CDMA SYNCHRONIZATION STUDY

Final Report

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Acknowledgments

Much of the work reported in this document has been completed by Xing Zhang and Derek Thille. Graphs of acquisition time vs. matched filter length and correlation length have been computed by Ms. Zhang. The transition detection method and performance comparison with other methods have been reported in her M.Sc. theses. The almost coherent segmented MF work forms part of the M.Sc. project of Mr. Thille and his work is still in progress. He has supplied graphs of detector output versus codephase and Doppler offset frequency.

The guidance of Mr. Mike Moher in this project is gratefully acknowledged. He has provided applications information and the section on CDMA personal communication satellite includes material provided by Mr. Moher.

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Contact

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Preface

The objectives of this project are to propose a strategy for both code and frequency acquisition in a multiple access DS-SS system and to characterize the performance of the proposed strategy through analysis and/or simulation. The proposed strategy should be appropriate for personal communication satellite systems. The project requirements are summarized below.

Target system parameters are:

- BPSK modulation with a chip rate of 1 MHz
- PN sequence length on the order of 2^{16} (65,535) chips
- data symbol period is suggested to be 256 times the chip period
- co-user interference equivalent to additive white Gaussian noise
- strategy should be appropriate for fading channel applications
- no side information is available for narrowing the search range
- unknown carrier frequency offsets up to ±10 kHz
- local clock accuracy no better than 10^{-6}
- user transmissions are synchronized on forward but not reverse path

Requirements of the acquisition system are:

- time to acquire: < 5 seconds
- minimize the required E_c/N_0' for acquisition
- E_c/N_0' as high as -5dB may be acceptable
- minimize local clock accuracy requirements
- implementation size power and cost suitable for a handheld unit

Investigations to be performed:

- characterize acquisition time as a function of the SNR
- characterize sensitivity of acquisition time to PN code length
- suggest how tracking should be performed once acquired

Questions to be considered:

- is it better to synchronize to a single long sequence of 2^{16} chips or to one of several length 2^{11} sequences?
- the use of a pilot sequence without data modulation would seem to be necessary approach for the long sequence approach but is it necessary with the short sequences?
- due to the power limitation, it may not be appropriate to have a signaling/synchronization channel in all beams at all times. Could this be combined with the pilot sequence?

1. Introduction to Spread Spectrum Synchronization

The purpose of this project is to investigate different codephase synchronization methods for a CDMA receiver and to evaluate their characteristics in satellite mobile and personal cellular CDMA network applications. In a DS-SS receiver, synchronized replica of the transmitted spreading code is necessary so that the data signal can be despread and the data sequence can be recovered. The process of initial coarse acquisition is the focus of this report.

Several levels of synchronization are required in a direct sequence spread spectrum CDMA receiver. These include carrier synchronization, codephase synchronization and symbol synchronization. Carrier frequency / phase synchronization is a prerequisite for some methods of codephase synchronization, while other schemes can work with "nearly coherent" detection. These latter schemes tolerate small Doppler frequency offsets.

Principal techniques for codephase acquisition are non-coherent correlator serial search and coherent or I-Q almost-coherent correlators or matched filters.

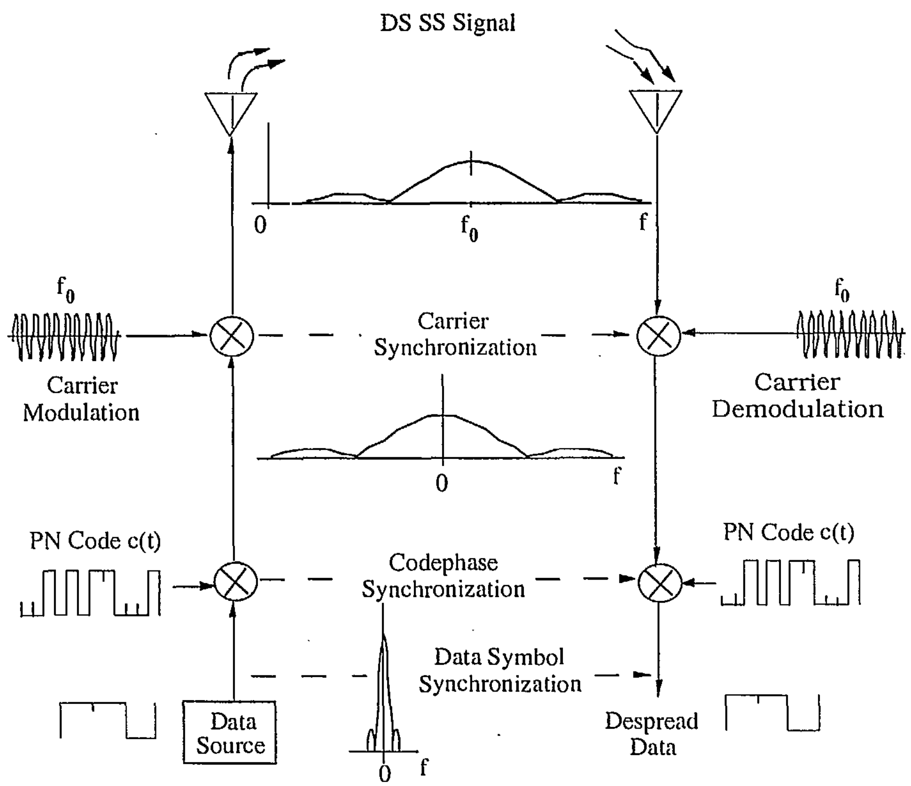

Fig. 1 Basic SS synchronization

After coarse code phase acquisition has occurred, code phase tracking takes over. This is the process of continually adjusting the locally generated reference code phase (and frequency) using a feedback loop to maximize correlation. A digital delay lock loop (DDLL) is the usual method of tracking.

