder operated with its microwave output saturated at 112-dBm effective irradiated power (EIRP) from the ground station. Varying degrees of IF limiting was accomplished in the satellite transponder by varying the waveform format and varying the ground transmitter power from 3 kW to 75 W.

The following sections of this paper provide descriptions and results of digital encoded transmission using FM-chirp surface-wave filters, the description and results of the satellite ranging experiments, and comparisons of the CW and FM-chirp ranging techniques.

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II. Digital Encoded Transmission Using FM-Chirp Surface-Wave Filters

The purpose of a digital encoded FM-chirp carrier system is to provide a low profile intercept or spread spectrum transmission offering the attractive attributes of low cost and minimal Doppler sensitivity. This is accomplished by using large time–bandwidth product surface-wave filters. Additional applications provide chirp ranging, range-rate correcting, and variable data-rate transmission with both multiple slope and coding bit flexibility.

The system can encode and decode digital information using surface-wave dispersive delay lines. Digital coding is accomplished by the assignment of positive and negative chirp pulses for the 0's and 1's of a binary logic. The system described herein uses a 3-port surface-wave filter to generate positive or negative slope time–frequency pulsed waveforms at IF's for transmitter coding selection, and a second 3-port filter for passive separation of the received multislope coded pulsed waveforms. Referring to the simplified block diagram of the system (Fig. 1), this is accomplished in the following manner. The digital message controls the position of the single-pole double-throw switch (SPDT) through a switch driver. The output of the pulse generator, which is triggered by the clock pulse, is directed by the SPDT switch to either the wide-band input transducer on the left (1's), or the one on the right (0's) of the dispersive delay line. If the output is taken off the long chirp transducer, situated between the two input transducers, the digital encoding will exist on an IF carrier as a sequence of overlapping frequency modulated pulses with the information being contained in the form of amplitude and frequency variations with time. When the data rate is such that no overlap occurs, then multiple carriers are not present and the amplitude modulation component is removed. In the receiver the sequence of incoming signals is separated in the surface-wave delay line, appearing as compressed intermediate frequency pulses at the appropriate 1's and 0's output terminals. Video detection of the compressed pulses allows with suitable thresholding recovery of the binary digital message in, for example, a bistable "flip-flop" detector.

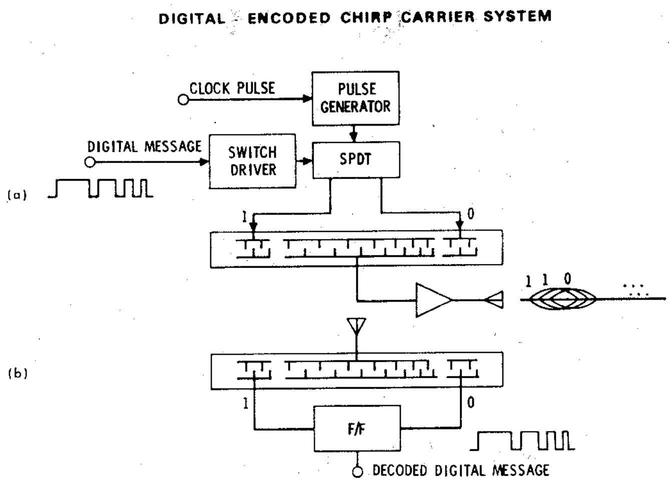

Fig. 1. Digital encoded chirp carrier system. (a) Transmitting mode. (b) Receiving mode.

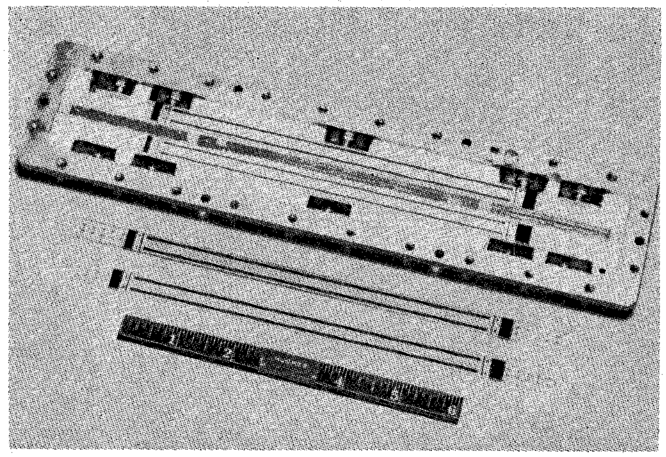

Fig. 2. Experimental surface-wave dispersive delay-line filters used on *ST*-cut quartz. (Designed and fabricated at Hughes Aircraft Company, Culver City, Calif.)

