1984

12. BINARY PULSE COMPRESSION TECHNIQUES FOR MST RADARS

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INTRODUCTION

In most MST applications pulsed radars are peak power limited and have excess average power capability. Short pulses are required for good range resolution, but the problem of range ambiguity (signals received simultaneously from more than one altitude) sets a minimum limit on the interpulse period (IPP). Pulse compression is a technique which allows more of the transmitter average power capacity to be used without sacrificing range resolution. As the name implies, a pulse of power P and duration T is in a certain sense converted into one of power nP and duration T/n. In the frequency domain, compression involves manipulating the phases of the different frequency components of the pulse. A short pulse consists of contributions from a wide band of frequencies, all of which are in phase at one point in space-time. Changing the phase relations on transmission lengthens the pulse, but it can be reassembled into a short pulse upon reception by proper processing <u>if</u> the phase has not been perturbed in some unknown way in the meantime (i.e., by the scattering process). This is essentially the idea behind frequency "chirping".

Another way to compress a pulse is via phase coding, especially binary phase coding, a technique which is particularly amenable to digital processing techniques. This method has been used extensively in recent years in radar probing of the atmosphere and ionosphere, and it is the method we will discuss here. The general topic of pulse compression is dealt with in COOK and BERNFELD (1967), BARTON (1975), BROOKNER (1977), and other texts. The operation performed on reception to produce the compression of phase coded radar signals is referred to as decoding. The signal is decoded upon reception by passing it through a "filter" whose impulse response is the reverse in time of the transmitted pulse (the pulse "played backwards", so to speak). Such a filter is said to be "matched" to the pulse. In practice these matched filters are usually specially designed acoustic surface wave devices or conventional filters plus digitizers, digital delay lines, and some add/subtract circuitry or equivalent software.

From another point of view, the decoding process consists of cross-correlating the received signal with a replica of the transmitted pulse; hence, if an echo from a single scattering target is passed through such a decoder, the output is the autocorrelation function (ACF) of the pulse delayed by a time corresponding to the distance to the target. As an example, the phase coding sequence and the ACF of a 5-baud Barker coded pulse are listed below.

+ + + − +

. . . 0 0 0 1 0 1 0 5 0 1 0 1 0 0 0 . . .

If the compression process were perfect, only the 5 would be present in the above ACF; the 1's represent undesired range sidelobes.

In this paper we discuss many aspects of codes and decoding and their applications to MST experiments. This includes Barker codes and longer individual codes, and then complementary codes and other code sets. We discuss software decoding and also hardware decoders and coherent integrators.

BARKER CODES

A class of codes known as Barker codes (BARKER, 1953) has been used extensively in ionospheric incoherent scatter measurements. The Barker coded pulse is considered to be made up of n "bauds", each of duration T, so the total duration is nT, with the maximum value of n being 13. The phase of each baud is 0 to 180 degrees (±1), in a sequence that depends on n.

We have seen above the properties of a Barker code of length 5. A property of all Barker codes is that the center peak of the correlation function is n, and the sidelobes are either 0 or 1. For ionospheric application the sidelobes are generally not a problem since, for n equal 13, say, the power corresponding to the central peak is 169 times greater than that in each of the 12 sidelobes. (Note that the signal-to-noise ratio in the central peak is increased by the compression by a factor of 13, not 169, since the noise is the sum of 13 independent samples.)

The above discussion is valid for scatter probing of the atmosphere as long as the correlation time of the scattering medium is long compared to the total (uncompressed) duration of the coded pulse. In practice this is always the case of MST applications but may not be true for incoherent scatter from the ionosphere, for example. Detailed calculations of what happens in the latter case are given by GRAY and FARLEY (1973), and a general discussion of the "ambiguity function" of a Barker coded pulse as a function of target induced Doppler shift is given in COOK and BERNFELD (1967). Gray and Farley also discuss the use of multiple coded pulse sequences in the measurement of the ACF of the scattering medium. The effect of the coding is usually minimal; in typical situations the "true" ACF is convolved with a function whose width is about one baud. Finally, although 13 baud is the longest possible Barker sequency (unity sidelobes), there are many longer sequences with sidelobes that are only slightly larger. As an example, a 28-baud sequence with a maximum sidelobes level in the ACF of 2 is listed by Gray and Farley. In fact, this code sequence was used in the first application of phase coding techniques in MST radars (Woodman, MPI-SOUSY (SCHMIDT et al., 1979) and the Arecibo 430 MHz radar (WOODMAN, 1980a).