Data symbol synchronization is required to detect the information content of the despread signal. A unique word is often sent at the beginning of a data sequence and this can aid in symbol synchronization. The spreading code cycle is often designed to be equal to the data symbol period thus code synchronization can also be used for symbol timing.

The basic spreading and despreading processes are illustrated in Fig. 1 for a single user system. Note the requirement for synchronization at three levels.

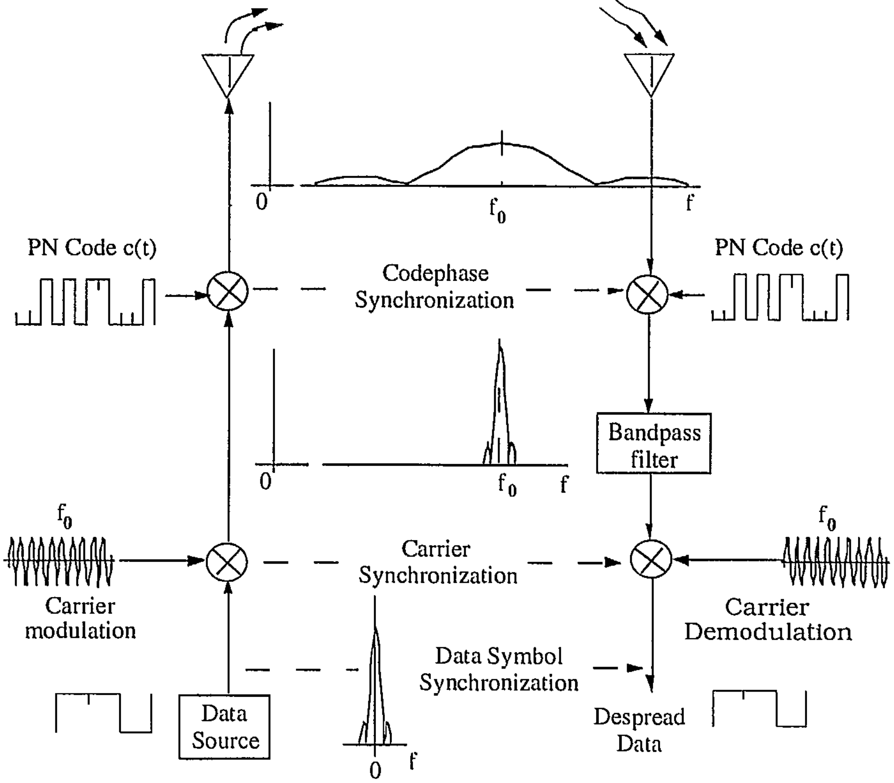

Fig. 2 Multiple access CDMA system

Fig. 2 shows similar spreading and despreading processes as in Fig. 1 except that the order of operations has been reversed. For a multiple access receiver, it is desirable to despread only the signal of interest. For the case of multiple access, the carrier phase or frequency may vary for signals from different transmitters and therefore the receiver must despread only the desired signal before performing carrier synchronization.

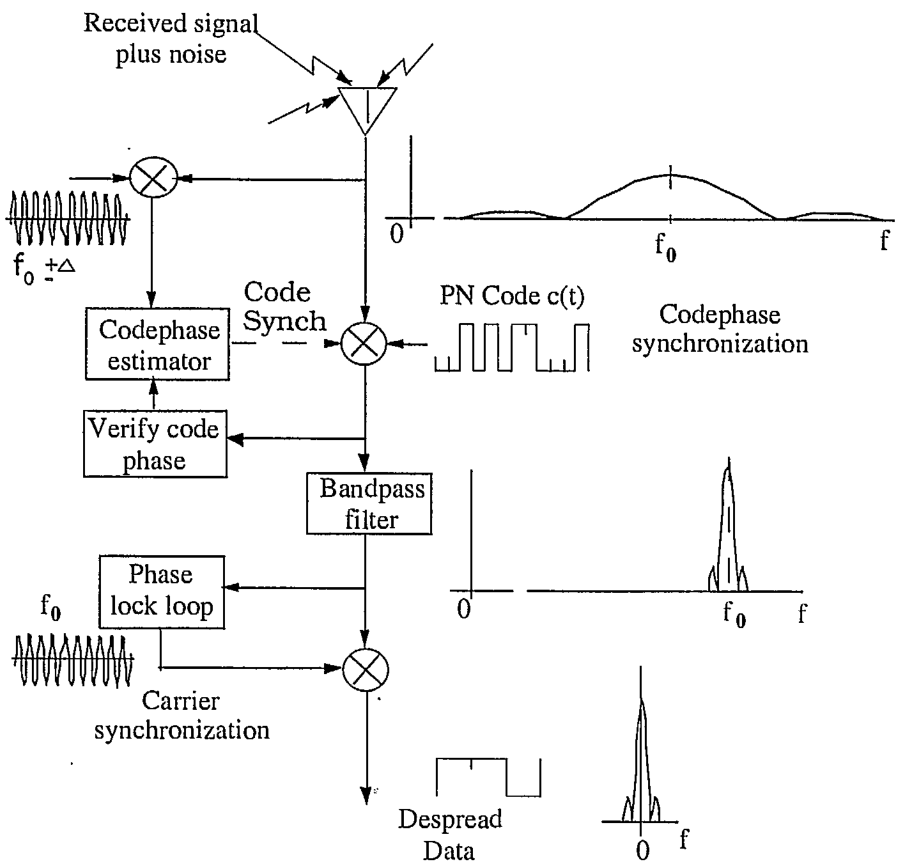

Fig. 3 Receiver with code phase estimation

The code phase estimator illustrated in Fig. 3 uses an almost-coherent local carrier and quadrature demodulation to translate the spread signal to baseband. This allows more rapid acquisition through the use of a matched filter. Since there is some frequency error in the local carrier, the baseband signal will "rotate" in the I - Q quadrature outputs. To address this rotation, several methods have been proposed in the literature. This report proposes an alternative method of detection using only the I channel and using a segmented matched filter or a transition based matched filter.

