SAW Quadraphase Code Generator

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Abstract—Radar emmissions are now subject to regulation that limits their spectral splatter. Quadraphase codes are used to make an MSK-like pulse that has a narrower spectrum than a biphase pulse, and furthermore, it is tolerant of filtering. The design of a SAW pattern is described which yields a device that is both an encoder and filter. When it is energized by an impulse a valuable, spectral limited radar pulse is generated. The performance of this pulse is shown to meet certain desirable criteria. The correlator for the pulse is simply derived from the code generator and its performance is shown. Comparison with non-SAW methods of performing the same functions is given.

I. The Need for Quadraphase

Biphase encoded radar pulses exhibit excessive spectral splatter due to almost instantaneous phase reversals which occur when bit polarity changes. These pulses then fail to meet the requirements of the Office of Telecommunications Policy (OTP) [1] which wants to limit the spectral width of radar emmissions. Seeking to bring radar emmissions into compliance with their regulations, J.W. Taylor proposed the transmission of four-phase codes instead of biphase codes, but he proposed that the codes be interpreted as continuous phase shift instead of step phase shift sequences. Communications engineers will recognize such a modulation as frequency shift keying. The distinction in the radar use is that the encoding is not applied to a continuous bit stream but is limited to the duration of a typical short radar pulse. More particularly, the encoding described here is minimum shift keying (MSK), a form of offset-keyed quaternary phase-shift-keyed modulation [2] (OK-QPSK).

This signal can be synthesized by summing in-phase and quadrature encoded signals. Each of these signals is biphase encoded and, additionally, the bits are sinusoidally weighted, not rectangular. They are summed after delaying one signal by half a bit length and shifting its carrier by 90°. This is illustrated in communications literature several places; in particular, Pasupathy's [3] article shows it well. Conventional generation of quadraphase encoded pulses literally follows the recipe above. This requires much circuitry, modulators, logic, etc. A typical encoder has 350 components, consumes 11 W of prime power, and occupies more than 100 in^3.

II. The Better Approach to Encoding

By properly interpreting the resulting quadraphase encoded pulse, a simpler and more direct implementation of a quadraphase encoder can be realized with a SAW delay line. In fact, the result of performing the summation mentioned above is, on a finite length basis, a pulse having certain favorable features.

1) It does not rise from zero level instantaneously.

2) During a bit, the phase of the sine wave in the bit shifts linearly over a span of 90°.

3) The direction of these shifts (+ or −90°) is governed by a chosen biphase code sequence.

4) The linearly changing phase shifts mean that the encoded sine wave is allowed to exhibit only two frequencies during the encoded portion of the pulse; these are given by some center frequency, f_0, + and $-\Delta f$, where

$$\Delta f = \frac{\text{phase shift per bit (part of a cycle)}}{\text{bit length (seconds)}}.$$

All these features can be synthesized directly in a SAW interdigital pattern.

The beginning of one of these devices requires a choice of coded sequence. The peak-to-sidelobe ratio of the autocorrelation function is the principal (but not only) feature governing the choice of code. Besides the Barker codes, shown for instance in Blinchikoff and Zverev [4], Turyn [5] has published tables of "good" codes up to length 34. These biphase sequences can be converted to n-phase sequences using the procedure of Golomb and Scholtz [6]. For converting biphase codes to quadraphase codes, each kth biphase bit is multiplied by $j^{(k-1)}$. The longest Barker code is the 13 bit sequence $1,1,1,1,1,-1,-1,1,1,-1,1,-1,1$. This converts to $1,j,-1,-j,1,-j,1,-j,1,-j,-1,j,1$ in quadraphase using the algorithm above with the conventional identity $j=\sqrt{-1}$.

At the top of Fig. 1, the quadraphase sequence is shown. Just below that, the phase history of the pulse is plotted. The step-shaped figure is the phase response if the sine wave were to dwell at each phase state; this is true OK-QPSK. The spectral spread of such signals is no better than biphase encoded signals. The compression of spectral spread is brought about by letting the phase change linearly during the bit as shown by the "continuous" phase plot; this is the feature that makes this encoding MSK.

MSK has lower spectrum sidelobes than OK-QPSK, furthermore, the radar pulse that we will generate can be filtered because MSK is tolerant of filtering. The filter will not introduce high amplitude ripples, which a radar transmitter would find very undesirable; it will only introduce a slow rise and fall function. The bottom diagram of Fig. 1 plots the frequency history during the pulse when the phase is changing linearly; a radar pulse exhibiting this feature of Fig. 1 satisfies the four points given above.