Complementary codes are again binary phase codes and they come in pairs. They are decoded exactly as are Barker codes, by a "matched" filter/delay line combination whose impulse response is the time reverse of the pulse. The range sidelobes of the resulting ACF output will generally be larger than for a Barker code of comparable length, but the two pulses in the complementary pair have the property that their sidelobes are equal in magnitude but opposite in sign, so that when the outputs are added the sidelobes exactly cancel, leaving only the central peak; i.e., the compression is perfect. As the simplest possible example, consider the 2-baud complementary pair below.

Code:	+ +	(first pulse)
	+ −	(second pulse)
ACF:	0 +1 +2 +1 0	(first pulse)
	0 −1 +2 −1 0	(second pulse)
	0 0 +4 0 0	(sum)

Representing the above pair as (A,B) it is easy to show that the sequence (AB, $\overline{AB}$), where $\overline{B}$ is the complement of B, is also a complementary pair that is twice as long. Proceeding in this way one can obviously generate long n-baud code pairs, where n is any power of two. It turns out that n can also be ten, or ten times any power of two. Further properties of these sequences are given by GOLAY (1961). In the first reported measurements (SCHMIDT et al., 1979; WOODMAN, 1980) n was 32, compressing a 32 μsec pulse to 1 μs to achieve an altitude resolution of 150 m.

There are two practical limitations on the maximum value of the compression ratio n: (1) as n increases the effect of ground clutter extends to higher and higher altitudes; (2) the computing requirements for decoding increase with n. The first is the most serious limitation; the computing requirements usually can be handled one way or another. One process that often simplifies the computing is coherent integration (summing N successive voltage samples from a given altitude before doing any other processing). Since coherent integration and decoding are linear operations they can be interchanged (WOODMAN, 1980a); e.g., samples from 100 pulses, say, can be coherently integrated and then decoded all at once. In dealing with the first limitation one must achieve some compromise between three competing goals: (1) the desire to confine strong ground clutter effects to the lowest possible range of altitudes (i.e., use short pulses; (2) the desire to avoid range ambiguity (use a long IPP); and (3) the desire to use the full average power capabilities of the transmitter to achieve maximum sensitivity.

MORE COMPLEX COMPLEMENTARY CODING SCHEMES

WOODMAN (private communication, 1980) suggested that it might be useful to transmit more complex complementary sets in order to partly alleviate the ground clutter/range ambiguity problem. The cross correlation function (XCF) of the basic complementary transmitted sequence, A, B, A, B, ... with the decoding function A, B is periodic with a period 2T, where T is the interpulse period (between A and B), but there are also substantial nonzero values of the XCF in the vicinity of T. For example, for the 4-baud pair (+++- ++-+) the XCF is

... 0008000 ... 0040400 ... 0008000 ...

At delays near T from the "wanted" return, the range sidelobes of the individual pulses add rather than cancel, whereas the main peak does cancel. The 4's in the above represent the most important source of range ambiguity. These can be eliminated by transmitting the more complex sequence, A, B, A, $\bar{B}$, A, B, A, $\bar{B}$, ... and decoding by cross correlating with A, B, A, $\bar{B}$. The XCF for this scheme consists of single identical spikes at intervals of 2T; i.e., the first range sidelobe is pushed out to twice the interpulse spacing. By extending this idea the first sidelobe can be pushed out to even higher multiples of T. In this way a substantial range of altitude could be probed at a very high pulse repetition frequency (PRF). In actual practice, though, some altitudes would be lost because of the necessity of blanking the receiver during actual pulse transmission and because of receiver saturation by ground clutter. GONZALES and WOODMAN (1981) have used long sequences of complementary codes for a partial reflection experiment using the Arecibo HF-Heating facility. They were able this way to avoid multiple reflections from F and E region heights.

