PSEUDO-RANDOM CODE SIDELOBE CANCELLER

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Summary. - During acquisition of direct sequence pseudo-noise signals, time sidelobes are produced at the correlator output which will degrade detection performance. These sidelobes may be the result of additive noise, channel distortions, deliberate jamming or the non-ideal correlation function of truncated code sequences. In order to minimize these sidelobes, special codes can be selected based on their low sidelobe levels, or some special sidelobe reduction or cancellation algorithm may be devised.

A sidelobe cancellation algorithm for use with LSI correlators has been simulated. Segments of a maximum length code word as well as a totally random bit stream were tested. The simulation results show that the largest sidelobes are reduced by a small amount; however, the majority of the sidelobes are reduced by as much as 6 dB. Consequently the false alarm rate for a particular threshold setting may be reduced. A compatible technique for the derivation of a CFAR reference from the same correlator was also successfully simulated.

Introduction. - Pseudo-random code sequences have found considerable application in modern communication, radar and navigation systems. Synchronization acquisition of pseudo-random code words and sequences is often hampered by the non-ideal correlation function which manifests itself in the form of time sidelobes. These sidelobes may be the result of additive noise, channel distortions, deliberate jamming or truncation of the code sequence into isolated code words. Sidelobes can be overcome by raising the detection threshold, by selecting special codes (such as maximum length sequences or Gold codes), or by devising a sidelobe reduction or cancellation algorithm. The latter approach is more useful when noise, dispersion or jamming destroys the ideal characteristics of the specially selected codes.

In order to control the false alarm rate in the presence of fluctuating noise and sidelobe levels, a Constant False Alarm Rate (CFAR) detector may be used. The noise reference channel for the CFAR detector should consist of channel noise and code sidelobes, but be insensitive to the desired signal energy.

To develop a CFAR detector, techniques such as an offset noise IF, delayed code correlator or orthogonal code correlator can be used. An orthogonal code CFAR detection algorithm, which also provides some sidelobe cancellation, has been developed and simulated on a digital computer. The performance results for this algorithm are presented in this paper.

Pseudo-Noise Signals. - Pseudo Noise (PN) or pseudo random noise, also known as direct sequence modulation, can be generated by multiplying the carrier signal shown in Figure 1 by a bipolar pseudo random sequence. The resulting PN-PSK signal bandwidth is spread to the reciprocal of the width of the narrowest keying pulse. Alternatively the pseudo random sequence can be used to ON-OFF key the carrier for spectrum spreading. The receiver must be knowledgeable of the bandwidth spreading sequence, so that a synchronization detection or correlation process can efficiently make use of all the transmitted energy.

Pseudo-random sequences are periodic binary sequences which have certain random like properties. The most important property of such sequences is the unambiguous autocorrelation function, which permits fine time measurements after signal synchronization and tracking. One class of pseudo random sequences is generated by linear feedback shift registers. If the feedback taps are properly chosen, the sequence produced by the shift register will have a length of period of $N=2^n-1$, where n is the number of shift register stages. Such maximal length sequences (MLS) will always exhibit a difference between the number of positive and negative sequence elements of one. For a normalized correlation function peak value of one, a constant side lobe level of $-1/N$ will be found for all delays in excess of one sequence element.

Code Sidelobes. - A computer generated ideal autocorrelation function for a sequence of 63 element length is shown in Figure 2. These ideal autocorrelation properties apply only in a noiseless environment. As soon as sequence element or chip detection errors are made, as a result of channel noise or jamming, large time sidelobes will be generated at the correlator output. This is illustrated in Figure 3, where the autocorrelation function for the 63 element length sequence is plotted, assuming alternate sequence element detection errors. These detection errors not only result in increased sidelobe levels but also directly reduce the amplitude of the in-phase correlation peak.

A similar sidelobe pattern would result if correlation were performed over a single code word preceded by noise, even though the word is equal to the maximal length reference code word. A different input sequence of code words, or channel noise without input signal, could produce similar results. In each of these cases sidelobes result, since the input sequence for all time is not a repeated version of the reference as shown in Figure 4. Each of these cases typify conditions at the receiver during various phases of synchronization acquisition.

For a fixed detection threshold, time sidelobes result in false alarms whenever the threshold is exceeded. Falso alarms can increase the signal search time, or even result in missing the true correlation peak. Raising the threshold above the largest sidelobe level to reduce the false alarm rate degrades the probability of detecting the desired correlation peak under low signal-to-noise conditions. This effectively reduces the realizable processing gain at the receiver.

Code Sidelobe Reduction. - Three solutions to the code sidelobe problem include:

1. Raise the detection threshold and take the performance loss. This is obviously unattractive.

2. Use special codes or sequences which exhibit low auto and cross-correlation properties. These include Gold codes, or specific maximum length codes or sequences.

3. Use some form of sidelobe cancellation processing at the receiver which permits lowering of the detection threshold by reducing the correlator's response to code sidelobes. It is desired to attenuate the code sidelobes sufficiently so the detection threshold is only determined by the Gaussian noise floor. In some applications it may be feasible to use a canceller in conjunction with specially selected codes/or sequences for improved performance.

