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A SEGMENTED MATCHED FILTER FOR PN CODE ACQUISITION

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ABSTRACT

This paper presents a segmented matched filter (SMF) for PN code acquisition in direct sequence spread spectrum systems. While conventional matched filters provide fast acquisition in the presence of high co-user interference, they are unable to handle the problem of Doppler shift. This problem is alleviated by filter segmentation with non-coherent summation. The paper focused on validating the segmented matched filter codephase synchronization method. Simulations were carried out to verify the functionality of the SMF from a statistical point of view. The simulated performance of the SMF was compared with that of the traditional synchronization techniques and was found to show an improvement over the traditional methods when Doppler shift was present.

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1- INTRODUCTION

In spread spectrum systems, synchronization of the PN codes that are used in signal spreading at the transmitter and signal dispreading at the receiver is necessary before successful data recovery can occur. This synchronization process is split into two levels, coarse synchronization (acquisition), which brings the two PN codephases into approximate alignment, and fine synchronization (tracking), which brings the codephases into more precise alignment. In this paper we focus on the acquisition methods.

There are several existing techniques for achieving code acquisition. The most widely used one is called serial search method [1]. This is in essence a 'trial and error' method that does not retain any past signal history from trial to trial. In a serial search strategy a locally generated replica of the PN code is used to despread the received signal. Detection is achieved if the replica codephase is in a good alignment with the codephase of the PN code that was used to spread the signal at the transmitter. If it is determined that the codephases are aligned, a verification process is entered. If it is determined that they are not aligned, the codephase of the locally generated replica is advanced and the process is repeated. The serial search method can take many trials before achieving acquisition, resulting in a long setup time before the communication link is usable. Long acquisition time is undesirable in satellite based personal communications system (PCS).

Another acquisition technique known as sequential estimation method [2], which uses a portion of the incoming code signal to estimate the PN codephase and hence can achieve acquisition very quickly. However, sequential estimation is extremely sensitive to noise, interference and cannot operate in systems that use code division multiple access technique (CDMA) where there is a large amount of interference from the co-users in these systems.

Matched filter acquisition systems retain past signal history from trial to trial and can therefore achieve fast acquisition [3]. The transversal matched filter (TMF) is shown in Fig. 1. The TMF, while performing well with low signal-to-noise ratios (SNR), exhibits a degraded performance if it is operated non-coherently. In a non-coherent receiver the demodulating sinusoid is not phase locked to the transmitter carrier sinusoid. For example, a phase difference of 90^0 between the modulating and demodulating carriers has a devastating effect since no signal is detected by the demodulator. To make the matched filter function in a non-coherent system, two TMFs can be used in an in-phase and quadrature (I-Q) structure as shown in Fig. 2. However, traditional matched filtering methods are not capable of handling a problem present in satellite communications systems called Doppler shift. Doppler shift is caused by the motion of the satellite relative to the receiver, resulting in a slight difference between the transmitter modulation frequency and the receiver demodulation frequency. In this application, the term Doppler shift is used to refer to any frequency mismatch between the carrier sinusoid used at the transmitter and the sinusoid used for demodulation at the receiver.

This paper introduces an acquisition technique that is both fast and capable of handling a substantial amount of Doppler shift, making it suitable for satellite-based PCS systems using (Direct sequence spread spectrum). This acquisition method uses a modified matched filter structure called a segmented matched filter (SMF). The SMF is equivalent in operation to the serial search method, but it achieves acquisition much faster because it retains past signal history from trial to trial. The advantages of the SMF are that it performs fast with reliable acquisition in the presence of interference from co-users and it can handle a substantial amount of Doppler shift. Since this system can handle data transitions during the acquisition process, no special startup sequence is required prior to communication.

The paper is organized as follows: following this introduction, sec.2 provides the segmented matched filter structure and its principle of operation. Sec.3 explains how Doppler shift tolerance is achieved by the SMF. Sec.4 introduces the digital implementation of SMF. Sec.5 provides simulations for both TMF and SMF for PN code acquisition. Finally, sec.6 gives the conclusion.

2- SEGMENTED MATCHED FILTER STRUCTURE

The underlying principle of operation of the SMF was first introduced in a paper exploring various DS-SS acquisition structures [4]. The performances of the different acquisition structures were compared based on the average time to achieve acquisition and the tolerance to Doppler shift. In the paper, it is proposed that the Doppler tolerance of the non-coherent serial search method shown in Fig. 3 can be combined with the fast acquisition of matched filtering by using a SMF approach.