QUASI-COMPLEMENTARY CODE SETS

The results presented so far have all been based on the assumption that the transmitted pulses were perfectly coded. In practice of course this won't be true; the phase shifts will require a finite amount of time and will not be exactly 180 degrees, etc. As a result the range sidelobes for the complementary code pairs will not cancel exactly; the location of the sidelobes will depend on what sort of error is made by the transmitter. SULZER and WOODMAN (1983) have developed a technique to minimize this problem. Rather than transmit just a pair of complementary 32-baud codes, they transmit a sequence of 48 different

32-baud pulses. Each is decoded individually and the results are combined coherently, so in a sense the whole sequence can be considered to be a single code. But from another point of view we can think of the sequence as 24 quasi-complementary pairs, each with a different set of small range sidelobes, due partly to errors in transmission and partly to the fact that the pairs are not perfectly complementary. Because the sidelobes produced by the individual pairs have a more or less random distribution, the resultant sidelobes of the entire sequence are lower and more uniform than those of a single (imperfect) complementary pair. This is no accident of course; the codes were chosen by an extensive computer search requiring about 350 hours (1) using a Harris computer and an FPS AP120B array processor. The major disadvantage of this technique is that no coherent integration before decoding is possible; at present only the Arecibo Observatory has the digital preprocessing equipment required for the extensive high speed decoding.

A similar idea has been developed by the same authors for mesospheric observations at Arecibo. To achieve the desired resolution of 600 m (4 μs) and fully utilize the transmitter, one would ideally use a 52-baud Barker code, which unfortunately does not exist and 4 msec IPP to avoid range fading of F-region scatter. A good approximation to this can be achieved by a pseudorandom sequence of pseudorandom 52-baud codes found by a 10 hour computer/array processor search. Sidelobe power is not reduced to the level of a Barker code, but its coherence from pulse to pulse is destroyed by use of a different code on each transmitted pulse. The only range retaining the spectral information of the medium is the main lobe.

PSEUDORANDOM PERIODIC CODES

In bistatic experiments, one can transmit continuously, but one wishes to code and decode in such a manner that it is equivalent to transmitting a sequence of short pulses. Codes used for this are the so called maximal - length pseudorandom sequences which can be generated by relatively simple configurations of flip-flcps. These are excellent codes, having a uniform sidelobe level of -1 and lengths of any power of two minus one.

DECODING

As mentioned in the introduction decoding can be considered either as a matched filter operation, i.e., a convolution, with a filter which has an input response equal to the pulse shape, or as a cross correlation of the input with the pulse shape. We can perform this operation by proper software in a general purpose computer, in a special purpose processor or in an analog device (filter). Here we shall describe different techniques which have been used for this purpose as well as some new ideas that have been proposed with special advantage for MST radar applications.

SOFTWARE DECODING

Decoding, if done in a straight forward manner, would normally take several tens of operations per μsec and would be out of the possibilities of moderately priced computers. Decoding is used to obtain optimum resolution, which in the case of MST radars is of the order of a few hundred meters, this corresponds to one complex sample per one or few μsec. If we take a 32 baud code with a bandwidth of 1 μsec, as an example, straight decoding would take 64 adding or subtracting operations per μsec; certainly a requirement beyond the capabilities of even the fastest computers and a very demanding one even for specially built digital equipment. Fortunately, MST radars at VHF and lower UHF frequencies produce highly redundant information. Correlation times in the medium, and hence of the echo signals they produce, are of the order of a fraction of a second (≃0.5 to 2 sec for 50 MHz). This calls for coherent integration from pulse to pulse, when performed, reduces the amount of information by as much as two orders of magnitude.

Decoding and coherent integration are linear operations and, as pointed out by WOODMAN et al. (1980), they commute. One can perform the coherent integration first, and decode afterwards, with identical results. This possibility permits performing decoding operations in just about any mini- or micro-computer available in the market.