It is well known that maximum length binary sequences have desirable autocorrelation functions. For single maximum-length binary code words the time sidelobes are higher than for maximum length binary sequences (where the input is a repeated version of the same code word). The time sidelobe amplitude also depends upon the starting point of the code word. Each maximum length code word of length 2^n-1 has 2^n-1 starting positions, with some of these positions exhibiting lower sidelobe levels than others. A summary of desirable code words and their starting positions is given by Taylor[1]. Further information on longer code words can be obtained from Roth,[2] Braasch and Erteza,[3] or from tables of irreducible polynomials.

Figure 5 shows the maximum and minimum values of binary phase coded residues (or time sidelobes) for codes of length 2^n-1. By selecting a desirable code, a 4.5 dB to 4.8 dB reduction in peak sidelobe level is possible over a worst case randomly selected code word of the same length. Figure 5, therefore, illustrates the potential improvement in sidelobe level achievable through code selection.

Several years ago, it was observed by Gold[4] that the <u>cross</u>-correlation function of two maximal length sequences was generally poor despite their good <u>auto</u>-correlation characteristics. Gold derived a theorem for selecting sets of maximal length linear sequences which exhibit reasonably good cross-correlation characteristics. This allows several users to share the same channel through the use of "good" multiple access code sequences. The upper bound on cross-correlation sidelobes with certain maximal length Gold code sequence pairs is

$$\left|\theta(\tau)\right| \leqslant 2^{\frac{n+1}{2}} + 1$$

Gold also constructed non-maximal length sequences with good cross-correlation properties. The upper bound on sidelobes of these non-maximal length sequence is:

$$\left|\theta(\tau)\right| \leqslant 2^{\frac{n+2}{2}} + 1$$

Randomly selected code sequences would have $|\theta(\tau)| = 2(2^n - 1)^{1/2}$ as a 2σ correlation sidelobe. About 4.5% of the sidelobes would thus be higher than this value. The cross-correlation performance of maximal length Gold code sequences and random code sequences is compared in Table 1. For the Gold codes the upper bound is shown, while for the random codes 2σ values are shown.

TABLE 1. Sidelobe Levels of Codes (Cross-correlation)

Code Length	Correlation Peak	Gold Code $\lvert\theta(\tau)\rvert$ max.	Random Code $\lvert\theta(\tau)\rvert$ 2σ
2^7 -1	127	17	23
2^{13} -1	8191	129	181

Gold codes have an advantage over randomly selected code sequences provided:

integration takes place exactly over the code sequence's period,
Interfering signals are Gold codes of equal length,
Noise or jamming do not seriously degrade the code's correlation function,
Security devices are not used for code encryption, since these would destroy the properties of the Gold codes.

It is not always possible to employ Gold codes for antijam communications. Therefore, the primary application of Gold codes is in a Code Division Multiple Access (CDMA) system.

Pseudo Noise Signal Correlation. - Correlators for pseudo-random signals can be separated into two classes. Serial correlators, also known as real time multipliers (RTM), perform a code chip by code chip comparison over some period of time. The correlator is implemented with a mixer or exclusive OR (Modulo-2 adder) and a filter to perform the integration. To obtain a maximum correlation output the signal input and the reference code input must be synchronized. Synchronization is achieved by slightly offsetting the reference code rate from the input signal rate. The obvious disadvantage is that synchronization may take a long time for a relatively large time uncertainty.

Parallel correlators, also known as matched filters, store the entire reference code word for one integration period in a digital register. This reference code word is then compared against an equivalent length signal sample. The signal may be stored in another register made up of digital or analog storage elements. In an all digital implementation half adders are connected between corresponding register stages and their outputs are summed. Figure 6. illustrates the basic configuration of such a matched filter correlator. The RCA TCS-040 CMOS/SOS correlator expands on this basic configuration in that it has a 32-stage reference register and two 32-stage data registers. A set of Modulo-2 adders are connected between each data register and the reference register to provide two current sum outputs, as illustrated in Figure 7. This type of correlator has proven to be very useful in simultaneously correlating inphase and quadrature signals following coherent demodulators. Similarly, it can be operated in a two phase clock mode to achieve higher correlation speeds.

Waveforms obtained with a matched filter correlator and a 63 chip long code are illustrated in Figure 8 for several input signal-to-noise ratios.

Demodulator. - When digital matched filter correlators are used for detection of pseudo randomly coded signals, the carrier phase is generally unknown. To assure efficient signal detection and data demodulation, quadrature carrier demodulators are used. An “inphase” (I) and a “quadrature” (Q) baseband channel are then obtained. A typical coherent demodulator with I and Q baseband correlators is illustrated in Figure 9. The carrier may be either biphase or quadraphase modulated. In the biphase case the VCO could be driven so that all of the desired energy is contained in the I channel. With quadraphase modulation there will always be desired signal components in the I and Q channels regardless of the VCO phase. Appropriate combining of the I and Q output signals will then be required to maximize the signal delectability.