2. Coarse Acquisition Schemes

In DS-SS systems code phase synchronization includes the two steps of coarse acquisition and fine tracking. In some cases, synchronization must be accomplished quickly at very low SNR and also in the presence of a jamming signal.

The serial search scheme is relatively simple and achieves good performance in the presence of strong interference or noise. The major disadvantage of serial search is its long acquisition time which results in an excessive setup time before the communication link is usable. The serial search "trial and error" correlation method does not retain any past received signal. In contrast, matched filter schemes retain a history of the input signal and can therefore reduce the acquisition time. Unfortunately, frequency error in the demodulator complicates the use of a matched filter.

Demodulation of the desired signal carrier can be non-coherent, almost coherent or coherent depending on the accuracy of the receiver's local oscillator. Section 2 contrasts non-coherent and coherent techniques and provides background for the almost-coherent schemes discussed in Section 3.

2.1 Correlator serial search (non-coherent)

The block diagram of a non-coherent direct-sequence serial search (*SS*) system is illustrated in Fig. 4. The spread spectrum signal acquisition is performed at intermediate frequency (IF) or even at radio frequency (RF). If a cell is within ±1/2 chip alignment with the incoming code, most of the energy will be within the passband of the bandpass filter (BPF). On the other hand, if the cell is more than 1/2 chip apart from the correct phase, only a small portion of energy is passed through the filter. The BPF is followed by an envelope detector or a square law detector which is then followed by an integrate-and-dump filter. The integrator output is compared with a predetermined detection threshold to determine if the correct codephase has been reached. The integration time is also known as the dwell time, T_d. It will be shown later in Section 2.5 that an optimum dwell time is in the order of 300 chip periods.

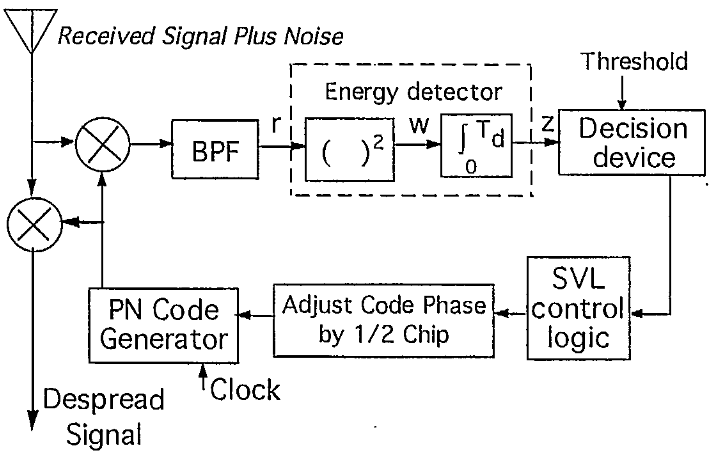

Fig. 4 The non-coherent serial search scheme

In serial search, this test is performed sequentially at all code phases of the PN spreading code (which might be a ML code or a Gold code). In search mode, the receiver PN code phase is adjusted by steps of 1/2 chip or 1/4 chip and each codephase step is called a cell. Local PN codephase alignment for a particular cell is tested against the incoming (antenna) signal by serial correlation over a dwell time, T_d. If the correlator output fails to exceed the detection threshold, the cell is rejected and the next cell (1/2 chip advance) is tested. If the detection threshold is exceeded, the process moves toward lock mode by entering a verification state. The dwell time in the verification mode is several times longer than in search mode; this increases decision accuracy and reduces false lock (probability. The transition from search mode to lock mode occurs only after the verification test is passed. Coarse acquisition is declared once entering into the lock mode. The next mode is the "fine synchronization" in which the tracking loop is activated.

In the absence of a priori information, the code phase acquisition process starts with an arbitrary code phase and steps through all phases of the entire code cycle. A uniform distribution is assumed for the expected codephase position and a "straight line" serial search strategy is used. If the codephase is roughly known through a priori information, the serial search is conducted over an uncertainty region which is a portion of the PN code cycle. The a priori distribution of the assumed code phase position is usually modeled with a triangular or truncated Gaussian probability density function. More sophisticated search strategies such as Z-search or expanding-window search are suitable to the applications where a priori information is available and where it is not feasible to search the entire code period since the PN code period is very long. The selection of the best search strategy for the receiver is dependent on the available prior information and the search test accuracy (P_d and P_{fa}).

The envelope detector or the square law detector of the non-coherent correlator operates at intermediate frequency and has a detection bandwidth which must exceed both the reciprocal of the dwell time and the user data bit rate. This bandwidth gives some tolerance to receiver carrier frequency offset or to Doppler shift, however, the detection bandwidth must be limited to reduce the noise at the detector. In systems where the Doppler offsets are large, a bank of parallel BPF filters and detectors can be used to cover the Doppler frequency range. Alternately, the receiver IF frequency can be varied to perform codephase searches in sequential Doppler "bins". This compound search method adds a second dimension and therefore increases the acquisition time in proportion to the number Doppler bins.