The advantages of this implementation are many. For example, expensive modulators and demodulators are eliminated and digitized data can be retrieved in a spread spectrum format without the use of synchronization pulses. There is an inherent flexibility in the data rate for either nonoverlap or overlapping chirp transmission through the use of multiple slope filters using wide and narrow bandwidths. In addition, there is the reduction or minimization of multipath problems encountered in aircraft-to-satellite communication systems. Video detection of the compressed pulses allows, with suitable thresholding, recovery of the binary digital message in a bistable flip-flop pulse stretcher. The advantages of this implementation are: elimination of expensive carrier modulators and demodulators; retrieval of digitized data in a spread spectrum format with relatively low-cost surface-wave "slope" separators; and with self-synchronous detection. This is significant as compared to the acquisition and tracking problems associated with the phase coded spread spectrum implementation [4]. There is also the flexibility in the data rate for either nonoverlap or overlapping chirp transmission through the use of multiple slope filters using a combination of wide and narrow bandwidths. The disadvantages of this implementation for high data-rate overlapping transmission is the use of a linear transmitter with low intermodulation with its inherent low efficiency. The disadvantage of the implementation for overlapping frequency modulated pulses is the generation of an amplitude and frequency modulated waveform. Removal of the amplitude modulation limiting and the power sharing of frequency modulated pulses is the generation of an amplitude and frequency modulated waveform. Removal of the amplitude modulation limiting and the power sharing of multiple carriers reduces the achievable compressive gain obtainable in some applications.

For spread spectrum applications it is interesting to note some of the significant differences between the digital encoded chirp and the binary phase coded pseudonoise (PN) sequences. The chirp slope coding of $+1$ or -1 is analogous to the phase change in binary coding. In the chirp case the spectral utilization being box-like for large dispersion ratios is more efficient.

In the binary case the spectral distribution is similar to AM, with the carrier being present or suppressed, depending on the PN sequence being even or odd. The AM type of modulation generates spectral bandwidths with null-to-null intervals of 2 times the reciprocal of the bitwidth. Thus, for assumed double sideband transmission with the carrier, twice the bandwidth is required as compared to the chirp (FM) transmission. Obviously, if carrier reinsertion with the proper phase and single sideband transmission is employed, the bandwidth can be reduced with the penalty of additional receiver complexity. Compression of the PN sequence also requires a specific bitwidth-to-carrier-cycle ratio and critical frequency control to insure in-phase addition in the IF delay line. The correlated time response, or compressed pulse with its time sidelobes, can be optimized to a -30-dB level by frequency weighting in the chirp case; however, time weighting with transversal filtering, an additional complexity, is required in the PN case. Doppler sensitivity in reception is minimal in the chirp case so long as adequate bandwidth is maintained for the Doppler shift. However, severe distortion and sequence decorrelation is encountered with the binary phase coder if precision frequency control is not employed. Compressive gain in the 30-dB range has been obtained with dispersive delays in the chirp case, without excessive loss when time delays of 130 μs have been required. Currently reported binary phase coding sequences suffer from extreme tolerance control when similar dispersions are attempted with equivalent time delays. Hybrid processing, as employed in the digital encoded chirp system, offers the benefits of both time and frequency spreading. In addition, compression again is provided and a detectable binary phase code.

The implementation of the 3-port surface-wave filter used in digital chirp coding can be seen in Fig. 2. This is one of a number of filter configurations developed at Hughes (Culver City) using exceptionally long interdigital transducers with low distorition and large time–bandwidth products. These lines operate at 30-MHz center frequency with 5.6-MHz frequency deviation and 50-μs time dispersion. The linearity is of the order of $\pm\frac{1}{4}$ percent with sidelobe suppression performance in excess of -30 dB and leakage suppression in access of -40 dB. The matched loss on *ST*-cut quartz is less than 50 dB.

Fig. 3 displays the input signal, the resulting stretched pulse, and the compressed pulse. The compressed pulse dis-

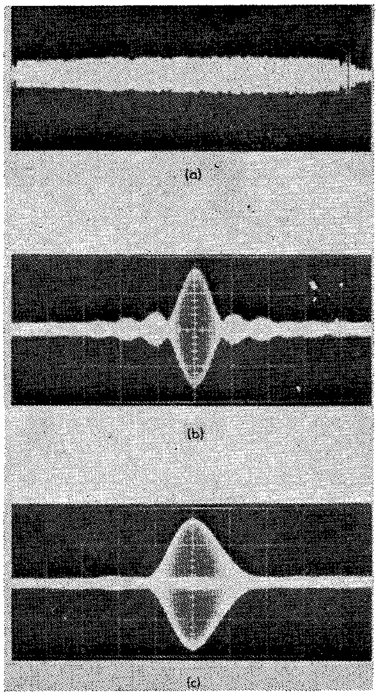

Fig. 3. Performance of 50-μs 5.6-MHz surface-wave chirp delay lines. (a) Dispersed and compressed pulsewidths. (b), (c) Expanded time scale for unweighted and weighted compressed pulses (0.5 μs/div).