In the coherent phase demodulator, the incoming signal from the last IF amplifier is applied to a pair of mixers. A VCO and hybrid generate inphase and quadrature references. The mixer outputs are thus the inphase and quadrature signal components. Each is applied to a bandpass filter, which is broad enough to pass all of the pseudo-noise signal but blocks low frequencies. By blocking DC and low frequency components, carrier leakage through the mixer is rejected. The output of the bandpass filters are supplied to threshold and sampling circuits, which are sampled at the chip rate, or at a multiple of the chip rate and provide a binary signal to each correlator. When the demodulator phase is exactly the same as the transmitted phase, the I correlator output will be the only active one. The Q correlator output signal, when used as an error signal, will drive the VCO to the correct phase and thus maximize the I correlator output. During acquisition, the I and Q outputs are generally combined noncoherently, since the initial phase is not known, thereby achieving the equivalent of an envelope detector.

Noncoherent Combining. - Normally the I and Q channels are combined in a root sum square (RSS) fashion to obtain the equivalent of envelope detection. The output of an envelope detector $(I^2 + Q^2)^{1/2}$ can be approximated by the larger of $|I| + 1/2\,|Q|$ or $|Q| + 1/2\,|I|$. This results in a loss in delectability of about 0.2 dB.

A further simplification is to use $|I| + |Q|$ or the larger of the two. This results in a delectability loss of less than 0.65 dB for a detection probability P_D =.5 and false alarm probability $P_{FA} = 10^{-6}$. For lower signal-to-noise ratios the loss can usually be neglected. Table 2 summarizes the losses over envelope detection from the second approximation (magnitude combining). These alternate implementations for noncoherent combining of I and Q signals are illustrated in Figure 10.

TABLE 2. SNR Penalty Relative to Envelope Detection (dB)

	Prob. of False Alarm, P_{FA}					
Prob. of Detection	10^{-1}	10^{-2}	10^{-3}	10^{-4}	10^{-6}	10^{-8}
$P_D = 0.5$	0.15	0.35	0.40	0.55	0.65	0.70
$P_D = 0.9$	0.25	0.40	0.55	0.70	0.75	0.85

In order to relate P_D and P_{FA} to signal-to-noise ratio, consider that synchronization acquisition typically requires at least a 12 dB signal-to-noise ratio, which corresponds to a $P_D = 0.9$ and $P_{FA} = 10^{-4}$. For this case the signal-to-noise loss for $|I| + |Q|$ is 0.7 dB. It can be shown that in the limit the loss for $|I| + |Q|$ is 0.9 dB for very large P_D and very small P_{FA}.

Signal Detection Criteria. - In conventional communications systems the signal detection criteria are simply based on signal-to-noise ratio. In a pseudo random noise system, the effects of code sidelobes must also be considered; they can equally produce false alarms. Furthermore, when the PN system is used to combat jamming, the noise or interference level can fluctuate drastically and cause numerous false alarms, if a fixed threshold setting is used. A commonly used solution takes advantage of a Constant False Alarm Rate (CFAR) detector. This adjusts the threshold level based on the background noise level. The noise reference level should include code sidelobes if possible, to make the CFAR more effective. From the various CFAR implementation options, three examples will be discussed: an offset noise IF, a delayed code correlator and an orthogonal code correlator.

To achieve a valid CFAR, the detection threshold can be modified in accordance with an uncorrelated noise reference signal. One form of such a noise reference channel is a noise IF, which is slightly offset in frequency from the desired correlation channel. The frequency offset is made large enough so that no signal correlation at that particular offset can occur for expected values of carrier frequency errors and doppler shifts. Another form of a noise reference channel uses a correlation channel on center frequency, but with a reference code, which is “very late” with respect to the desired signal and will, therefore, not produce a high correlation value. This channel will have outputs due to noise and code sidelobes, which are uncorrelated with respect to the desired signal, channel noise and sidelobes, provided long code sequences are used. With the transmission of short, repeated code words the “very late” correlation might include a valid code word and produce a high correlation peak.

As an extension of this approach, an orthogonal reference code is generated, which will not correlate with the desired signal, even if it were repeated. One approach to obtain orthogonal codes is to complement half of the code chips in a code word. This may be achieved simply by dividing the code word in half and complementing the chips in the last half of the code word. Figure 11 illustrates an example of the desired code word for a data or sync symbol, and the orthogonal code word, for the noise reference.

The orthogonal code CFAR implementation is illustrated in Figure 12. It consists of two cascaded correlators of a total length equivalent to one data symbol’s duration, with appropriate output combining circuits. During data demodulation the sub-correlation values for each block are added linearly; for example, the output is

$$\Sigma\,(S + N) = A + B$$

For initial sync acquisition this linear addition, in the I channel, which corresponds to coherent detection, cannot be used. First, the carrier phase is not known and second, the frequency error may not permit integration over the data symbol duration. Non-coherent detection must therefore be used, and this is ordinarily implemented with an envelope detector in each correlation channel. The I and the Q channel are therefore combined according to the RSS algorithm:

$$\sum (S + N) = \left[(A_I + B_I)^2 + (A_Q + B_Q)^2 \right]^{1/2}$$

where the subscripts identify the I and Q components.