To design a device that yields this frequency history on a

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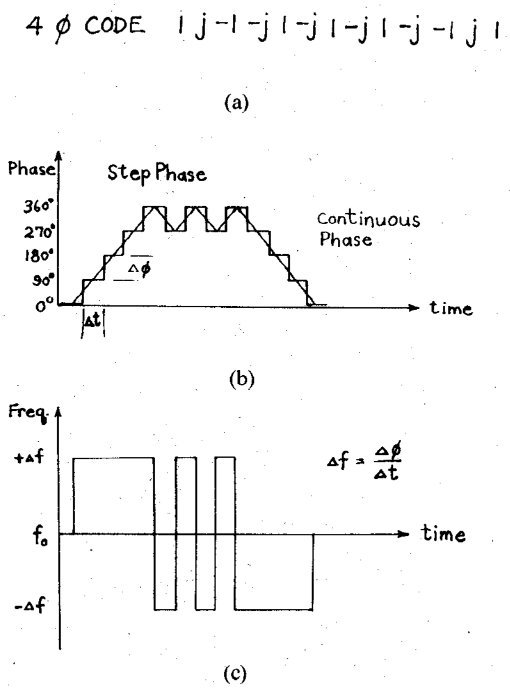

Fig. 1. The quadraphase code and its continuous shift interpretation. (a) The four-phase Barker sequence. (b) The step and continuous phase history. (c) The frequency history derived from continuous phase interpretation.

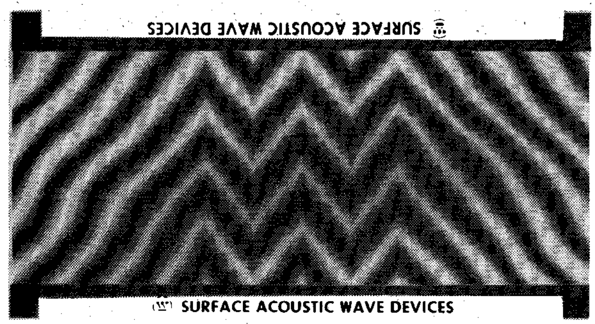

Fig. 2. The Moire pattern formed by a mask and its reverse shows the continuous phase pattern of Fig. 1(b).

SAW substrate is straightforward. If the chosen substrate has surface wave velocity v_s then standard interdigital finger patterns are designed with the space width = finger width $= v_s/(4(f_0+\Delta f))$ during the first four bits of the encoded pulse. An impulse wavefront passing under this interdigital grid will yield a sinusoidal output signal of frequency $f_0+\Delta f$. During the next bit, the interdigital finger and space dimension will be $v_s/(4(f_0-\Delta f))$; an impulse passing under this pattern will yield a sinusoidal signal having frequency $f_0-\Delta f$. The design of the encoding grid continues until the frequency inside each bit has been simulated with an interdigital finger pattern.

Quartz is a commonly available, inexpensive substrate material with many desirable features. It is temperature stable to within ± 0.01 percent of f_0 from $-55°C$ to $+95°C$. $v_s = 3157$ m/s for this temperature-stable orientation (*ST* cut). If the bit length Δt of Fig. 1 is 1/2 μs, and $\Delta\phi$ is 90° (1/4 of a cycle) then $\Delta f = 1/2$ MHz. At $f_0 = 30$ MHz, the finger and the space between dimensions would be 25.8771 μm for the high-frequency bits and 25.75425 μm for the low-frequency bits. A half-bit at the center frequency f_0 begins and ends the pattern. Our approach is to make a blowup of this desired array. This can be made to include a filter launching pattern whose impulse response is the duration of a bit and whose apodization yields a functional shape that effectively reduces the spectral width of the quadraphase encoded pulse. The finger and space widths of the filter are at the center frequency f_0.

A large computer-generated pattern was made on a Gerber plotter to fit on the copy board of a reduction camera. The burden on the Gerber plotter was to represent accurately finger spacings corresponding to a center frequency f_0 (30 MHz in our case), and frequency excursions of $\pm\Delta f$ about f_0. Measurements on the full size mask indicated that when photoreduced, the desired frequency shift of ± 500 kHz could be realized with an error or less than 600 Hz. The changes in finger spacing are so small that they cannot be detected visually. To aid in showing the frequency shifts and the intervals over which they exist, the Moire patterns are formed by reversing one mask and laying it on top of another. Fig. 2 shows the characteristic interference pattern; indeed, this Moire pattern is the continuous phase versus time pattern of Fig. 1(b). The Moire pattern is functionally equal to the phase history only when the quadraphase sequences are symmetrical; this is the case for all the Barker codes (except the four unit) and some of Turyn's desirable codes as well.