The SMF, shown in Fig. 4, is formed by dividing the typical TMF acquisition structure into segments. The SMF makes use of charge coupled devices (CCDs) to form an analog shift register to the samples of the incoming signal chips. Correlation between the PN code of interest and the incoming signal is performed on a segment-by-segment basis. The correlation outputs from the individual segments are then combined in a non-coherent (polarity insensitive) manner by squaring them prior to addition with the outputs from other segments.

The operation of the SMF when used in an I-Q structure with both sine and cosine demodulators can be related to the operation of the non-coherent serial search structure of Figure 3 as follows. In the serial search correlator, part of the integration is performed in the BPF prior to the detector; the summing within each segment of the SMF performs the same function as this pre-detection integration. The remainder of the integration in the serial search correlator is performed after the square law detector (the dwell integrator); the equivalent SMF function is the addition of the SMF segment outputs.

The effective length of the SMF is shorter than the TMF that performing correlation detection based on the same number of signal samples [4]. Since the noise performance of a matched filter structure is a function of its effective length, the SMF will have less immunity to co-user interference than the TMF. This disadvantage can be overcome by increasing the length of the SMF (at the cost of greater hardware complexity).

The great advantage that the SMF exhibits over a conventional matched filter is its improved ability to tolerate Doppler shift. How this Doppler tolerance is achieved is explained in the following section.

3- DOPPLER TOLERANCE

The SMF requires a pre-acquisition system in front of it to convert the signal coming into the antenna to a baseband signal and take samples of the result. The demodulating sinusoid used for the downconversion is not frequency locked to the transmitter carrier sinusoid because the two sinusoids are created with separate crystal oscillators that will have small differences in frequency. However, the crystals have a high degree of accuracy and the frequency difference will be very small in the absence of Doppler shift; thus the demodulation operation can be considered 'almost coherent'.

Once Doppler shift due to the satellite movement is added to the system, the frequency mismatch between the modulating and demodulating sinusoids becomes larger. The larger frequency mismatch, which causes the phase difference between the carriers to rotate at a higher rate, will result in a sinusoidal modulation of the received user signal. This amplitude modulation periodically changes the demodulated polarity relative to the transmitted polarity, as shown in Fig. 5.

This periodic reversal in demodulated signal polarity will alter the product of the signal samples and the coefficient taps in both the TMF and SMF. Consider the case where the code used to spread the incoming signal and the code represented by the PN code coefficient taps are aligned. Under normal conditions, the result of the signal sample and PN coefficient multiplication will be positive for each sample, resulting in a large positive output sum. However, when the polarity reversal in the demodulated signal as shown in Figure 5 is considered, the result of the multiplication between the reverse polarity demodulated signal samples and the PN code coefficients will be negative.

The TMF, which sums the multiplication result over the length of the entire filter, will produce a zero output because the first half of the coefficient multiplications produce positive results and the second half produce negative results. This is shown conceptually in the top rectangle in Figure 5 where the white half represents the positive outputs and the shaded half represents the negative outputs. Since the TMF will output zero even when the codephases are aligned, it will be incapable of performing acquisition in the presence of Doppler shift.

Under the same conditions, the SMF will be able to detect the alignment of the codephases. It sums the multiplication results over small segments, then squares each segment sum and adds the squared result to the outputs from the other segments. This means that the segments in the first half that contain all normal polarity signal samples will have large positive outputs. The segment in the middle which experiences a polarity transition halfway through its signal samples will have a zero sum output. The segments in the second half that contain all reverse polarity demodulated signal samples will have large negative sums which are squared to produce large positive sums. When the results from all the segments are added together, the output from the SMF will be large and it will detect the codephase alignment. This is shown conceptually in the bottom rectangle of Figure 5.

There is a limit on the amount of Doppler shift that the SMF can handle. If the Doppler shift is large enough that the amplitude modulation envelope changes the demodulated polarity in the middle of each segment, the SMF will be incapacitated. However, the SMF shows a great deal of improvement over the TMF in handling Doppler shift. The SMF can tolerate the same amount of phase rotation due to Doppler shift over one segment that the TMF can tolerate over its entire length. For example, a 16 segment SMF can handle 16 times more Doppler shift than an unsegmented TMF.