The orthogonal noise reference channel uses the same correlator and the same reference code as for signal detection. The algorithm selected for the noise channel is:

$$\sum (N) = \left[(A_I - B_I)^2 + (A_Q - B_Q)^2 \right]^{1/2}$$

Under low noise conditions the sub-correlation peak for block A is expected to be the same as for block B. If both are positive, the difference will be zero; if both are negative, the difference will also be zero. Since the coherent subtraction is only made over a short time interval, doppler will have a negligible effect on the cancellation of the signal correlations. The sync acquisition and the noise reference channel are thus orthogonal with respect to each other as far as the desired signal is concerned.

Since a digital implementation is a desirable objective, magnitude combining is often preferred over RSS combining. It can be implemented with trivial arithmetic operations compared to the squaring and square root operations needed for RSS combining. The detection algorithm is then:

$$\sum (S+N) = \left| A_I + B_I \right| + \left| A_Q + B_Q \right|$$

while the algorithm for the noise channel becomes:

$$\sum (N) = \left| A_I - B_I \right| + \left| A_Q - B_Q \right|$$

The magnitude combining implementation is illustrated in Figure 13.

Sidelobe Cancellation. - When noise or sidelobes are predominant during either block A or block B, the magnitude of the sum and the magnitude of the difference would be nearly alike. In that case, the noise and sidelobe output in the sync detection channel and in the noise channel would be similar or alike. Subtraction of the noise signal from the sync acquisition signal will thus cancel part of the noise and sidelobes, but it will not significantly reduce the magnitude of the desired correlation peak. A new sync signal is derived as follows:

$$\text{New Sync} = \sum(S+N) - \sum(N) = \left|A_I + B_I\right| + \left|A_Q + B_Q\right| - \left|A_I - B_I\right| - \left|A_Q - B_Q\right|$$

If a synchronization preample or complementary data are sent, the expected output will be a positive pulse. Any negative pulses can be ignored, since they represent noise or code sidelobes. For coherent decoding or complementary data modulation, the correlators must be summed linearly to extract the data signal.

Figure 14 illustrates the combined CFAR and sidelobe cancelling algorithms implemented with the magnitude combining approach. It is not expected that this represents the optimum algorithms to achieve the intended objective. However, the premise was that the same correlation channel and the same reference code should be used for both signal and noise measurements. The noise and the code sidelobes in the signal channel and the noise channel should be correlated.

One should then be able to subtract from the signal correlation channel the noise and sidelobes, which also appear in the noise reference channel. As the noise and sidelobes in the two channels become more correlated the cancellation of the undesired components should improve. The challenge is then to find a set of algorithms where the desired signal is completely orthogonal to the noise reference, and yet have the noise and sidelobes fully correlated with each other in both channels.

Simulation Approach. - In order to quantify the potential of the sidelobe cancellation algorithm, a digital computer simulation was run. The sidelobe canceller model which was simulated is shown in Figure 15. It can be configured for magnitude or RSS combining by selecting the appropriate combiner output. This permits determination of canceller performance variation with the RSS and magnitude combining approaches shown in Figure 12 and 13.

In the interest of simplicity only the I channel was simulated, however, the codeword was segmented into 4 block length to permit the configuration of a proper RSS combining model. Each block is 32 elements long, resulting in a reference codeword of 128 elements. At the start of the simulation the first 128 elements of a 255 element linear maximal length code sequence (8 bit code) were loaded into the correlator reference registers. This code is one of the desirable codes discussed previously. The simulated LSRG, which generated this code, is shown in Figure 16. The reference was then correlated against the same 255 element code sequence, which was used as the correlator input sequence. Correlation was performed on a discrete element by element basis, from $\tau = 1$ to $\tau = + 127$. The code phasing for each of the extreme τ values is shown in Figure 17. For each of the 127 values of τ the output of the sync symbol subtractor (point A in Figure 15) was recorded, resulting in 127 sample values. It should be noted that correlation over only half the 255 maximum length sequence was attempted. A codeword representing a segment of a MLS will produce a substantial number of sidelobes.

This process was repeated using a purely random 255 element code derived from the Rand table of random numbers. A + 1, -1 random code sequence was generated depending upon whether the random number was odd or even. This simulated the canceller performance with a noise or truly random sequence input. For each of these two input codes the normalized (relative to the correlation peak at $\tau = 0$)* sample mean $\sum_{i=1}^{127} \frac{X_i}{127}$ and

standard deviation $\sqrt{\frac{\sum_{i=1}^{127} \left(X_i - \bar{X}\right)^2}{127}}$ were computed

for the code sidelobes at various points in the sidelobe canceller.

Simulation Results

The simulation's sidelobe statistics are summarized in Table 3. These long term statistics show the magnitude of the sidelobes to be nearly the same in the noise and sync channels as postulated. This similarity demonstrates a potential for developing a CFAR reference from the noise channel.