Coherent integration is so efficient in reducing the amount of information, that this can be put in a few tapes per hour of observation. In this case decoding, statistical processing and parameter estimation can then be performed off-line. This approach has been taken for the SOUSY radar - using a hardware integrator - (SCHMIDT et al., 1979) and by the same group at Arecibo with a portable 50 MHz transmitter.

Coherent integration usually requires a special purpose digital equipment, but can be done with fast array processors when used at the front end of the processing system. The M-mode at Millstone Hill uses this approach using an AP-120B for the integration and decoding (RASTOGI, 1983 private communication).

Jicamarca, at present, performs low resolution decoding off-line, by coherently integrating on line with a Harris/6 general purpose computer.

HARDWARE DECODERS: A COHERENT INTEGRATOR-DECODER

Coherent integration by general purpose computers usually takes most of the computer power to perform this task, leaving no CPU time to perform the decoding and statistical processing. Therefore, it is highly recommended to perform the coherent integration in a dedicated device. A device constructed to do this task can perform the decoding operation with very little added complexity.

A coherent integrator and decoder has been designed and built for the SOUSY radar (WOODMAN et al., 1980). To the authors' knowledge, it is the only device specially designed for MST applications which has been built for this purpose. The device is described in some detail in the cited reference. Here we shall limit ourselves to reproducing the block diagram (Figure 1) and discussing some of its features.

The W.K.R. decoder is a programmable device. It also performs the function of a radar controller. All the parameters which control the radar and data taking sequence are programmed in 4 computer addressable PROMS. The parameters include Transmitter IPP, baud length, code length, sequence, sampling rate and groupings. For this purpose it interprets 16 program instructions which are accompanied by a number that specifies a time delay before next instruction is executed. Maximum sampling rate is 0.5 μsec, it can process 1024 altitudes (before decoding) with single codes, or 512 in the case of code pairs like complementary codes.

THE ARECIBO PLANETARY DECODER - A CROSS-CORRELATOR

Apart of the W.K.R. decoder, the only other hardware decoder that has been used for a MST radar is the Arecibo Planetary Decoder (schematics and some text in a maintenance document at the Arecibo Observatory). The device was built for planetary radar, a more demanding application. It is capable of performing 1000 additions or subtractions per μsec and has a maximum sampling rate of 1 μsec. It has been used for decoding at the front end of any previous processing, since it has sufficient computing power to perform decoding (as much as 256 x 1 μsec bauds) without any previous coherent integration.

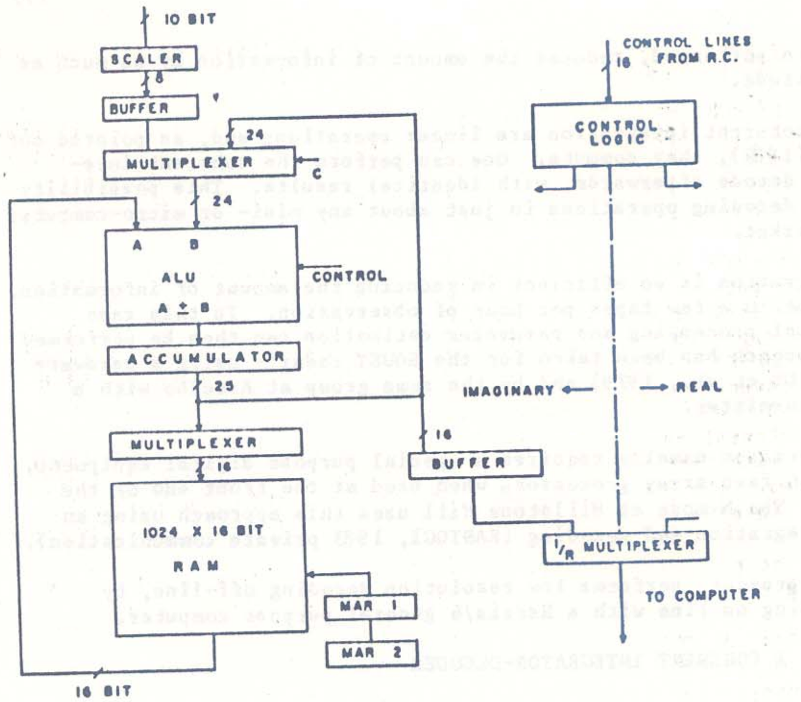

Figure 1. Block diagram for WKR Decoder and coherent integrator for SOUSY radar.