4- DIGITAL IMPLEMENTATION OF THE SMF

To simplify the implementation of the SMF, digital shift registers can be used in place of analog charge coupled devise (CCD) shift registers [5]. In the digital SMF, instead of taking in analog samples of the received signal amplitude, the SMF takes in signal samples that have been quantized. These quantized signal samples are placed in a digital shift register, called the "Signal Register", shown in Fig. 6. Instead of taps weighted with the PN code of interest, there is another shift register called the "Code Register" which is pre-loaded with the user PN code prior to operation. A 'match detection' (an exclusive-NOR gate comparison) is performed between the contents of the Code Register and the Signal Register in place of multiplication with weighted taps.

Like the results from multiplication between the signal samples and the PN code coefficient taps in the analog SMF, the results from the match detection operations in the digital SMF are added over each segment and then squared. The squared outputs from all the segments are added together to form the final output of the SMF.

The signal input to the digital SMF acquisition system is the result of demodulating and then quantizing the received signal. The pre-acquisition front end circuitry required to do so is shown in Fig. 7. The carrier-modulated signal coming into the antenna is multiplied by a demodulating sinusoid to translate the signal down to baseband. The baseband signal is then passed through a 'chip matched filter' which is an integrate-and-dump circuit. In the absence of co-user interference and channel noise, the result of integration will be positive if a '1' was sent, and negative if a '0' was sent.

The output from the 'chip matched filter' is dumped into a quantizer which decides if a '1' or '0' was sent and sends the appropriate digital level(s) to the input of the SMF. The SMF can be implemented using varying numbers of quantizing bits for converting the incoming signal chip amplitudes to digital levels. Two and four levels quantization were taken into account.

Table 1. Match Output Assignation (2 Levels Quantization)

Code	Signal	Match Value
0	0	+1.5
0	1	-1.5
1	0	-1.5
1	1	+1.5

The digital SMF match detection output assigned to the result of the comparison between the Code Register and Signal Register contents is summarized in Table 1 for the case where 2 levels quantization is used. The values of ± 1.5 are used instead of ± 1 so that the same hardware implementation can be used for 2 and 4 level quantization. This assignment is equivalent to weighting the PN code coefficients in the analog SMF with ± 1.5 instead of ± 1. Assigning ± 1.5 to matches and mismatches amplifies the output but does not change the operation in any way as both the signal and noise outputs are amplified equally.

If 4 levels quantization is used, the Signal Register of the SMF will have two bits depth. There will be two inputs, Signal 1 MSB (the most significant bit) and Signal 0 LSB (the least significant bit), that receive the 4 levels quantized signal chips as shown in Fig 8. The match detection output assignation for 4 levels quantization is outlined in Table 2.

Table 2. Match Output Assignation (4 Levels Quantization)

Code	Signal1	Signal0	Match Value	Meaning
0	0	0	+1.5	'Strong Match'
0	0	1	+0.5	'Weak Match'
0	1	0	–0.5	'Weak Mismatch'
0	1	1	–1.5	'Strong Mismatch'
1	0	0	–1.5	'Strong Mismatch'
1	0	1	–0.5	'Weak Mismatch'
1	1	0	+0.5	'Weak Match'
1	1	1	+1.5	'Strong Match'

The reason for assigning the ± 1.5 to matches and mismatches for the 2 levels quantization SMF now becomes clear. The 4 levels quantization circuit can be used 2 levels quantization by tying the two Signal Register inputs, Signal 1 and Signal 0, together and driving them with a 2 levels quantized input. This would correspond to the entries in Table 2 where the Signal 1 and Signal 0 states are the same and ± 1.5 is assigned to matches and mismatches.

5- SIMULATIONS FOR PN CODE ACQUISITION FOR TRANSVERSAL AND SEGMENTED MATCHED FILTERS

A series of MATLAB simulations were carried out in order to verify the noise immunity and Doppler tolerance performance of the SMF. The simulated SMF performance is compared against the simulated TMF performance. A 512 samples length digital SMF made up of 16 segments each containing 32 samples was simulated using both 2 and 4 levels quantization. Simulations for the TMF were run for a digital TMF with 2 levels quantization for 512 sample length filters. The 512 samples length TMF was chosen because it has the same length of registers as the 512 samples SMF and is therefore similar in hardware complexity.