These results indicate considerable reduction in mean sidelobe amplitude with the canceller (SYNC-NOISE) using both RSS and magnitude combining. This reduction applies relative to the sync channel which would normally be used during acquisition. However, the cancellation algorithm increases the sidelobe amplitude's standard deviation. It is thus possible that little reduction will be provided to the peak sidelobe amplitudes represented by the tails of the sidelobe amplitude distribution. To accurately determine the canceller's effect on the tails of the distribution, the simulation data was reviewed in greater detail. Using a printout of the 127 normalized sidelobe samples for each run, the largest, next largest . . . Nth largest sidelobe were determined. The observation time is one-half period of the 255 element code.

The 20 largest normalized sidelobe amplitudes are shown in Table 4 and relative sidelobe levels in Table 5. The results show that for this code peak sidelobe levels can be reduced by 0.67 dB with magnitude combining to 0.84 dB with RSS combining, with the cancellation algorithm.

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* Sidelobe normalization factor is 90.5 for RSS combining, 128 for magnitude combining

TABLE 3. CANCELLER OUTPUT STATISTICS

	Output	Mean η		σ	
		Noise	PN	Noise	PN
RSS Combining	Sync $\sqrt{(Y1)^2 + (Y2)^2}$	0.108	0.102	0.049	0.041
	Noise $\sqrt{(X1)^2 + (X2)^2}$	0.110	0.111	0.057	0.050
	Sync-Noise	-0.002	-0.008	0.075	0.072
Magnitude Combining	Sync $\lvert Y1\rvert + \lvert Y2\rvert$	0.097	0.092	0.046	0.038
	Noise $\lvert X1\rvert + \lvert X2\rvert$	0.098	0.100	0.053	0.050
	Sync-Noise	-0.001	-0.009	0.069	0.066

TABLE 4. Normalized Sidelobe Amplitudes

Sidelobe	MAG.	RSS	RSS With Cancel	MAG. With Cancel
Largest	0.203	0.218	0.198	0.188
2nd	0.188	0.203	0.168	0.156
3rd	0.188	0.198	0.156	0.156
4th	0.172	0.198	0.156	0.125
5th	0.172	0.178	0.155	0.125
6th	0.172	0.178	0.154	0.125
7th	0.156	0.178	0.139	0.125
8th	0.156	0.178	0.125	0.125
9th	0.156	0.178	0.0972	0.0938
10th	0.156	0.178	0.0873	0.0938
11th	0.156	0.168	0.0873	0.0938
12th	0.141	0.159	0.0862	0.0938
13th	0.141	0.159	0.0862	0.0938
14th	0.141	0.159	0.0818	0.0625
15th	0.141	0.156	0.0696	0.0625
16th	0.141	0.156	0.0685	0.0625
17th	0.141	0.156	0.0685	0.0625
18th	0.141	0.156	0.0674	0.0625
19th	0.125	0.155	0.0641	0.0625
20th	0.125	0.148	0.0586	0.0625

TABLE 5. Relative Sidelobe Level (dB)

Sidelobe	Magnitude ÷ RSS	Mag + Cancel ÷ RSS	Mag + Cancel ÷ Magnitude	RSS + Cancel ÷ RSS	Mag + Cancel ÷ RSS
Largest	-0.619	-1.29	-0.607	-0.84	-0.45
2nd	-0.667	-2.29	-1.62	-1.64	-0.65
3rd	-0.450	-2.07	-1.62	-2.07	0
4th	-1.22	-4.00	-2.77	-2.07	-1.93
5th	-0.298	-3.07	-2.77	-1.20	-1.87
6th	-0.298	-3.07	-2.77	-1.26	-1.81
7th	-1.15	-3.07	-1.92	-2.14	-.93
8th	-1.15	-3.07	-1.92	-3.07	0
9th	-1.15	-5.56	-4.42	-5.25	-.31
10th	-1.15	-5.56	-4.42	-6.19	+.63
11th	-0.644	-5.06	-4.42	-5.80	+.74
12th	-1.04	-4.58	-3.54	-5.32	+.74
13th	-1.04	-4.58	-3.54	-5.32	+.74
14th	-1.04	-8.11	-7.07	-5.77	-2.34
15th	-0.878	-8.11	-7.07	-7.01	-1.1
16th	-0.878	-8.11	-7.07	-7.15	-0.96
17th	-0.878	-8.11	-7.07	-7.15	-0.96
18th	-0.878	-8.11	-7.07	-7.29	-0.82
19th	-1.87	-7.89	-6.02	-7.67	+0.22
20th	-1.47	-7.49	-6.02	-8.05	+0.56

Lower order sidelobes are reduced by 7-8 dB. These results indicate the cancellation performance depends upon the acceptable false alarm rate (set by detection threshold) for the application.

The simulation showed that magnitude combining generally produced smaller sidelobe levels than IRSS combining in the sync channel and at the canceller output.