The device is more powerful than required by MST applications, but it was already available. It has been used to decode and coherently integrate (in that order) 32 baud complementary code sequences in conjunction with the 430 MHz radar (WOODMAN, 1980). Its speed has allowed the implementation of quasi-complementary sequences. These are long sequences of transmitter pulses, each coded with a different code which present advantages that have been discussed by SULZER and WOODMAN (1983) but which do not permit coherent integration before coding.

The Arecibo Planetary Decoder consists, essentially, of four parallel cross correlators. Each cross correlator consists of a selectable number 2048 maximum of lag-product integrators, depicted schematically in Figure 2. Each integrator register is associated with a given lag, and integrates the product of the present, 10 bit digitized analog signal, "multiplied" by the delayed value of a 2-bit (3 level, also called 1 1/2 bit) sample of the other signal (code in our case). The delay for the delayed signal, is produced by a 2-bit shift register. Each integrator is associated with one of the registers of the shift register and, therefore, to a given lag delay. The necessary multiplications and updating additions are performed by 64 parallel adders per cross-correlator. Speed is achieved by parallel operations and by the fact that multiplying a 3-level signal involves only additions, subtractions and no-operations (1,-1,0).

When a correlator is used as a decoder as shown in Figure 2, the length of the code is in principle unlimited, although the accumulator must eventually overflow. However, the number of ranges is limited to the number of lags. If the code length is equal to or less than the number of lags, the number of

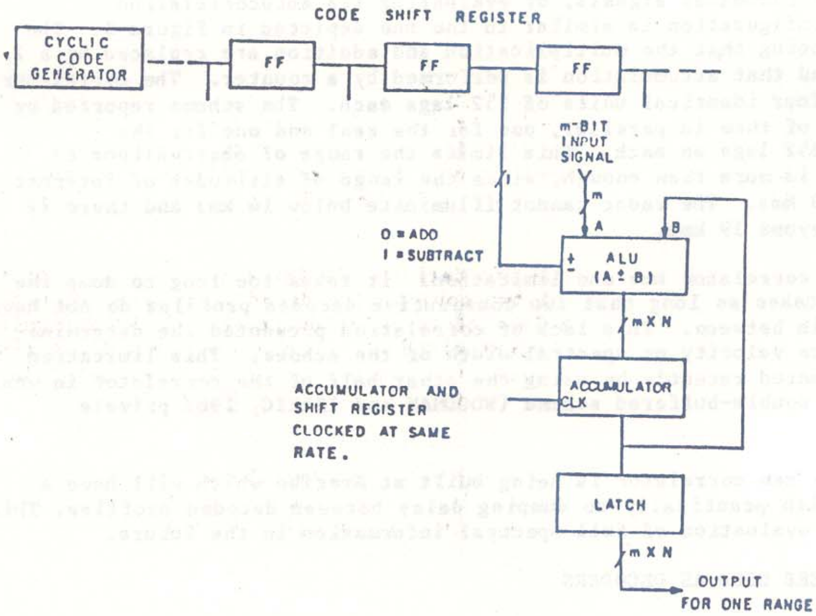

Figure 2. Decoder using correlator architecture.

ranges is also unlimited since each accumulator can be dumped as it completes the last bit of the code, and the code can be fed to the correlator again so that the lags are used for another set of ranges.