There are two results desired from the simulations. The first is the pdfs of the amplitudes of the SMF and TMF response when the codephase of the user signal is correctly aligned with the contents of the Code Register. The second is the pdfs of the output amplitudes when the codephases are misaligned. These pdfs were obtained using the relative frequency statistical method where the output was simulated for a large number of occurrences and the resulting output amplitudes were used to construct the pdfs.

There are two main parameters that determine the performance of the acquisition system, probability of missed detection (P_{MD}) and probability of false alarm (P_{FA}). Missed detection occurs when the codephase is correct but the correlation output is below the threshold value. False alarm occurs when the correlation is above the threshold value but the codephase is actually incorrect. The P_{MD} is given by the area under the aligned pdf that falls to the left of the threshold. The P_{FA} is the area under the non-aligned pdf that falls to the right of the threshold. The threshold has been set at the intersection of the two carves, which is the optimal threshold when the missed detection and false alarm conditions have equal time penalties.

Since P_{MD} and P_{FA} depend on the threshold, which in turn depends on the acquisition system surrounding the SMF and TMF, the metric chosen to indicate performance is the overlap area between the aligned and non-aligned output amplitude density functions. This will indicate to what degree the aligned and non-aligned outputs can be distinguished; the larger the overlap area, the more likely the unwanted events of missed detection and false alarm are to occur.

Doppler performance simulations were carried out for a 512 samples length TMF with 1 bit (2 levels) quantization, and for a 512 samples length SMF with 1 bit (2 levels) and 2 bits (4 levels) quantization. The simulations were carried out for 3 kHz, 5 kHz, and 10 kHz Doppler with 10, 20, 50 and 80 co-users.

Plots for 3 kHz of Doppler with 10 co-users are shown in Fig. 9, 10 and 11. The term 'unusable' has been used to represent cases where the performance is so poor that the aligned and non-aligned cases are indistinguishable. It can be readily seen that the conventional TMF is rendered unusable with just 3 kHz of Doppler, while the SMF still exhibits a clear distinction between the aligned and non-aligned cases. A summary of the results for all simulated cases is contained in Table 3.

Table 3. Simulation Results for varying Doppler shift

Doppler	Co-users	512 Sample TMF 2 Level Quantization	512 Sample SMF 2 Level Quantization	512 Sample SMF 4 Level Quantization
3 kHz	10	unusable	0.3263	0.2225
3 kHz	20	unusable	0.5670	0.4880
3 kHz	50	unusable	0.8018	0.7449
3 kHz	80	unusable	0.8653	0.7869
5 kHz	10	unusable	0.3283	0.2572
5 kHz	20	unusable	0.6005	0.5054
5 kHz	50	unusable	0.8205	0.7332
5 kHz	80	unusable	0.8824	0.7967
10 kHz	10	unusable	0.4397	0.3615
10 kHz	20	unusable	0.6609	0.6079
10 kHz	50	unusable	0.8374	0.7757
10 kHz	80	unusable	0.9040	0.8282

It can be seen that while Doppler shift is the limiting factor for the TMF which is unusable with even small amounts of Doppler shift, for the SMF it is the co-user noise which degrades its performance most severely. As previously mentioned, a longer SMF will exhibit more co-user interference immunity. The SMF can be made longer to improve its performance in the presence of co-user noise without decreasing its immunity to Doppler shift. The cost for better noise performance is greater hardware complexity, but there is no tradeoff in terms of Doppler shift performance.

Finally, as a future work concerning extension of the analysis, is to find a relationship between the overlapping areas of the probability density functions and the average acquisition time of the proposed system since it is an important measuring parameter for the initial synchronization schemes.

6- CONCLUSION

Simulations were run to verify the performance of the SMF acquisition method. The simulation results showed a great improvement in the performance of the SMF over the TMF in the presence of Doppler shift. In fact, the TMF becomes unusable with even small amounts of Doppler shift while the SMF performance is degraded only slightly. The simulated SMF performance exhibited degraded acquisition performance in the presence of co-user interference when compared with the TMF acquisition method. This degraded noise performance can be overcome by increasing the length of the SMF, at the cost of greater hardware complexity.