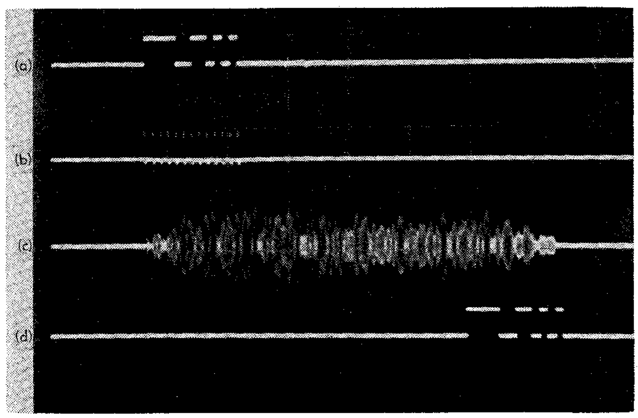

Fig. 4. Laboratory simulation of digital encoded chirp transmission using multiple chirp filters (10 μs/div). (a) Digital input logic. (b) Impulses to line. (c) Composite digital encoded output. (d) Decoded digital input logic.

played is for the case of uniform spectral weighting. The sidelobes are down -13.6 dB from the main lobe, which is in good agreement with the theoretical sinc x response. The compressed pulse is also displayed with Gaussian weighting. The first sidelobe is now 30 dB below the main lobe. However, note that the penalty for sidelobe reduction is an increase in the pulsewidth of the main lobe. For information transmission, sidelobe reduction is desirable so that interbit interference is reduced. For ranging, the rise time of the compressed pulse is the primary consideration, hence weighting is not desirable for this application. Using a laboratory test position similar to that shown in Fig. 1, a PN code of 12 was transmitted and received by chirp filters. Trace 1 of Fig. 4 shows the PN code that is transmitted from the word generator. Trace 2 shows

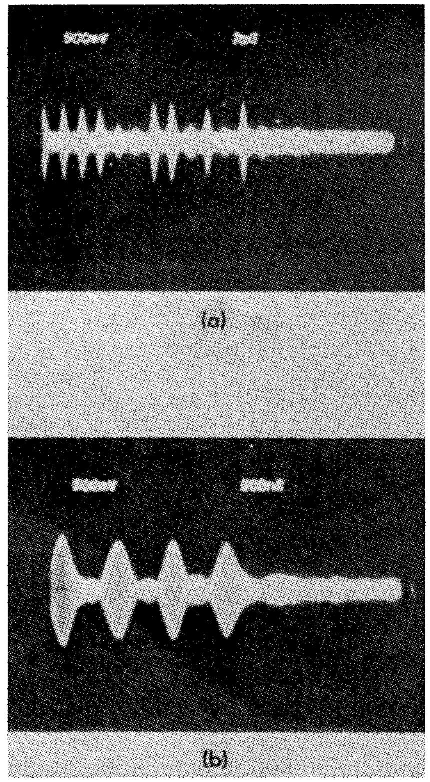

Fig. 5. Received, decoded FM signals from synchronous communication satellite. (a) Compressed positive slope pulses (2 μs/div). (b) Expand time scale (0.5 μs/div).

the sequence of clock pulses used, and Trace 3 shows the composite output signal from the positive- and negative-slope chirp filters. Trace 4 demonstrates the recovered logic after the signal in Trace 3 was compressed using another chirp filter similar to that shown in Fig. 2. Received satellite signals after compression are provided in Fig. 5, indicating the return compressed positive pulses from the satellite.

Note the low interpulse interval indicating the reverse-slope rejection level. The rejection is the order of 14 dB (voltage ratio).

Fig. 5 displays four 1's being received in the IF section. Note that the bit rate is approximately 1.25 Mbit/s and only a small amount of interbit interference is apparent. At this rate, the chirped pulses are forced to share the output power of the satellite. For example, a 50-μs dispersed pulse at a bit rate of 1.25 Mbit/s, results in a power sharing loss of the pulse length times the inverse of the bit rate, equivalent in this case to approximately -18 dB. An additional loss is associated with the limiting of the AM caused by beat notes between overlapping chirp bits. The resulting compressive gain is 13 dB. This signal-to-noise ratio (13 dB) agrees with the observed compressed signals.