DECODING PERIODIC PSEUDORANDOM CODES

Decoding of bi-static radars, where the atmosphere is illuminated by a continuous but coded wave, requires special discussion. The length of the code in this case is much larger than in the case of a pulsed radar. It has to be at least as long as the range of altitudes one expects to receive echoes, to prevent range folding of the echoes. This can put demands on the decoding operation which are orders of magnitude higher than in pulsed MST radars.

So far, the only ST radar which works in a bi-static coded mode is the Arecibo 2380 MHz radar (WOODMAN, 1980b). In this case pseudorandom phase coded sequences with a baud length of 0.2 μsec and a period of 1024 baud where used. This corresponds to a resolution of 30 meter with a range ambiguity of ≈30 km. Decoding at all ranges would involve ≈500 operations per μsec, a formidable requirement, even when we reduce the number of ranges. Coherent integration does not help much in this case since the coherence time at 2380 MHz is nearly two orders of magnitude shorter than at 50 MHz. Nevertheless, decoding of these signals was accomplished by means of the Arecibo 1008 channel correlator (HAGEN, 1972).

Decoding becomes possible, because of the time stationarity of the radar returns, since the codes have constant amplitude and are transmitted continuously. This time stationarity permits the use of one-bit and three level (1 1/2 bit) cross correlators, which can be economically implemented by parallel integration with a large number of lags at a high sampling rate. The Arecibo correlator has 1008 lags and a maximum sampling rate of 20 MHz. This represents 20,000 operations per sec, which in this case involves simply the counting up or down of 1024 parallel counters.

The correlator at Arecibo was built for the spectral analysis of broadband (10 MHz) radio astronomical signals, by evaluating its autocorrelation function. Its configuration is similar to the one depicted in Figure 3. The main difference being that the multiplication and addition are replaced by a 2 bit multiplier and that accumulation is performed by a counter. The correlator is divided into four identical units of 252 lags each. The scheme reported by Woodman used two of them in parallel, one for the real and one for the imaginary, with 252 lags on each. This limits the range of observations to 7 1/2 kms, which is more than enough, since the range of altitudes of interest is from ≈14 to 19 kms. The radar cannot illuminate below 14 kms and there is no sensitivity beyond 19 kms.

The Arecibo correlator has one limitation: it takes too long to dump the information. It takes so long that two consecutive decoded profiles do not have any correlation in between. This lack of correlation prevented the determination of either the velocity or spectral width of the echoes. This limitation has been circumvented recently by using the other half of the correlator in what corresponds to a double-buffered scheme (WOODMAN and IERKIC, 1983 private communication).

At present a new correlator is being built at Arecibo which will have a buffered dump, with practically no dumping delay between decoded profiles. This should allow the evaluation of full spectral information in the future.

TRANSVERSAL FILTERS USED AS DECODERS

A decoder performs a convolution

$$o(t) = i(t) * h_j(t) \tag{1}$$

where $i(t)$ is the input signal
$h_j(t)$ is the impulse response of the decoder programmed for code j

and $o(t)$ is the output signal.

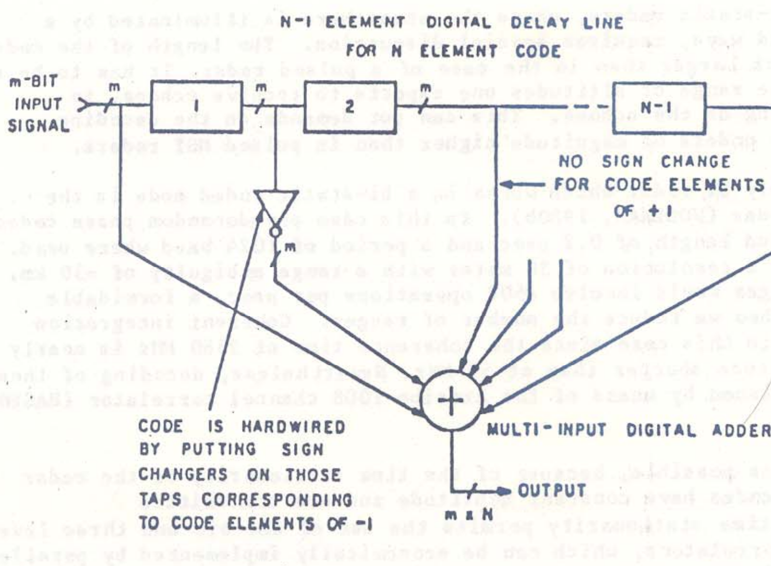

Figure 3. Transversal filter decoder for code length N.