Serial search envelope detection is similar to on-off keying (OOK), when the received signal is despread, the desired carrier falls within the BPF filter bandwidth. When the codephase is incorrect, the IF filter bandwidth contains only a small portion the spread signal "noise". Although the non-coherent detector is versatile, it has a distinct disadvantage. When compared with coherent detection, non-coherent envelope detection results in a larger probability of false alarm. This is because noise components of a false codephase correlation always result in a positive envelope output. The bandpass filter output, |r|, shown in Fig. 4 has a Rayleigh distribution of the PDF when the codephase is incorrect and it has a Rician distribution when the codephase is correct. An example illustration envelope PDF is given in Fig. 5a where the noise component has variance $\sigma^2 = 1$ and the signal has amplitude S = 1.4. Following the bandpass filter in Fig. 4, the squaring function has output w which has the probability density function shown in Fig. 5b.

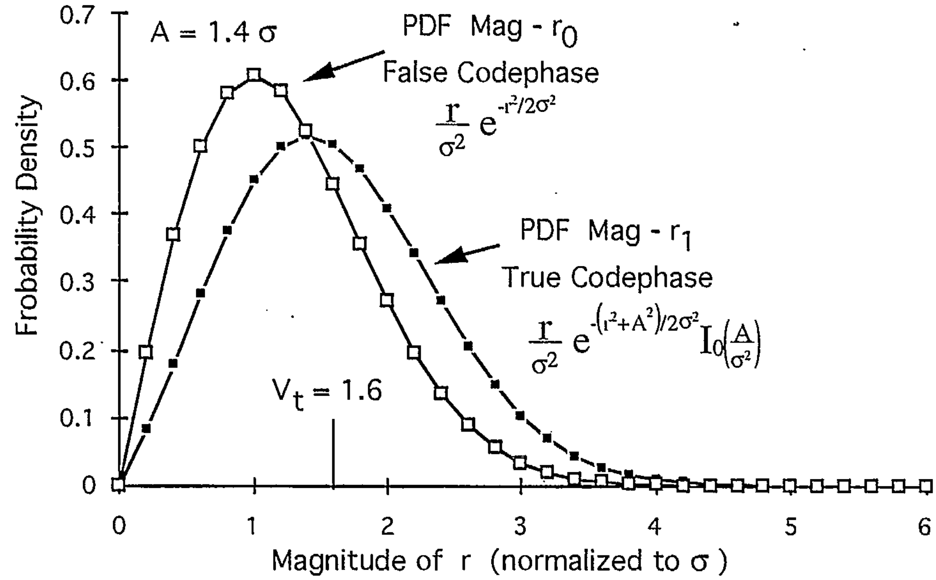

Fig. 5a PDF of bandpass filter envelope

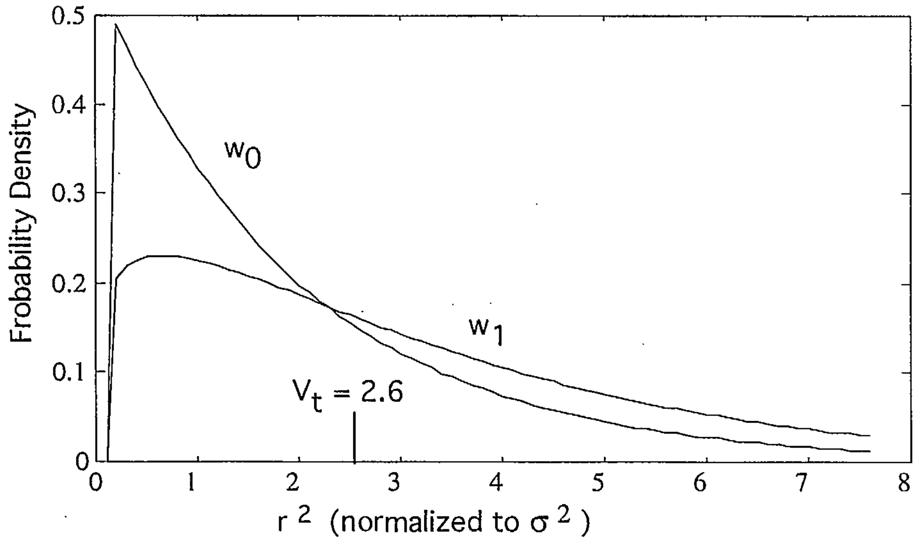

Fig. 5b PDF of squared bandpass filter output

The filter bandwidth, B, is in the order of one tenth the spread signal bandwidth ($B \approx 1/10T_c$) and it is convenient to represent the output of the bandpass filter by its samples taken at intervals $T_f = 1/B$ where T_f is the correlation time of the filter. The output of the squaring function is accumulated by integration over the dwell time T_d to form the decision variable z. The PDF of the integrator output z is shown in Fig.5c for $BT_d = 2$ or, in other words, $T_d = 2T_f$. The curves were obtained by convolution of the curves in Fig. 5b with assumption that, in either case H0 or H1, the samples of the squaring function output are independent and identically distributed. The variance of the decision variable z decreases relative to its mean and its distribution becomes more Gaussian as the integration time T_d is increased. The curves in Fig. 5d for the sum of 8 samples ($T_d = 8\ T_f$) were obtained by convolution of 8 PDF curves from Fig. 5b. As the integration time is increased there is increased accuracy in the decision between H0 and H1. Note that the PDF in Fig. 5b has been truncated and this has caused the small bump in the z1 curve of Fig. 5c.