III. Satellite Ranging Experiments

A. *Multitone CW Ranging*

The measurement of range with discrete tones is based on the simple principle that the two-way range must be known to at least $\pm\frac{1}{2}$ wavelength of the tone being used for the measurement. This is necessary since the measurement technique does not yield propagation time but phase delay which

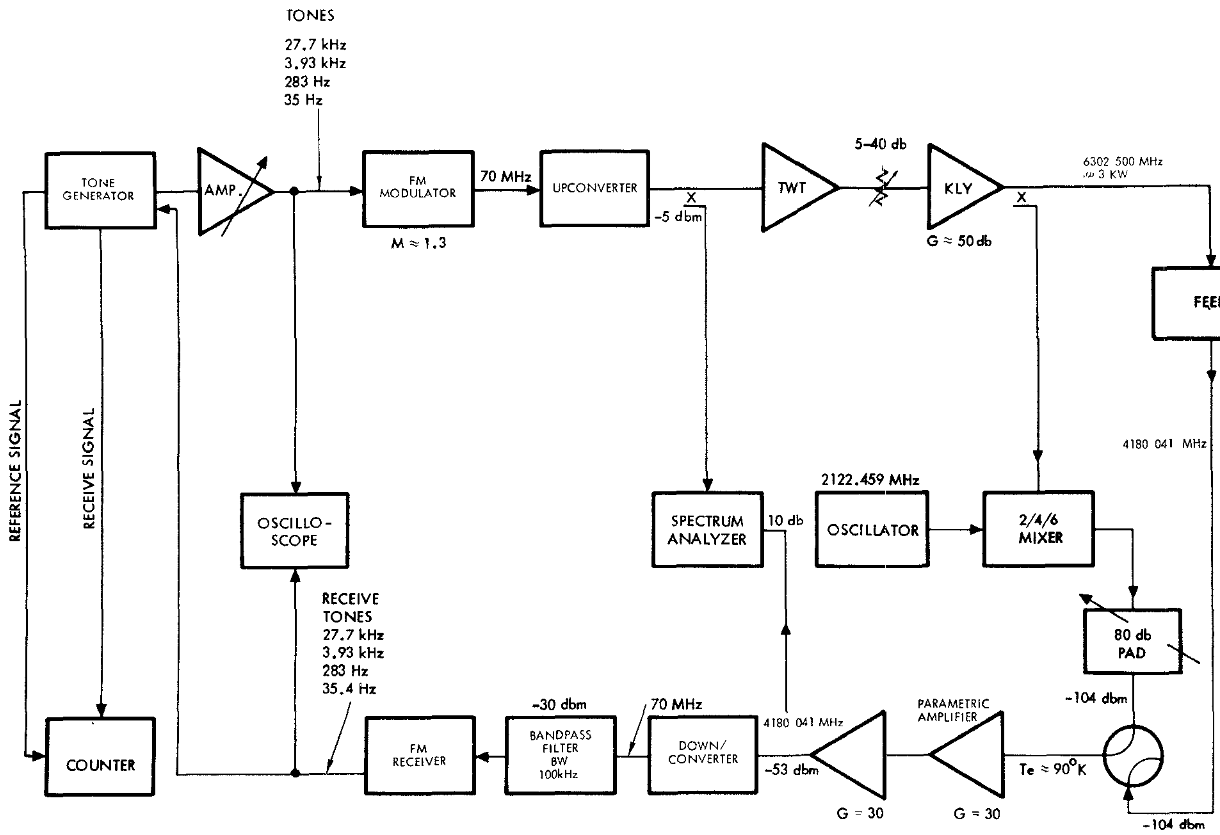

Fig. 6. Block diagram of earth station for tone ranging.

is always ambiguous to some multiple of 2π. Therefore, it is necessary to remove this ambiguity by first estimating the two-way range to $\pm\frac{1}{2}$ wavelength of the lowest ranging tone, and then designing the measurement to yield ranges by each tone that are at least accurate to $\pm\frac{1}{2}$ wavelength of the next higher tone. The accuracy of the ranging technique is inversely proportional to the highest tone frequency utilized. Theoretically, the one-way range accuracy is given by:

$$\Delta R = \frac{C}{4\pi f \sqrt{2\ S/N}} \tag{1}$$

where the error in the one-way range is ΔR, the highest tone frequency is f, the speed of light is C, and the received signal-to-noise ratio is S/N [6].

The tones are frequency modulated on a carrier with a low modulation index [1], [3] in order to minimize error due to group delay in the various electronic components through which the signal is transmitted. The frequencies utilized for the tone ranging experiment were 27.77 kHz, 3.968 kHz, 283.4 Hz, and 35.46 Hz.

From the link calculations, assuming the satellite transmitter is operating at saturation, the received signal-to-noise ratio is calculated to be 55 dB, yielding a theoretical range resolution of 1 m.

Fig. 6 shows the block diagram of the transmit and receive section used at the earth station for the tone ranging experiment. Frequencies from the tone generator are modulated onto a 70-MHz IF frequency and up-converted to 6302.500 MHz. A traveling-wave tube (TWT) acts as an intermediate power amplifier and a Varian klystron as the final high-power amplifier. In order to calibrate the time delay in the earth station itself, a 30-dB directional coupler with a 20-dB attenuator tapped off part of the transmit signal. The tapped signal was down-converted to the receive frequency of 4180.041 MHz and routed through the receive system. A Hewlett-Packard 5360A computing counter was used to measure the time delay between the reference signal provided by the tone generator and that modulated on the carrier and routed through the transmit and receive section. This calibration measurement was performed for each of the tones. The group delay through the satellite was not known, but it was estimated to be on the order of 40 ns, since the minimum filter bandwidth within the satellite is 25 MHz. This delay was not accounted for in the range measurement. The initial estimate of the two-way range was assumed to be 40 000 nmi. The range measurement with the highest tone yielded 20 325.374 nmi for the one-way range. The predicted range for the satellite using tracking data furnished by the Goddard range and range-rate system (GRARR) is 20 317 nmi. The measured range is well within the accuracy limits of the predicted range.