The convolution may be written as an integral

$$o(t)=\int_{-\infty}^{t} i(q)\ h(t-q)dq \qquad (2)$$

This integral becomes a summation for discrete samples. Figure 3 shows a block diagram of a decoder for a code N long. The present value and N-1 earlier values are available in a delay line. In the case of a binary phase code the impulse response consists of +1 or -1 and so the multiplications are replaced with additions of either the signal or the inverted signal. Figure 3 shows a digital implementation. This could be done with an analog line and analog inverter and adders.

If we wish to change the code quickly, we must add a shift register and selector switches. This is shown in Figure 4. Note that if the signal enters from the left, the code must fill the shift register from the right. This accomplishes the folding about the time axis which is indicated by the negative sign in equation 2.

For a practical decoder, we need one more shift register which is parallel to the one storing the phase code. This is the amplitude of the code and is "1" wherever the code is defined and "0" otherwise. It is also attached to the control lines on the switches and when it is "0" neither switch is connected. This makes it possible to use codes with the length less than N.

An example of the type of decoder described in Figure 3 is the Barker Decoder used at Arecibo Observatory. The device is completely analog and works at the 30 MHz I.F. frequency. The input signal is fed to a surface-wave delay line with taps. The appropriate taps are inverted and added. The device is good only for the 13-bit Barker code with 4 μsec baud.

The transversal filter decoder would probably not be built with a digital implementation because of the complexity of the multi-bit delay line, and the adder. With the correlator design described in the last section, only the 3 level signal need be delayed, and one can choose either unlimited code length or range. For digital implementation, the correlator configuration is clearly the best.

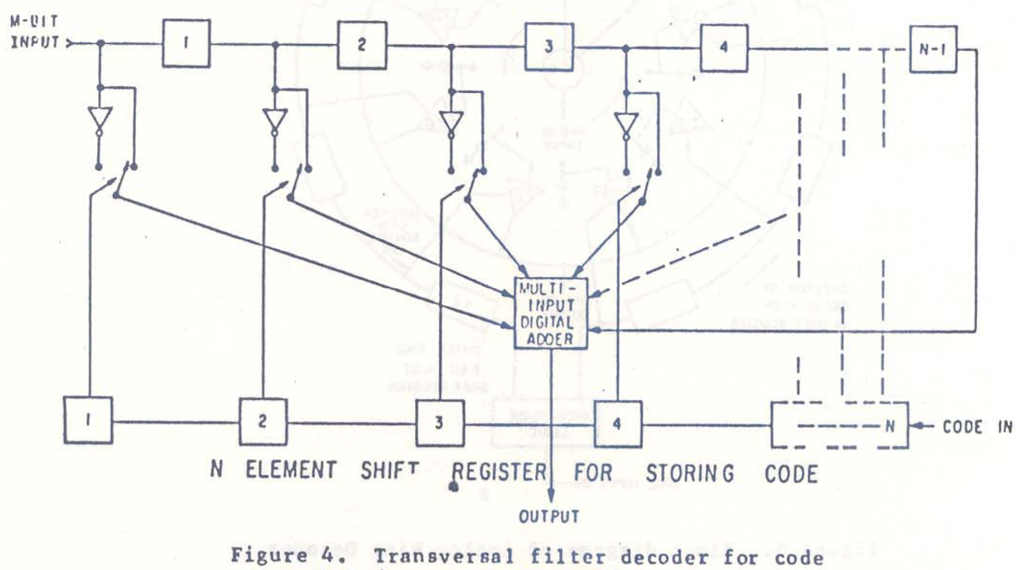

Figure 4. Transversal filter decoder for code length N (programmable code).