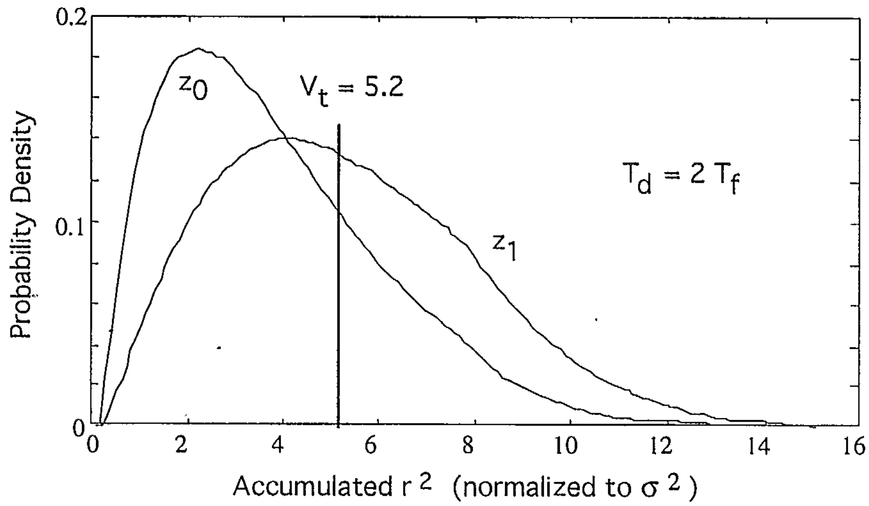

Fig. 5c Integrator output with $BT_d = 2$

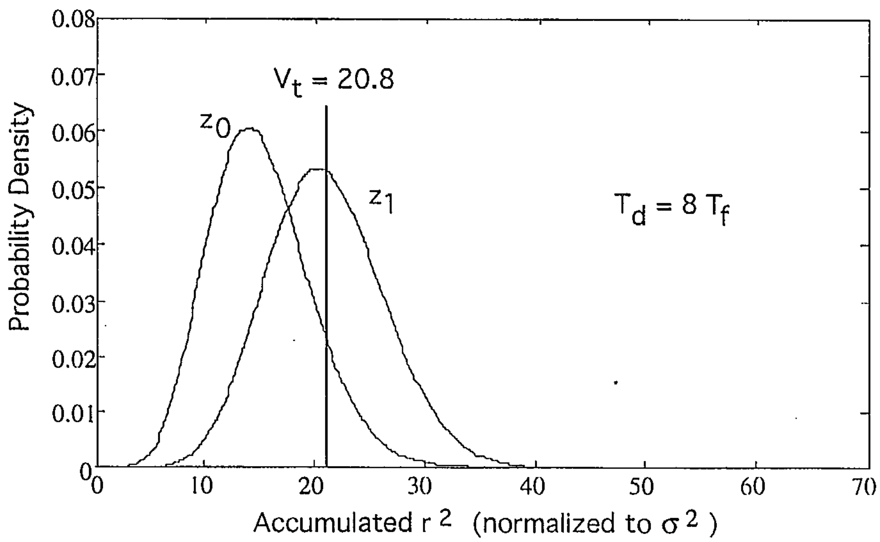

Fig. 5d Integrator output with $BT_d = 8$

2.2 Correlator serial search (coherent)

In cases where a synchronized carrier is available at the receiver, coherent demodulation can be used together with serial search at baseband. This is possible in the downlink where the portable unit receives a single carrier from the base station and all spread signals for the multiple users are modulated on the same carrier. Correlation at baseband is performed by simply multiplying the demodulated signal by the chip sequence of the reference PN generator and then integrating over the dwell time to produce the decision variable z. The dwell time is a multiple of the chip period, $T_d = MT_c$ where M is in the order of 300.

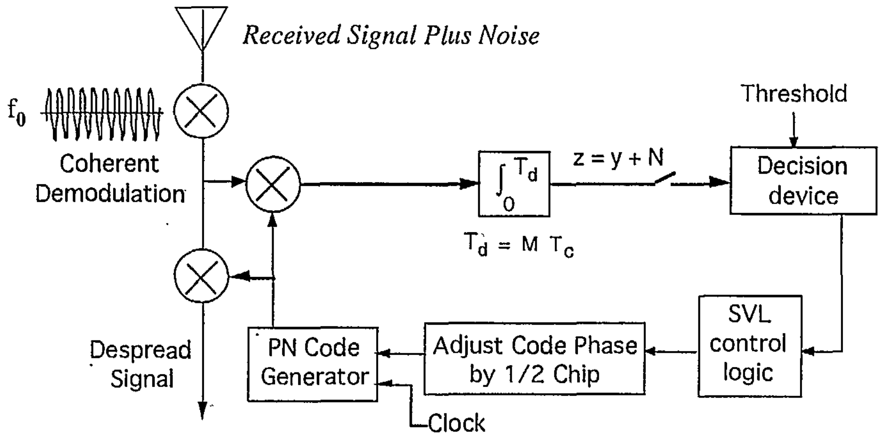

Fig. 6 Correlator serial search (coherent)

Decisions to adjust the local PN codephase are based on the random variable z which has components y and N such that z = y+N. The component y is the partial correlation of the desired received signal and the locally generated reference code. The component N is a random variable resulting from the correlation of the PN code sequence with the received random noise. The received random noise is in large part signals from other CDMA users and for this analysis the noise will be assumed to be white Gaussian noise with one sided power spectral density N_0.

The random variable y has mean m_i and variance s_i^2 where i=0 for a false codephase and i=1 for the true codephase. Assuming a spreading code with period L>M, the variance in y can be approximated using the autocorrelation of a random data stream. The variance has maximums when the chip boundaries align, relative minimums when the boundaries are offset by 1/2 chip time and is essentially zero when the codephases are aligned. The mean and variance of y are developed on page 554 of [Polydoros and Weber84] and, to facilitate comparison of the H0 and H1 cases, the expressions are presented in pairs as:

$$m_0 = 0 \qquad\qquad m_1 = \sqrt{S}\, MT_c\,(1-|p|) \tag{1,2}$$

$$\sigma_0^2 = S\, MT_c^2(1-|p|+2p^2) \qquad\qquad \sigma_1^2 = S\, MT_c^2(p^2) \tag{3,4}$$

where S is the power of the desired signal, T_c is the chip period, M is the dwell time in chip periods and p is the chip phase offset parameter. The parameter p is uniformly distributed in (-1, 1) and indicates the offset between the chip period of the local PN code generator and the chip period of the desired received signal. Ideally, the offset parameter p is zero.