Conclusions. - The foremost problem in the acquisition of pseudo-noise signals is to achieve a high probability of detection and a low probability of false alarm. While noise is often the source of false alarms, its statistics are well known. Code sidelobes are another source of false alarms, however, with much less known and more varied statistical characteristics. The feasibility of reducing code sidelobes appeared to be a worthwhile simulation project. Furthermore the desire for a low cost constant false alarm rate detector (CFAR) and inexpensive quadrature signal combining techniques led to the inclusion of additional options in the simulation program.

The simulation has demonstrated that for the algorithms and codes described, an orthogonal noise reference channel can be produced which is suitable for a CFAR reference. This CFAR reference can also be used to achieve a reduction of code sidelobes at the output of a matched filter correlator. Magnitude combining appeared to result in lower sidelobes than RSS combining, which is equivalent to envelope detection. Previous analyses have shown that RSS combining in the presence of Gaussian noise is slightly superior to magnitude combining. The simulation results seem to show that for the code sidelobe statistics magnitude combining is to be preferred.

It should be noted that only one algorithm for the sidelobe canceller has been tested. Further work is required to determine the performance of this cancellation algorithm with other pseudo-random sequences and codewords. Other algorithms should also be investigated, including the differencing of non-adjacent code blocks. The limited improvement shown by the first attempt promises greater performance enhancement when the “optimum” algorithm is found.

References

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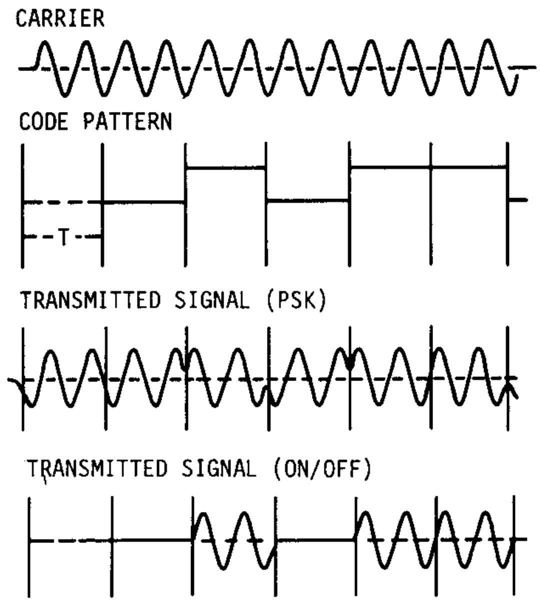

Figure 1. Pseudo-Random Keyed Waveform

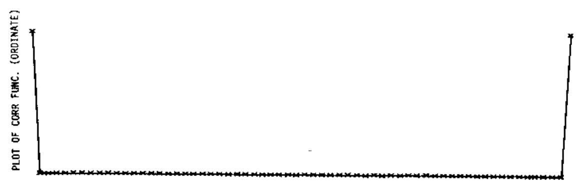

Figure 2. Correlation Function of Pseudo-Random Sequence (63 Chip MM)

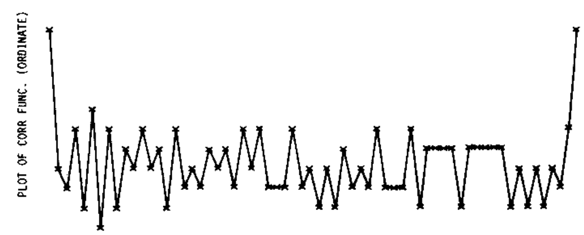

Figure 3. Correlation Function of 63 Chip Pseudo-Random Sequence; with Alternate Chip Detection Errors

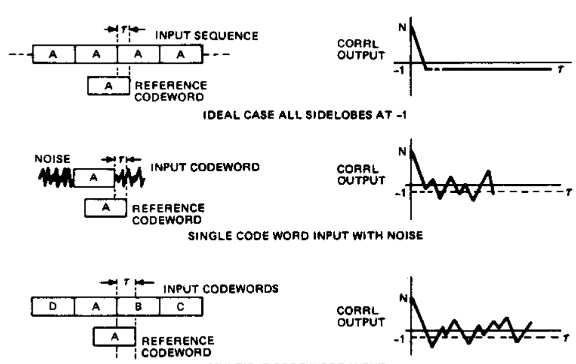

Figure 4. Code Sidelobe Generation

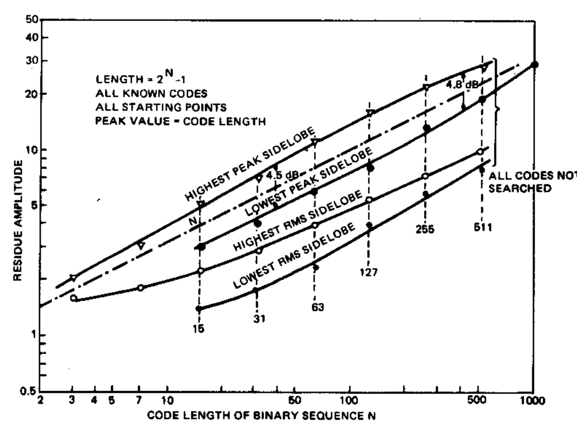

Figure 5. Maximum and Minimum Values of Binary Phase-Coded Residues[5]