III. Fabrication and Performance

The large pattern was accurately photoreduced to make a positive chrome photomask. Standard photoresist over 1000 Å of aluminum on *ST* quartz was exposed to UV light through the mask. Development and etching yields the device shown in Fig. 3. On the left is the launcher pattern which can be seen to have gaps in some fingers to yield a half-sine apodization. This then is a filter whose impulse response has half-sine shape and is 1/2-μs duration. The effect of the filter is to give integral-of-sine rise and fall shape to the 4ϕ encoded pulse; this serves to limit the spectrum of the encoded pulse. The encoding pattern is the large array to the right. Applying an impulse shorter than or equal to 1/2 a sine wave of the carrier f_0, starts a weighted acoustic wave train traveling left to right, which passes under the output array. The resulting encoded output pulse and its rise time is shown in Fig. 4. It is significant that the pulse is not obviously encoded because the phase transitions are smooth, not abrupt. This contrasts starkly with a biphase encoded pulse where the phase transitions are obvious [7]. Certain features of the pulse deserve comment. The small pulse preceding the encoded pulse by 1 μs is the electromagnetic feedthrough. This can be reduced to insignificance by proper shielding and grounding. The taper of the encoded pulse is due to energy under the encoding pattern reducing while the traveling wave is propagating under the pattern. This is correctable by putting a reverse taper in the finger overlap of the encoding pattern. The ripples on top are due to the finite bandwidth of the launcher-filter. The feedthrough and trailing time spurious (all more than 25 dB down) will be

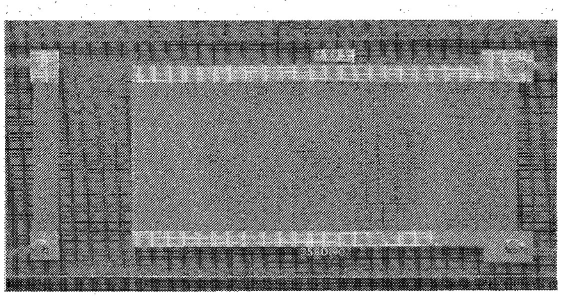

Fig. 3. Aluminum pattern on quartz substrate is the heart of the quadraphase encoder.

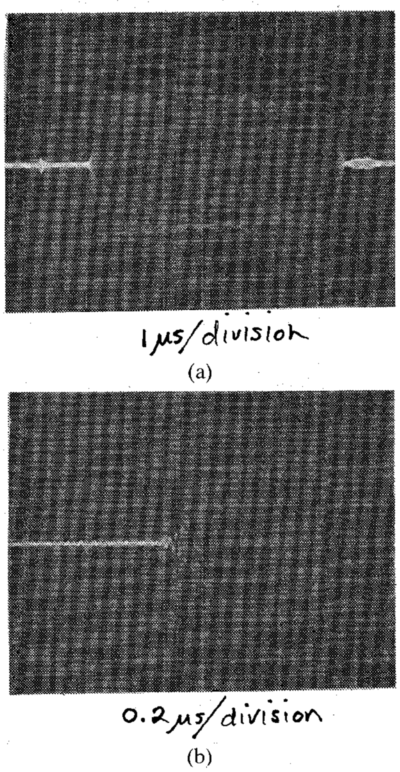

Fig. 4. Impulse response of quadraphase encoder. (a) Entire pulse showing feedthrough of the initiating impulse. (b) Rising edge.

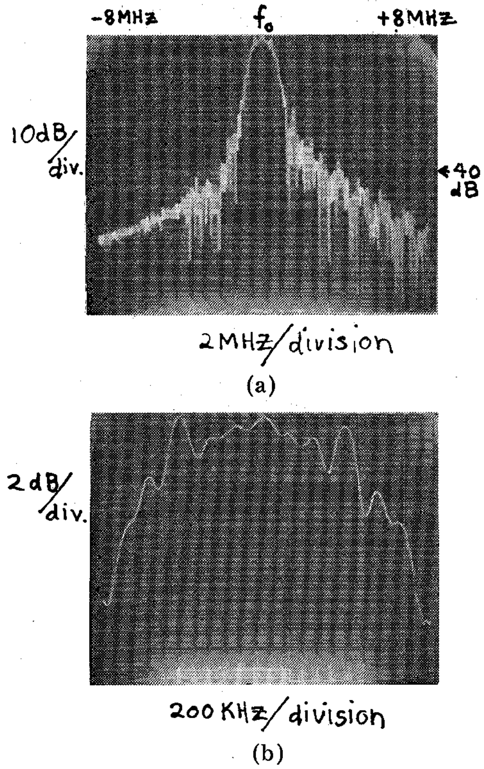

Fig. 5 Spectrum of the quadraphase pulse. (a) Broad view. (b) Detail in the center.