B. *Chirp Ranging Experiment*

Fig. 7 is the block diagram for the chirp ranging experiment and the information transmission experiment. For the ranging experiment, the word generator was not utilized. A single pulse generator was used to drive a positive-slope chirp surface-wave delay line. The line operated at an IF frequency of 30 MHz with a bandwidth of 5.6 MHz, and a delay of 50 μs.

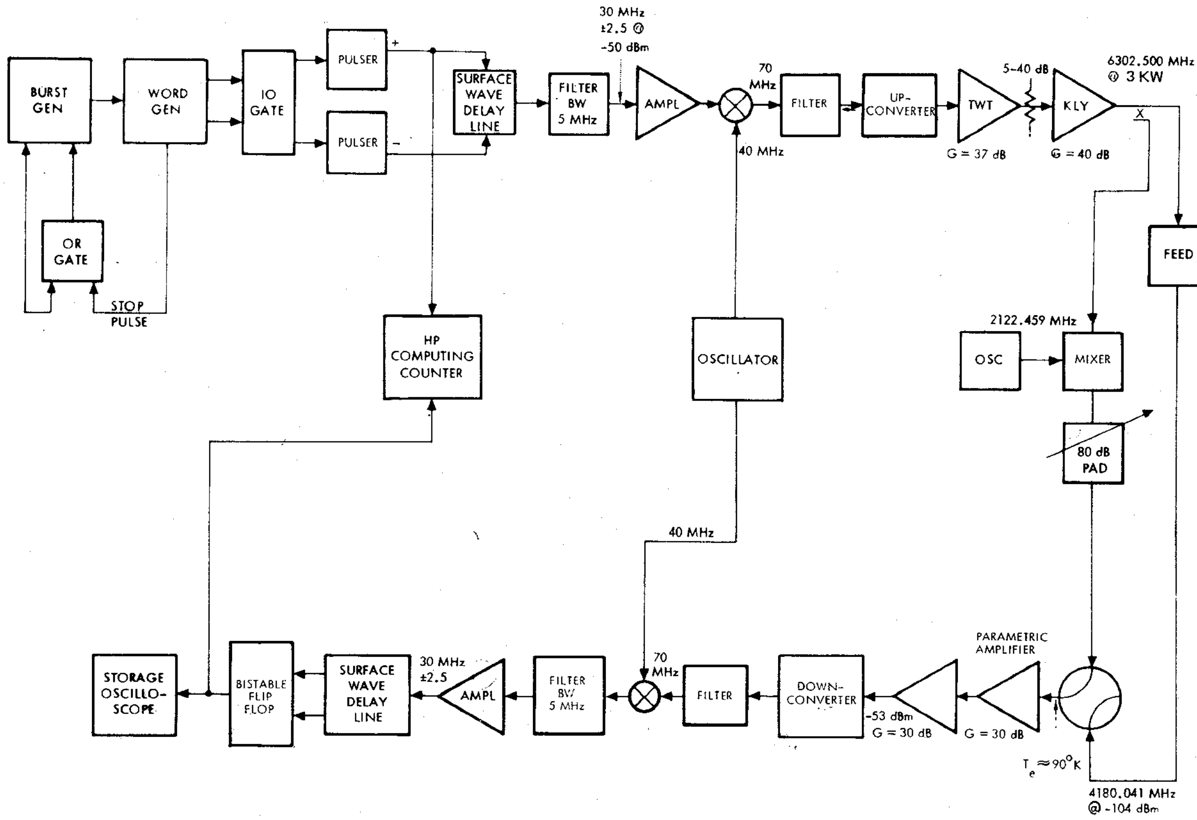

Fig. 7. Block diagram of earth station for chirp experiment.

The compression ratio is 280. The output of the delay line was up-converted to an IF frequency of 70 MHz and then translated to 6302.500 MHz where it frequency modulated the output power of the transmitter. A 30-dB directional coupler was again used to tap off the output signal for calibration of the delay through the ground station. As in the tone ranging experiment, the group delay through the satellite was neglected in the range calculation. The received chirp signal from the satellite was routed into the parametric amplifier and down-converted to 70 MHz. It was translated to 30 MHz and received by the negative-slope delay line. The compressed pulse was then transmitted to a bistable flip-flop which triggered a Hewlett-Packard Model 5306A computing counter. This counter measured the time difference between the pulses supplied to drive the transmit chirp filter and those that were compressed by the receive chirp filter.