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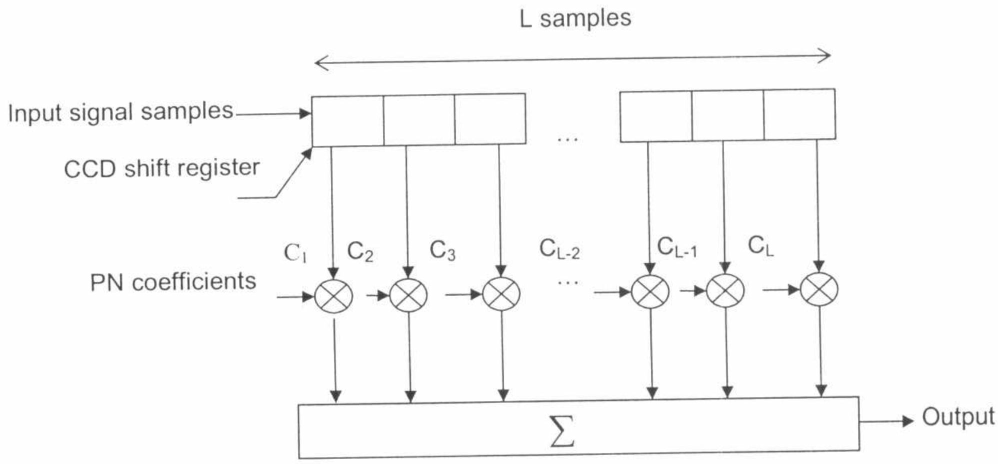