The component N of the random variable y is a zero mean, independent, identically distributed Gaussian random variable with mean and variance

$$m_n = 0 \tag{5}$$

$$\sigma_n^2 = N_0 MT_c/2 \tag{6}$$

where N_0' is the effective one sided noise power spectral density at the receiver input. The term N_0' includes co-user noise as well as background noise. The effective noise power spectral density can be approximated as $N_0' = (U-1)\, PTc + N_0$ where U is the number of active users and P is the average power per user.

The variance in the detection variable z is the sum of the partial correlation variance and the correlation variance due to receiver noise. The receiver noise has no contribution to the mean of the detection variable z.

$$m_{Z0} = m_0 = 0 \qquad\qquad m_{Z1} = m_1 = \sqrt{S}\, MT_c\,(1-|p|) \tag{7,8}$$

$$\sigma_{Z0}^2 = \sigma_n^2 + \sigma_0^2 \qquad\qquad \sigma_{Z1}^2 = \sigma_n^2 + \sigma_1^2 \tag{9,10}$$

In the case of multi-user CDMA systems, the noise exceeds the signal and the system operates with E_c/N_0' much less than one. The worst case partial correlation component, σ_S^2, is much less the noise component.

$$\sigma_S^2 = \frac{E_c}{T_c} MT_c^2 \;\ll\; \sigma_n^2 = \frac{N_0'/2}{T_c} MT_c^2 \tag{11}$$

An illustration of the probability density function of the detection variable z is shown in Fig. 7 for the two cases of false codephase (hypothesis H0) and correct codephase (hypothesis H1). The receiver bandwidth is assumed to be $1/T_c$ and the correlation length is assumed to be

M=100. The chip alignment is assumed to be exact and the partial correlation component of the signal is assumed small in comparison with the noise component ($E_c/N_0' << 1$). The ratio of chip energy to noise power spectral density is assumed to be 0.05, that is $E_c/N_0' = -13$ dB. The horizontal axis has been normalized to the standard deviation σ_n. The normalized mean value m_1 of z at the correct codephase is calculated as follows:

$$m_{1n} = \frac{m_1}{\sigma_n} = \frac{\sqrt{E_c/T_c}\, M\, T_c}{\sqrt{N_0'/2T_c}\,\sqrt{M}\, T_c} = \sqrt{\frac{E_c\, M}{N_0'/2}} \tag{12}$$

$$m_{1n} = \sqrt{(0.1)(100)} = 3.1 \quad .$$

A decision threshold is set to decide between the two hypotheses H0 and H1 based on the decision variable z. In Fig. 7 the threshold is shown at 2σ. Falsely assuming the true codephase is known as a false alarm and this has probability P_{fa} indicated by the hatched area. The probability of missing the correct codephase ($1-P_d$) is illustrated by the shaded area. When the codephase uncertainty region is long or the SNR is poor, it is desirable to adjust the threshold and decrease P_{fa} at the expense of P_d in order to get a minimum acquisition time.

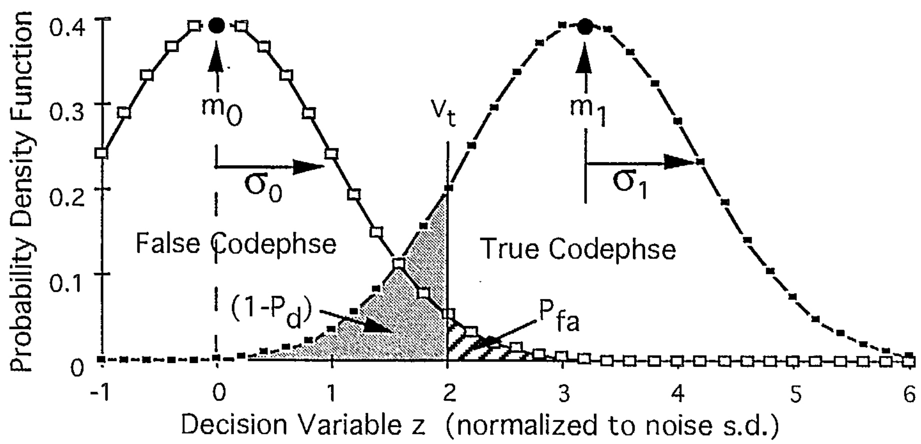

Fig. 7 Correlator output PDF (coherent detection)

2.2.1 SNR disadvantage of the non-coherent correlator

It is possible to compare the coherent and non-coherent systems by equating the matched filter time domain averager and the despreader/bandpass filter . The non-coherent detection system utilizes a despreading multiplier, a bandpass filter, a square law detector which is integrated for a dwell time to produce the decision variable, z. Expressions for the non-coherent decision variable mean and variance have been developed for the false codephase (H0) and true codephase (H1) cases [Hopkins77]. After some manipulation, the mean, μ, and standard deviation, σ, can be expressed as follows. The expressions for the non-coherent correlator are presented in pairs to facilitate comparison of the two cases.

$$\mu_0 = N_0' B \qquad \mu_1 = N_0' B + P \tag{13,14}$$

$$\sigma_0 = \frac{m_0}{\sqrt{B\,T_d}} \qquad \sigma_1 = \frac{m_1}{\sqrt{B\,T_d}} \tag{15,16}$$