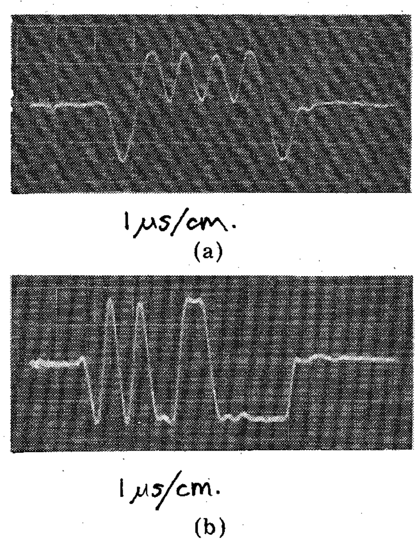

Fig. 6. Phase detector outputs (a) when compared to 31.0518 MHz; (b) when compared to 31.5815 MHz.

eliminated when the encoded pulse is enclosed in a gating signal.

The spectrum of the pulse is shown in Fig. 5. It easily meets the OTP [1] requirement that energy be 40 dB down before ± 8.2 MHz thus confirming the value of the 4ϕ code.

Measuring phase versus time through the pulse is done by comparing it with a coherent sine wave in a phase detector. The resulting filtered output pattern is shown in Fig. 6(a). Notable here are the ledges at the beginning and the end of the phase pattern. These are caused by the fact that during the rise and fall times, the pulse contains sine waves at the center frequency. The excursions in frequency only occur over the middle portion of the pulse. This pattern is used to separate a possible phase and frequency ambiguity. The frequency to which the pulse is compared is varied until there is no tilt to the phase pattern. This can be done accurately to within ± 1 kHz out of 30 MHz. Phase can then be independently varied to yield a sequence of expected phase patterns.

Offsetting frequency by about 1/2 MHz yields the pattern of Fig. 6(b). This is remarkable since the pattern is the original biphase Barker sequence of 13 bits 5,2,2,1,1,1,1 (in Turyn's notation) given above. The insignificant difference is a polarity change and a time reversal.

The actual measurement of relative phase jitter between the sine waves in the SAW encoded pulse and sine waves from which the impulse is derived has not been done.

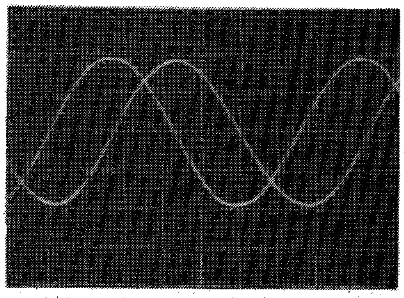

Fig. 7. Encoded sine wave compared to original STALO sine wave.

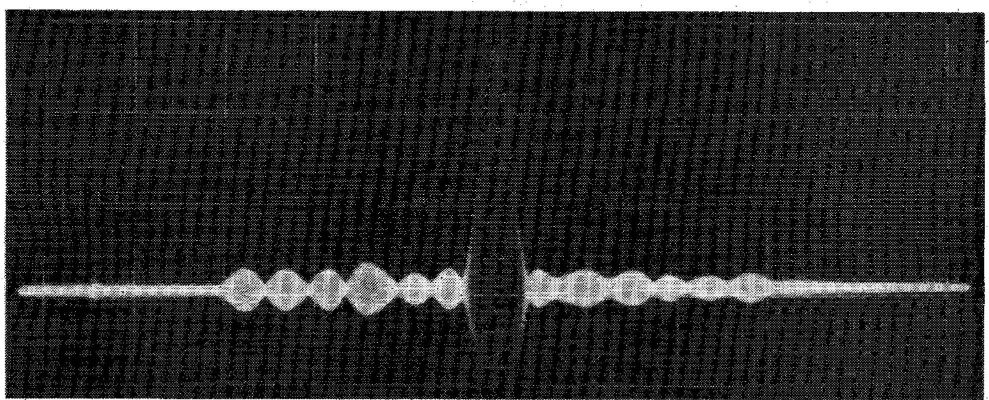

Fig. 8. Output from a SAW quadraphase decoder when the input is generated by a separate quadraphase SAW encoder.

However, Fig. 7 compares these two sine waves on a 5-ns/cm scale. In fact, nothing is detectable here; but when the scale is expanded horizontally to 1 ns/cm, a jitter *in both* sine waves of about 0.1 cm is detectable. This jitter must be about 100 ps; then the relative jitter must be less. Exactly what the relative phase jitter is must still be resolved.