Fig. 1. Transversal Matched Filter Structure

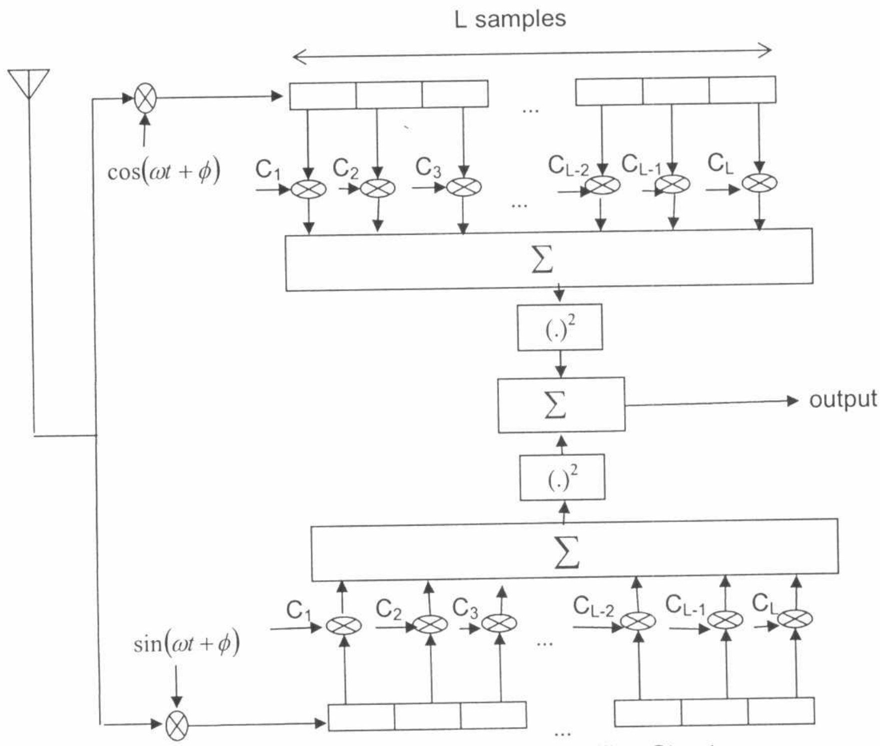

Fig. 2. I-Q Transversal Matched Filter Structure

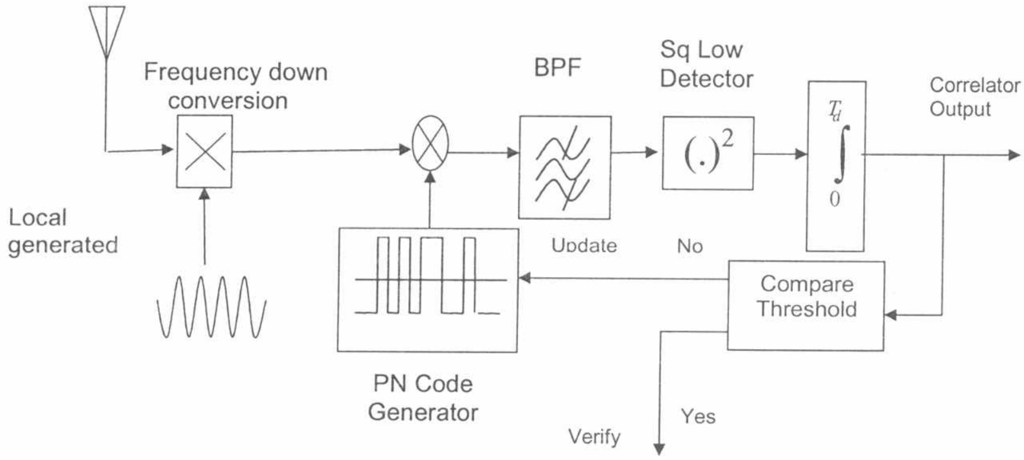

Fig. 3. Non-coherent serial search structure

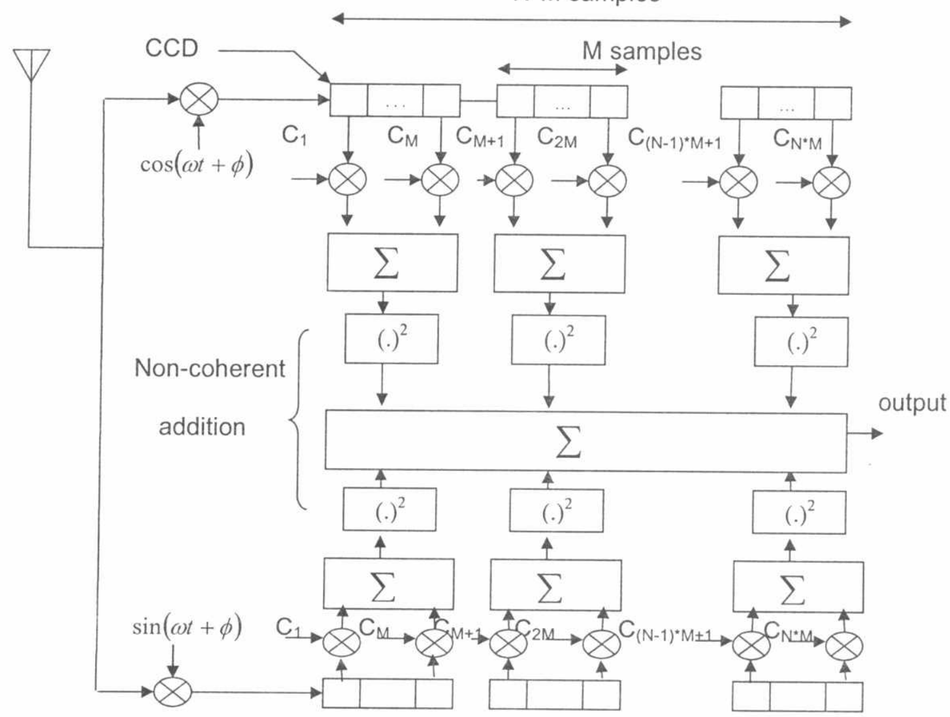

Fig. 4. Segmented Matched Filter Structure

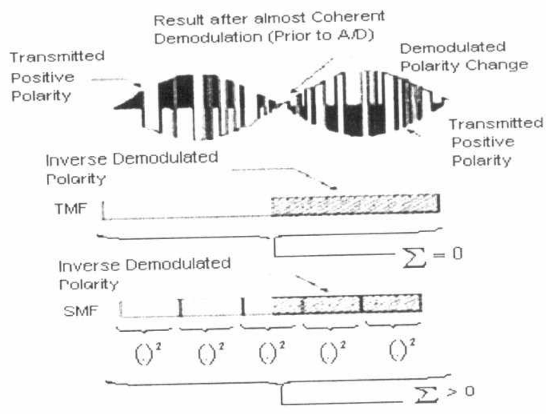

Fig.5. Effect of Doppler induced phase rotation on TMF and SMF