A link calculation for the chirp signal predicts a signal-to-noise ratio of 31.5 dB after compression, assuming satellite transmitter saturation. The one-way range error due to noise on the dechirped signal is given by:

$$\Delta R = \frac{\sqrt{3C}}{2\pi B\sqrt{2\, S/N}} \tag{2}$$

where ΔR equals one-way range error, C equals the speed of light, B equals the bandwidth of the chirp pulse, and S/N equals the received signal-to-noise ratio after compression by the chirp filter. Using the above formula [6], and the calculated signal-to-noise ratio, the theoretical range resolution for the link is 0.4 m. Fig. 8(a) displays an oscilloscope picture of the receive signal after pulse compression. The satellite has transponded and translated the chirp signal to the down-link frequency. The earth station transmitter had an output power of approximately 3 kW. The received signal-to-noise ratio is approximately 30 dB, which is in good agreement with the link calculations. The envelope of the power lines indicated in Fig. 8(b) represents the power spectrum of the chirp pulse.

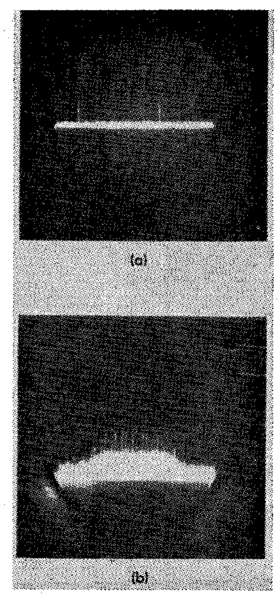

Fig. 8. Returned signals from satellite before and after compression. Transmit power, 3 kW. (a) Compressed pulse oscilloscope setting: 0.5 V/div; 50 μs/div. (b) Power spectrum of chirp signal above. Scale settings 1 MHz/div–horizontal; 10 dB/div–vertical IF bandwidth–100 Hz.

The sweep rate of the spectrum analyzer was adjusted for too few samples. Note the noise pedestal envelope which is approximately 20 dB below the chirp power spectrum envelope. This agrees with the link power budget calculations.

Returned signals were observed with transmitter powers of 300 and 75 W, respectively. The satellite output power did not reduce quite linearly in proportion to the earth station transmitter power, since the 3-kW figure causes the satellite transmitter to operate on the base of the saturation curve. At the lower power level, the spectrum of the chirp signal was not discernible from the noise. This feature of the chirp signal allows simultaneous chirp and other modes of information transmission with minimal intermodulation. During these experiments, a 27.7-kHz tone was FM modulated onto the carrier and transmitted through the satellite. The FM signal was demodulated by the receiver with no noticeable effect.

The chirp range measurements obtained on December 23, 1971, at 17:52 GMT, using a system calibration of 62.548×10^{-6} s, and measured time of flight of $251.263045 \times 10^{-3}$ s provided a range (R) of 40 637.097 nmi. The measured range ($R/2$) was 20 318.548 nmi as compared to the R (ATS-1) predicated from the GRARR of 20 325.8 nmi. The measured range of 20 318.548 nmi is well within the accuracy of the predicted range using the GRARR data taken one month previous to the chirp measurements.

Further measurements during June of this year, taken during 10-s intervals for periods of 8 min were used to fit the two-way satellite range to a second-order polynomial in time. A second-order polynomial was chosen since the satellite acceleration is almost constant over small time intervals. The standard deviation of this curve was approximately 3 ns, and is indicative of the experimental resolution of this particular chirp ranging system. Sequential measurements by the chirp and tone range systems indicate agreement to within a few hundred feet for satellite range. This slight disagreement is assumed to be due to errors in calibration of the ground station.

The delay time induced by a chirp filter is linearly dependent on the frequency. Hence, if the RF carrier has been Doppler shifted by motion of the satellite, the time at which the Fourier components are compressed by the delay line will also be shifted proportionately. The time shift is proportional to the slope of the delay versus frequency for the particular chirp filter. The first-order Doppler shift is proportional to twice the range rate of the satellite, and therefore the time shift of the compressed pulse is given by:

$$\Delta T = \frac{2vf_c}{C}\left(\frac{dt}{df}\right) \tag{3}$$

where v is the range rate of the satellite, C is the speed of light, f_c is the carrier frequency, and dt/df is the slope of delay line in seconds/hertz.