Where B is the bandwidth of the predetection filter, Td is the postdetection integration time and N_0' is the effective noise power spectral density. The difference in the means, normalized to the standard deviation σ_0, is

$$\frac{\Delta\mu}{\sigma_0} = \frac{\mu_1 - \mu_0}{\sigma_0} = \frac{P}{N_0' B / \sqrt{B\,T_d}} \tag{17}$$

The bandpass filter can be equated to a transversal filter of length T_f and bandwidth $B = 1/T_f$. Substituting T_d for 1/B in (17) yields

$$\frac{\Delta m}{\sigma_0} = \frac{E_c / T_c}{N_0' / T_f}\sqrt{\frac{T_d}{T_f}} = \frac{E_c}{N_0'}\left(\frac{T_f}{T_c}\right)\sqrt{\frac{T_d}{T_f}} \tag{18}$$

The bandpass filter therefore integrates over $M_1 = T_f/T_c$ chip times and the dwell time integrator dwells for $M_2 = T_d/T_f$ units of the filtering time. The total time taken for dwell at each codephase is thus $T_d = M_1 M_2 T_c$ Example values of M_1 and M_2 are 22.4 and 455 [Hopkin77] or 3.86 and 1586 [Alem et al 78]. Equ. 18 can thus be rewritten as

$$\frac{\Delta m}{\sigma_0} = \frac{E_c}{N_0'} M_1 \sqrt{M_2} \tag{19}$$

The performance of the non-coherent detector can be improved by increasing M_1 (relative to M_2) and thus decreasing the bandwidth, B, of the predetection filter. The reduction in bandwidth is limited by Doppler frequency uncertainty, data modulation bandwidth and filter output risetime during the dwell period.

For almost-coherent correlators such as the one discussed in Section 3.1, I and Q correlators integrate over the dwell time to produce two outputs which are then combined by squaring and summing to form the decision variable. The square root of the decision variable is illustrated in Fig. 8. In the true codephase position and under conditions which result in acceptably low P_{fa}, the squared output can be approximated by the square of the inphase mean.

$$m_1^2 \;=\; \frac{E_c}{N_0'/2}\,M \qquad (20)$$

where M is the correlator length in chip times. In a false codephase position, the squared output as shown by w_0 in Fig. 5b has mean, m_0^2, approximately 1.6 and its standard deviation σ_{20} is approximately 2. Under good conditions (low P_{fa}), m_1^2 is large and m_0^2 can be neglected. The probability P_{fa} can then be calculated from the ratio of standard deviation and the difference in the mean of the squared output and it can be seen from (21) that the coherent correlator represents an improvement of $\sqrt{M_2}$ when compared with the non-coherent correlator.

$$\frac{\Delta m^2}{\sigma_{20}} \approx \frac{E_c}{N_0'}\,M \qquad (21)$$

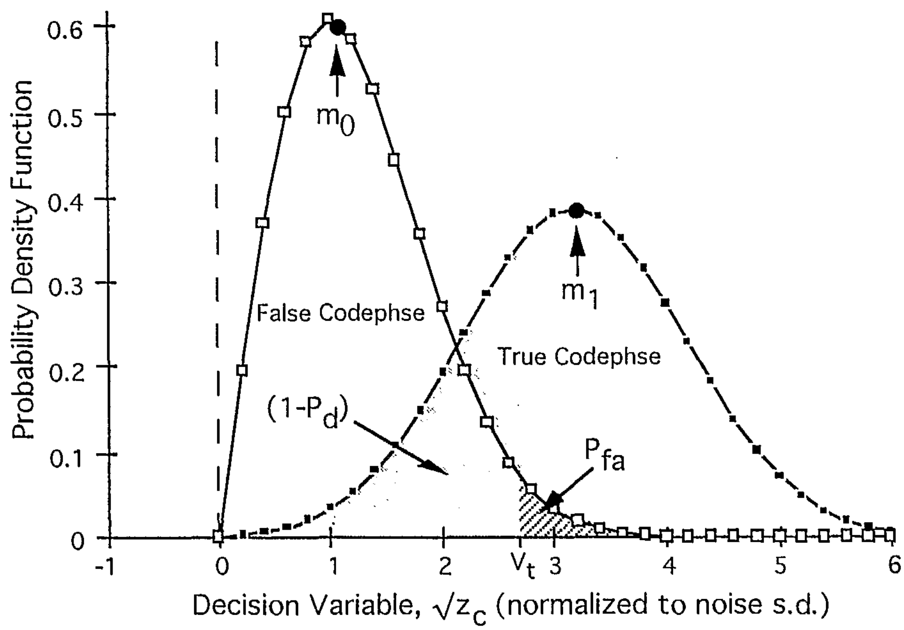

Fig. 8 Illustration of non-coherently combined I-Q output $\sqrt{z_c}$

2.3 Matched filter search (coherent)

Coherent demodulation and baseband processing is assumed in this section in order to discuss basic matched filter operation. The primary advantage of the matched filter (MF) codephase detection technique is that out of code phase cells are rejected at the rate at which they arrive and much faster codephase acquisition is achieved than for correlator serial search. As in correlator serial search, the MF can be operated on a fractional chip interval ΔT_C where $\Delta = 1/2$ or $1/4$, etc. For clocked delay lines, the demodulated signal is processed by a "chip matched filter" which then delivers samples to the analog delay line. The length of the matched filter must be less than a data bit duration and, for simplicity, it is assumed for the time being that no data transitions occur during the acquisition period.