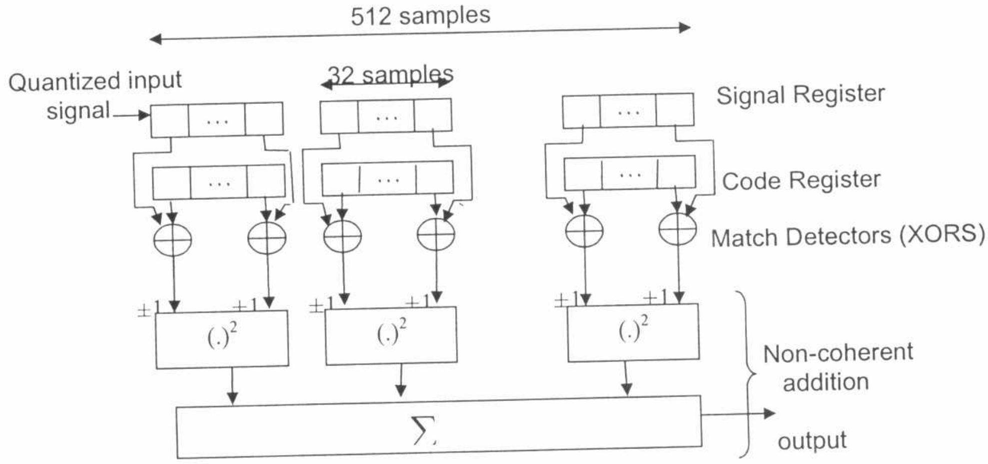

Fig.6. Segmented Matched Filter Structure with 1 bit depth signal register

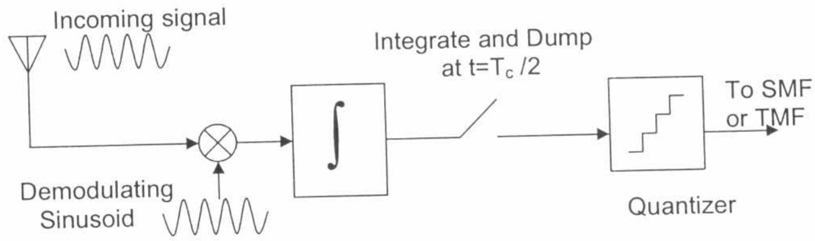

Fig.7. Pre-acquisition chip Matched Filter

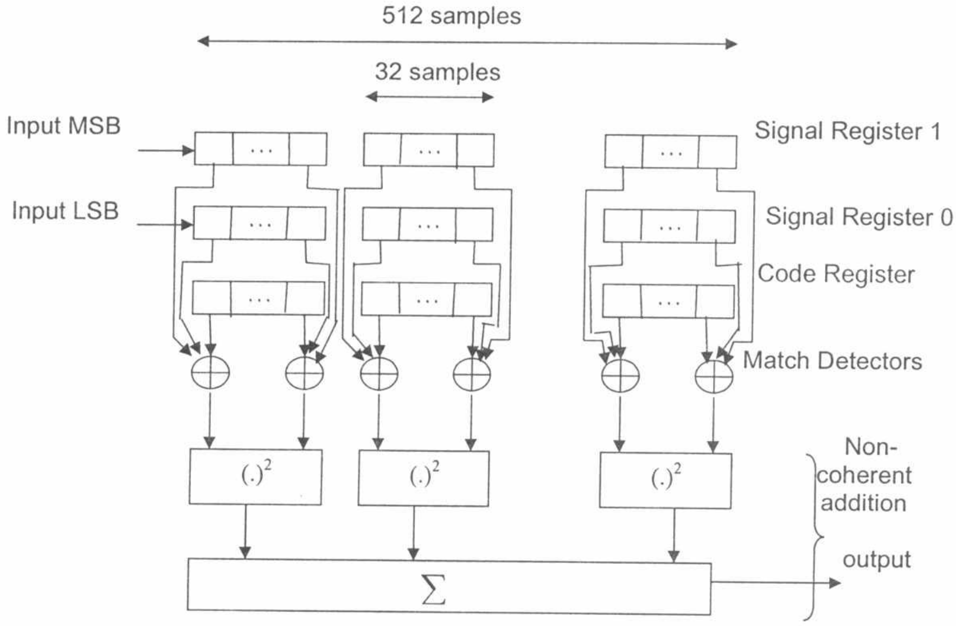

Fig.8. Segmented Matched Filter Structure with 2 bits depth signal register

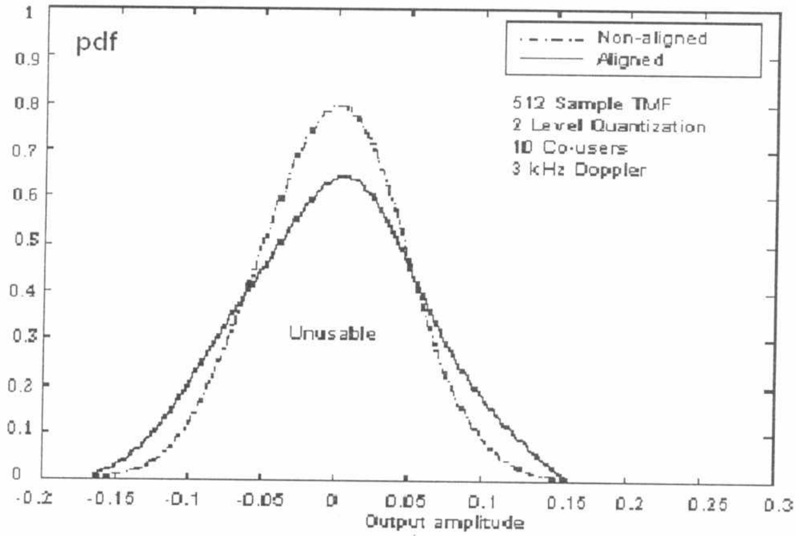