Since a positive chirp pulse is compressed on a negative-slope delay line and vice versa for a negative chirp pulse, the two types of pulses are time shifted in different directions. The error produced in the range measurement can thus be accounted for by transmitting positive and negative chirp pulses separated by a known time and measuring the separation time between these two pulses when they are recompressed. This technique can be used not only to correct the propagation time for Doppler shift, but to measure the range rate directly. The separation time between positive and negative chirp pulses will be increased by:

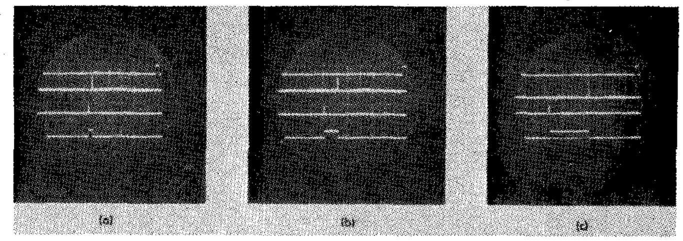

Fig. 9. Effect of frequency shift to positive- and negative-slope chirp pulses.

$$\Delta T_S = \frac{4v}{C} f_c \left(\frac{dt}{df}\right). \tag{4}$$

It has been assumed that the negative and positive chirp filters have equal but opposite slopes. The term ΔT_S is the time shift in the separation of the pulses. The magnitude of the range-rate shift that can be detected utilizing this method is given by:

$$v = \frac{\Delta T_S C}{4f_c}\left(\frac{df}{dt}\right). \tag{5}$$

For the range experiment reported on here, the minimum detectable ΔT_S was approximately 1 ns, the average carrier frequency f_c was 5 GHz, and delay line slope df/dt is given by $df/dt = 5$ MHz/50 μs $= 10^{+11}$ Hz. The range-rate shift or resolution obtained from (5) is therefore $v = 1.5$ m/s. Consequently, a range rate of 15 m/s will produce a shift of 10 ns, which can be easily measured.

This experiment was not attempted at the time the satellite was ranged; however, the effects of a frequency shift were observed when the local oscillator for the up-converter in the earth station was shifted by a few megahertz. The second trace of Fig. 9(a) is a received negative chirp pulse while the third trace is a received positive chirp pulse from the satellite. The width of the square pulse on Trace 1 represents the time separation between the two chirp pulses at transmit and the width of the pulse on the fourth trace represents the time separation between the received pulses. Fig. 9(b) and (c) displays these same traces with a progressive increase in the local oscillator frequency of the up-converter. This increase simulates the Doppler effect. Note that the compression of the positive pulse is advanced and the negative pulse retarded by a positive Doppler shift. The opposite is true of a negative Doppler shift. This shift can be used as a correction factor during ranging operations.

C. Comparison of Tone and Chirp Ranging

Both of the two methods that were used to range a synchronous satellite performed successfully. No decision could be made as to which method is experimentally more accurate since the predicted range to the satellite was only accurate to about 20 nmi. Both methods of ranging have the same potential precision utilizing the same bandwidth. The range resolution using the tone system is inversely proportional to the highest tone frequency and the square root of the signal-to-noise ratio. Hence it should be noted that the resolution is

TABLE I

COMPARISON OF MULTITONE CW AND CHIRP RANGING TECHNIQUES USED

Characteristic	CW	Chirp
Range measurement	Indirect measurement of phase delay of highest tone with ambiguity in phase removed by lower tones	Direct time delay measurement using FM dispersed and compressed pulse
Accuracy (function of frequency f, S/N)	1 m $(f\sqrt{S/N})^{-1}$	0.4 m $(\Delta f\sqrt{S/N})^{-1}$
Doppler sensitivity	Detected by Doppler shift of carrier or modulation	Detected by plus and minus chirp slopes—time interval measurement (~1.5-m/s resolution)
Equipment stability calibration (time delay)	Temperature stability of large time delay narrow-band filters requires routine test of station time delay.	Zero temperature coefficient surface-wave filter—essentially a time–frequency standard minimizes calibration
Receivers	Linear-phase precision-stabilized receiver	Linear-dispersive matched filters integrated with standard receiver components
Processing gain	Narrow-band filters in baseband, with averaging of zero crossings	Direct IF addition in chirp filter "in-phase" components provides 20–30-dB gain
Frequency modulation	Low FM deviation used to minimize time delay errors in system	Large time–bandwidth (efficient box spectrum)

more sensitive to tone frequency than signal-to-noise ratio. The tone method of ranging uses narrow-band filters in the baseband and averaging of numerous axis-crossing measurements to obtain processing gain. Narrow-band filters are generally characterized by large group delays, and hence must be carefully designed in order to minimize errors due to shift of the center frequency of the filter with temperature. Routine calibration of these filters is necessary to update the measured earth station delay. The tone method also requires a fairly expensive tone generator that can maintain a precisely known frequency. Since the range is measured in terms of the period or the wavelength of the highest tone, and since the synchronous range contains large numbers of these wavelengths (approximately 7000 for the 27.7-kHz tone), the error caused by frequency instability in the tone generator is multiplied by this large number. This problem does not occur with the chirp method of ranging used here since the pulse that triggers the timing counter is the same pulse that is delayed by the propagation time and turns the counter off. Range errors introduced by frequency shifts caused by Doppler effect or local oscillator drift can be eliminated by transmission of both positive- and negative-slope chirp signals. The tone method also requires a fairly expensive FM or PM receiver.