NEW IDEAS FOR NEW DECODERS: AN ANALOG RING DECODER

When designing a filter which will have an analog input and output, we must consider the possibility that using analog circuitry inside may be the best approach. High speed digital processors are very powerful machines, but they have some disadvantages. The most important disadvantage for our purpose is difficulty in maintenance. These machines are usually one, or a few, of a kind and the trouble-shooting procedures are often not well-defined.

We present here a new analog design which is similar to the normal transversal filter design in that the output is the simultaneous sum of N stored samples but is like the correlator design in that only the code, a simple three-level signal, need be shifted. This design has been proposed by SULZER and WOODMAN (1983, MAP Conference).

Figure 5 is a block diagram showing the concept of the analog ring decoder. To understand its operation first look at the rotor of the central rotary switch which is connected to the input signal. This switch deposits samples on the capacitors which are stored until the switch completes a revolution and deposits a new sample. As long as the code is shorter than the number of capacitors, we have enough information stored to decode the signal. All we have to do is add the signals from N adjacent capacitors with the correct polarity. To do this, we place the code in a shift register, and the outputs of the shift register control switches in the amplifiers to give the proper gain (+1, -1, or 0) to the signal. Everytime the switch moves forward one step, so does the shift register, and the correct signals are given to the delayed signals, which are added through the resistors to the output bus.

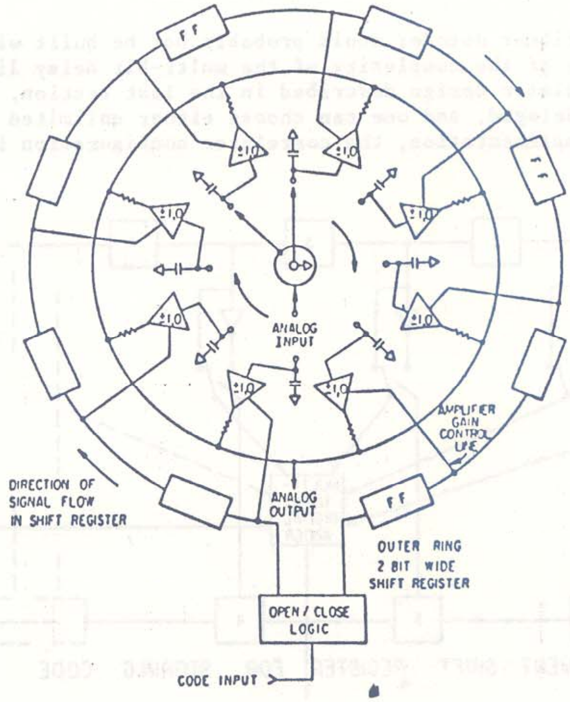

Figure 5. Block diagram of Analog Ring Decoder.

This sample block diagram depicts closely the way the actual machine operates. However, we can make a few simple additions which will allow the machine to run much faster with high accuracy. First let us make the number of capacitors and switch positions, M, somewhat larger than N, the length of the largest code we will use. Second, we arrange the phase of the switch rotor so that the code begins two positions after the switch pointer. This means that the present sample and the ones surrounding it will be connected to amplifiers with gain of 0. This allows the amplifier time to settle. Third we add a second rotor to the switch which runs one step ahead of the other. This rotor is connected to ground, and discharges each capacitor before it samples.

CONCLUSIONS AND RECOMMENDATIONS

MST radars should include hardware coherent integrators. This reduces the decoding efforts by many orders of magnitude. Once the decision to include a hardware coherent integrator has been made, a decoding operation can be included with little additional money effort. If coherent integration and decoding is performed with dedicated devices, the existing computing capacity can be used for statistical computations and parameter estimation.

Long sequences of codes, like the quasi-complementary sequences, can not be implemented with simple coherent integrators and decoders. Straight decoders are required, but devices simpler than the Arecibo decoder would be economical.

Bi-static CW radars should use continuous periodic pseudorandom codes. One bit correlators can perform the decoding operation economically.

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