Fig.9. Probability density functions for 512 bits TMF, with 2 level quantization , 3 kHz Doppler and 10 co-users

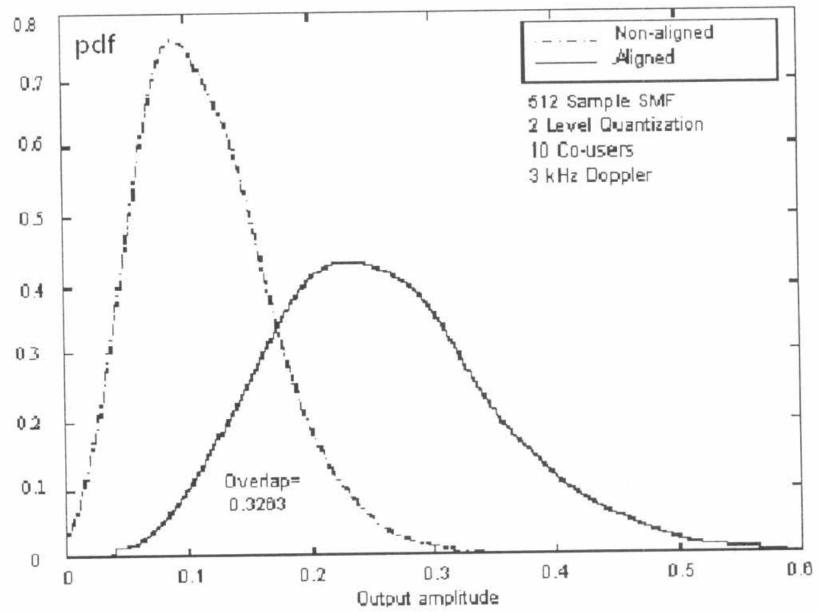

Fig.10. Probability density functions for 512 bits SMF, with 2 levels quantization, 3 kHz Doppler and 10 co-users

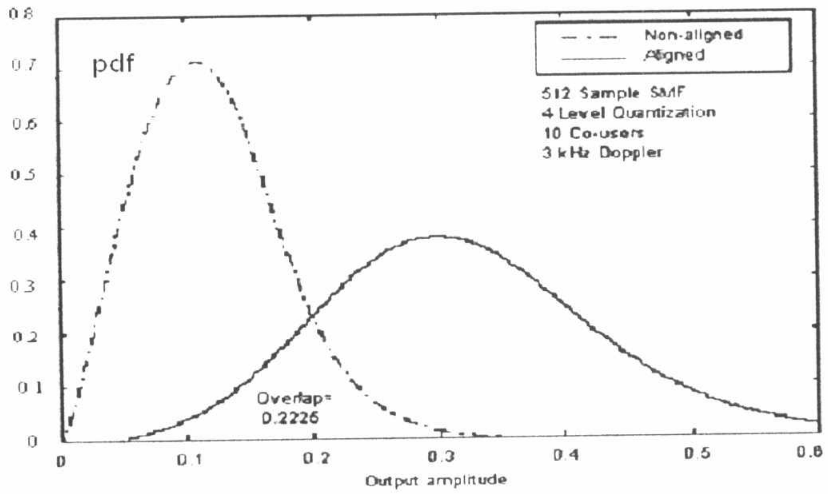

Fig.11. Probability density functions for 512 bits SMF, with 4 levels quantization , 3 kHz Doppler and 10 co-users