The chirp method measures the range directly with no possible ambiguities as inherent in the tone method. The resolution of the method is inversely proportional to the bandwidth of the chirp filter and the square root of the signal-to-noise ratio. Hence the range resolution is more sensitive to the bandwidth of the chirp filter than the compression ratio. The chirp signal is wide band and appears like noise so that interference with other modes of information transmission on the same channel are reduced. Chirp surface-wave filters, although with increased cost, can also be designed at RF frequencies so that no IF strip is necessary. In addition to ranging, the chirp signal can also be used to transmit information. The accuracy of both ranging methods is more dependent on the systematic errors of the ground station and satellite calibration, tropospheric, and atmospheric corrections rather than random errors introduced by noise. Considerable care must be taken to reduce these systematic errors. Modeling of the troposphere and atmosphere can account for path bending and velocity changes.

Table I, summarizes the significant remarks, appropriate for consideration in comparing the multitone, CW tone, and the chirp ranging techniques.

IV. CONCLUSION

The two significant applications discussed in this paper, namely chirp ranging and data transmission via a synchronous satellite, provide examples of the practical usefulness of the surface acoustic wave filters using the chirp waveform. The design flexibility of these devices, with their relatively low cost, stability, linearity, and dynamic range, makes a number of interesting future applications practical. The applications discussed here have indicated that the chirp coded waveform offers significant advantages over both the presently used multitone CW ranging system, and the more conventional binary phase coded transmission.

The state of the art developments in surface acoustic wave dispersive delay lines will facilitate in the future the implementation of chirp ranging systems with wider bandwidth filters (100 MHz) operating at higher data-transmission rates. Surface-wave chirp filters now considered available for practical implementation fall into two categories: 1) those providing dispersive time delays of 100–150 μs with bandwidths of up to 20 MHz; and 2) those with dispersive bandwidths of 100 MHz and time delays of 20 μs on piezoelectric quartz substrates.

In the next generation, low-cost reliable techniques for international synchronization of airborne time standards can be anticipated by using chirp carrier systems with surface acoustic wave filters. Other developments anticipated are improved aircraft-to-ground communications via satellite through chirp coding, and air-traffic control of aircraft using improved time-ordered sequencing systems with digital encoded chirp. In addition, the digital encoded chirp data transmission, with nonoverlapping carriers, is anticipated to be utilized more, as the need arises, for less spectral pollution of the communication bands. The asynchronous operation of these chirp filters with nominal processing gains of 30 dB and modest bandwidths exhibits exclusive advantages over the digital matched filter, with its inherent Doppler insensitivity, and should provide attractive compelling advantages in competing with present communication transmission techniques.

ACKNOWLEDGMENT

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A Review of Device Technology for Programmable Surface-Wave Filters

EDWARD J. STAPLES AND LEWIS T. CLAIBORNE

Invited Paper

Abstract—**A review of programmable surface acoustic wave filters is presented. The elementary theory, fabrication procedures, and device performance are described in light of recent technological advances. Both hybrid and monolithic structures are considered. The relative advantages of programming techniques which utilize diode switching are compared to those that make use of solid-state, three-terminal, and acoustic wave detectors.**

I. Introduction

PRACTICAL surface acoustic wave devices are a reality for a variety of VHF and UHF filter and delay-line applications, e.g., bandpass filters [1], [2], pulse compression filters [3], and biphase coded matched filters [4]. The devices which have received the most attention and have been most extensively utilized for current systems hardware are based on Rayleigh wave propagation in piezoelectric substrates with fixed pattern interdigital transducers. For future systems needs [5]–[7], it will often be desirable to have available filters whose transfer functions are electronically programmable. It is the purpose of this paper to review the techniques which have been under development to provide surface-wave devices which can be programmed in real time for a variety of impulse responses.

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In its simplest form, the programmable surface-wave filter consists of a delay-line substrate which has a wide-band input transducer and a series of taps which are located at arbitrarily specified intervals down the propagation path. Parallel summation of the output from each tap forms a transversal filter [6] configuration, and in the general case, both the amplitude and phase of each tap are programmable variables. In this paper, we shall limit our discussion to surface-wave filters which have equispaced and constant amplitude taps, the programmable function being only the phase at each tap.

Most of the programmable filter development has been directed at biphase pseudonoise (PN) sequence matched filters for use in applications such as spread spectrum communications, ranging, and identification sequence generation and recognition [7]. The discussion in Sections II and III is concerned primarily with filter configurations which could be used in these applications.

The key component in all programmable filters is the acoustoelectric transducers used for signal taps. Methods of implementing diode-switched phase programming with the interdigital transducer on piezoelectric substrates are discussed in Section II; following this in Section III is a discussion of solid-state techniques for implementing other programmable transducers including the Si–metal–oxide–semiconductor field-effect transistor (MOSFET) and piezoelectric