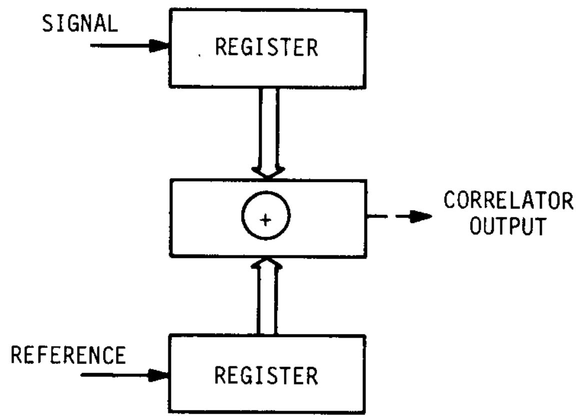

Figure 6. Basic Correlator

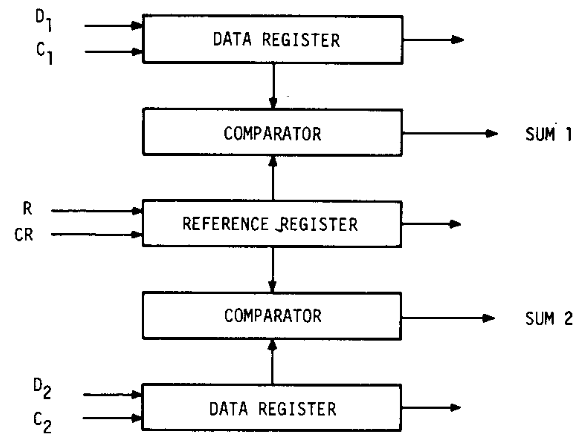

Figure 7. TCS-040 Dual 32-Stage Correlator

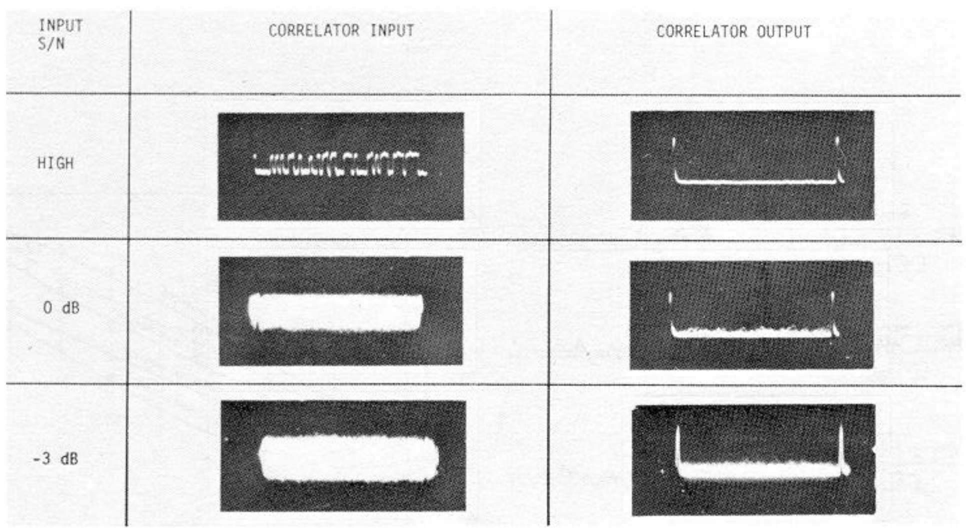

Figure 8. Matched Filter Correlator Inputs/Outputs

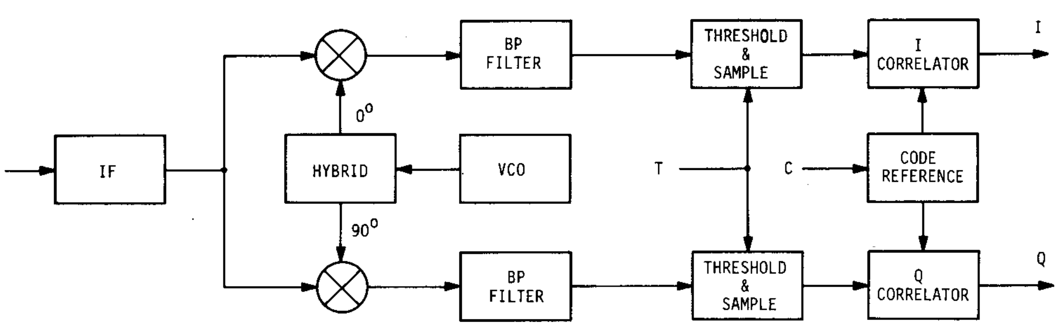

Figure 9. Coherent Demodulator

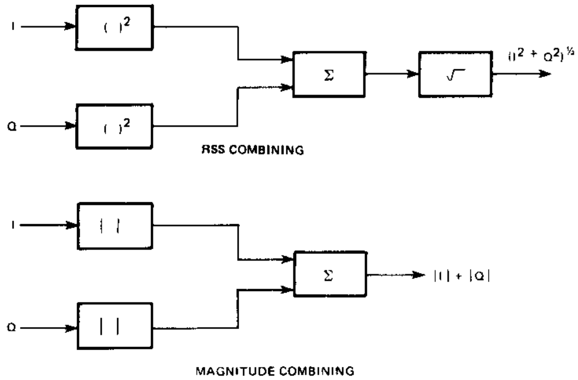

Figure 10. Noncoherent Combining

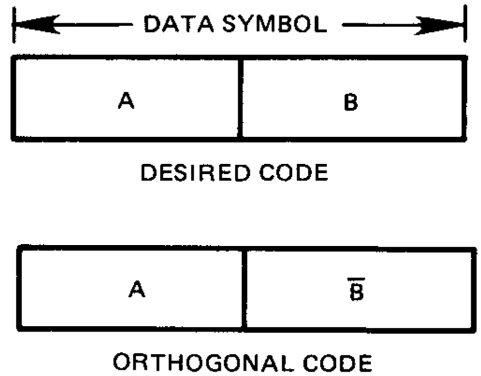

Figure 11. Orthogonal Code