A. Decoder

The decoder for this signal looks identical to the encoder; the difference is that the frequency sequence shown at the bottom of Fig. 1 is reversed. If decoding is a measure of the encoding, then Fig. 8 indicates that the surface-wave encoder and decoder are almost ideal. The peak-to-sidelobe ratio is 21 dB (that is, the third lobe before the mainlobe; all others are more than 22 dB down). The uneven level of the sidelobes is due to a combination of filtering in the encoder and decoder, and to the taper in the generated code. With a 50-Ω source and load impedance the decoder insertion loss is only 13 dB.

IV. Comparing

A traditional encoder was described in the beginning of this paper. It requires a separate spectrum control filter. The surface-wave encoder only requires a simple impulse as an input. It consumes 5 W, has 90 components, and occupies 20 in^3 and the filter is integral with the encoder.

SAW decoding contrasts even more with conventional methods. A digital decoder is a planar array which naturally requires power to operate. It occupies 96 in^3 and consumes 17 1/2 W. The SAW decoder is a passive device with low insertion loss requiring no power input at all. It occupies 0.7 in^3.

Conclusion

Matthews [7] shows several references to previous work in generating coded signals by impulsing SAW filters. Here we have shown that valuable radar pulses can be generated by impulsing SAW devices and that proper choice of encoding is crucial to meeting regulatory restrictions while still maintaining desirable properties required by radar. Once an encoding pattern has been made, a corresponding decoding pattern is simply derived from it. The required spectrum control filtering is integrated into the SAW device.

Besides resolving the phase jitter question mentioned above, there still remain some other questions to be answered. It will be necessary to know what frequency difference can be tolerated between an encoder–decoder pair before the autocorrelation characteristics degrade. Also, it is not yet clear that the minimum insertion loss has been attained. There are also possible procedures for generating initiating impulses that are inherently more immune to phase jitter. It is intended to continue our investigations along these lines.

Acknowledgment

The author would like to acknowledge many illuminating discussions with R. M. Morrison and H. J. Blinchikoff regarding quadraphase encoding and its satellite problems. Also, the technician skills of E. Heitman deserve thanks.

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Precision SAW Filters for a Large Phased-Array Radar System

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***Abstract*—The electronically steerable radar (ELRA) at the Forschungsinstitut für Funk und Mathematik is an experimental *S*-band phased-array radar system consisting of separate transmitting and receiving arrays employing several coherent and incoherent signal-processing and data-handling techniques, incorporating multiple beam and multifunction operation for target search and tracking, adaptive interference suppression, and target resolution.**

This paper deals with the development and application of two types of SAW filters for the IF amplifier channel of the receiving array. Compared to conventional filters with lumped elements, these filters have some important merits. By making use of a special tuning technique, the center frequencies of all filters were adjusted, resulting in an rms deviation of less than 1 kHz. One type of the SAW filters represents an almost ideal approach of realizing a matched filter for rectangular shaped pulses. The conformity of the frequency responses of several hundred filters improved the noise suppression capability of the system. The use of the filters described represents one of the applications where high-quality mass-produced SAW devices have been applied to improve system reliability and performance.

I. Introduction

The primary aim of the ELRA system is to prove the practical feasibility of theoretical studies of signal processing, data management, and system control in a multitarget environment, in connection with phased arrays. To achieve highest flexibility for these widespread demands, separate antennas with active modules for transmitting and receiving have been chosen. The following description of the system should only give an overview. More detailed information on the architecture and specifications can be found elsewhere [1]–[3].

Three-hundred printed dipoles, randomly distributed on planar circular apertures of 28 wavelengths diameter, and 768 dipoles on 39 wavelengths diameter, respectively, form the transmit and receive arrays.

Each transmitter dipole is fed by its own module consisting of a 3-bit diode phase shifter followed by a triode amplifier. One of the 768 receiver channels is illustrated in Fig. 1. A bipolar transistor preamplifier in connection with an image rejection mixer assures a low noise figure of about 3 dB. The IF amplifier uses either a narrow-band or a broad-band SAW filter adapted to the transmitter pulse length of 10 μs for search and 2 μs for tracking, respectively. The filtered IF signals are converted down to baseband by a synchronous detector with two orthogonal components for coherent signal processing. The IF reference is fed to the mixer through a 3-bit phase shifter which consists of an 8-channel analog multiplexer connected to a tapped delay line. The outputs of 16 neighboring elements are combined in summing operational amplifiers and converted from analog to digital representation. Parallel trees of binary adders combine these subarray signals forming

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