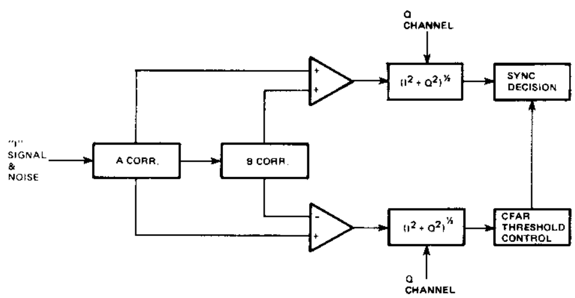

Figure 12. Orthogonal CFAR implementation

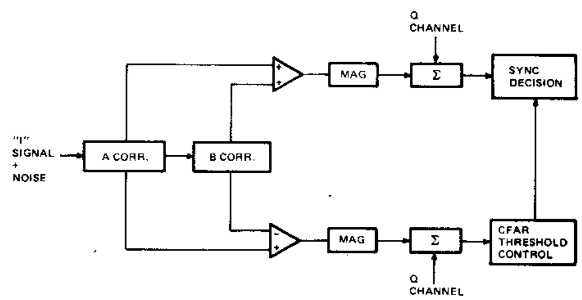

Figure 13. Magnitude Combined CFAR

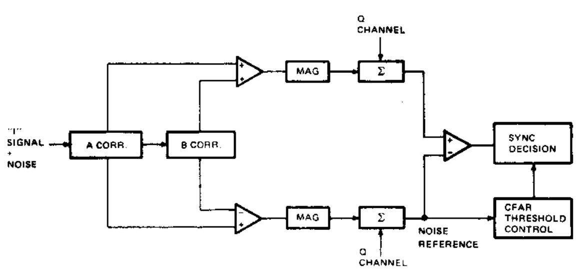

Figure 14. CFAR and Side Lobe Canceller Implementation

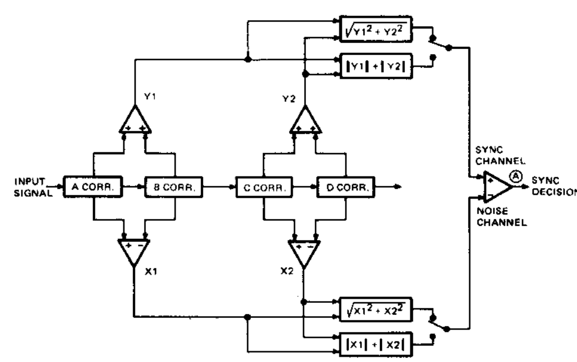

Figure 15. Sidelobe Canceller Model

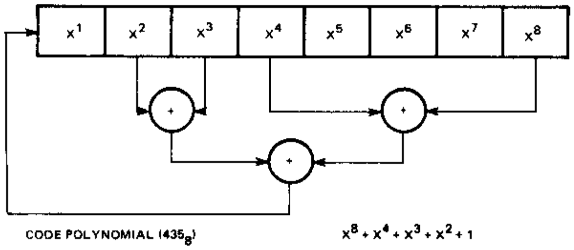

Figure 16. 8 Stage 255 Element LSRG

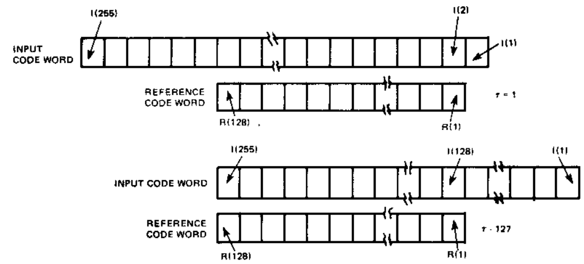

Figure 17. Simulated Code Element Correlation