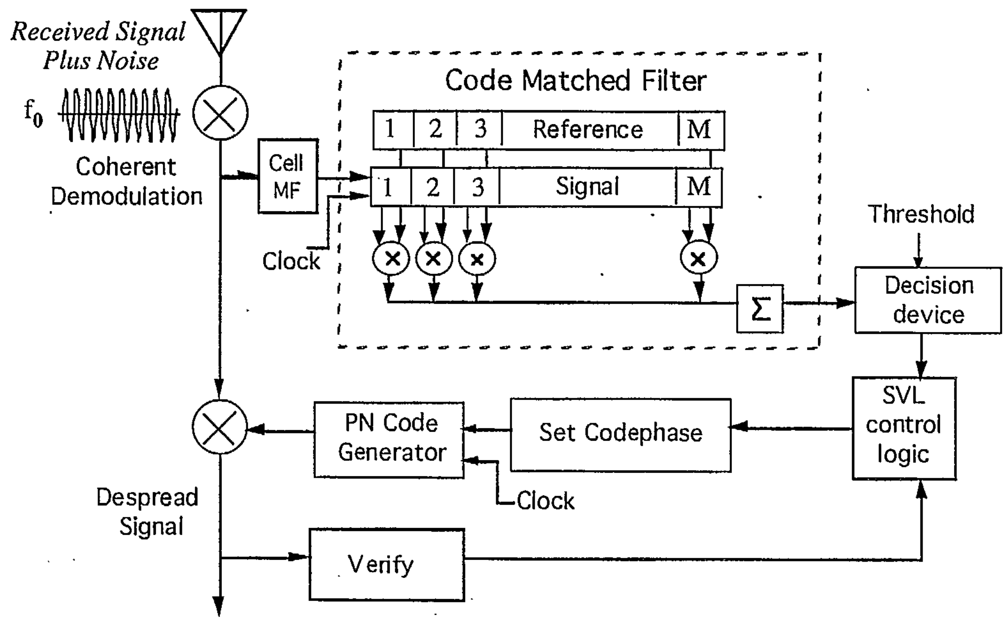

Fig. 9 Matched filter search (coherent demodulation)

The term matched filter is used to specifically mean a transversal matched filter in which a short history of the incoming signal is stored in an analog delay line. Past samples of the input signal are available at taps on the delay line and the retained history in the delay line is in the order of 300 chip times. These samples are first multiplied by chip elements of the reference code and then summed. This result is equivalent to a serial correlator operating on the same baseband input signal over the same 300 chip times. The MF is a "pipelined" rather than a "batch" process and thus correlation outputs are continuously available as the spread spectrum signal is received. In practice, an analog shift register or multibit digital shift register is used and the correlation output occurs at 1/2 or 1/4 chip intervals. The example PDF for a correlator of duration 100 chip periods as shown in Fig. 7 of Section 2.2 is also valid for a matched filter with length 100 chip periods if the same conditions and assumptions are applied in both cases.

An illustration of a MF is shown in Fig. 9. There are two shift registers, one is used to store the reference code and the other stores and shifts samples of the input signal. The data content of the reference shift register is static and the values stored are only changed if a different spreading code is to be acquired. To effectively use TMF synchronizers, the delay line clock period must accurately represent a submultiple of the chip period of the code sequence to be detected.

Matched filters with analog shift registers use surface acoustic wave (SAW) or charge coupled device (CCD) technology. SAW matched filters are discussed by Milstein and Das in [MiDa77] and Baier et al in [BaDoPa82]. Charge-coupled device (CCD) matched filters are dealt with by Grieco [Griec80] and Magill et al [MaGrDyCh79].

A digital implementation with eight level 64 tap I-Q matched filters is used in the low cost ($50) Stanford Telecom STEL-2000 spread spectrum processor chip. The device is designed for one data bit per code period and is designed for a maximum PN code length of 64. The signal shift register has 128 elements with a tap at every second element to allow processing with 2 samples per chip. A preliminary data sheet is included in appendix B.

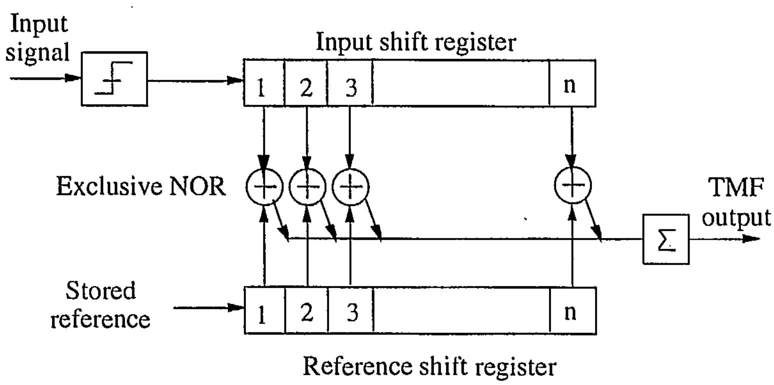

Fig. 10 Block diagram of Digital baseband TMF (from Dixon)

A simple digital implementation of a transversal MF is illustrated in Fig. 10. A hard limiter converts the incoming signal into binary levels as required by the digital shift register. This input quantizer, whether binary or multilevel, adds quantizing noise to the receiver input signal and further degrades E_c/N_0'. If the input signal is assumed to have a Gaussian amplitude distribution and the uniform quantizer is adapted to have an optimum step size relative to the received signal, the added quantizer noise can be obtained from [J. Max60]. A two level quantizer will have quantizer noise which is -4.4 dB relative to the received signal amplitude. Since the received signal is primarily noise, the desired signal can be assumed constant and the noise power to have increased by a factor of $1 + 10^{-.44} = 1.363 = 1.35$ dB. A four level quantizer adds a relative quantizing noise of -9.25 dB which increases the receiver noise by the factor of 1.119 (0.49 dB). An 8 level quantizer has only a 0.16 dB degradation on the receiver E_c/N_0. There is little advantage in using more than 8 levels of quantization in a digital matched filter.

It will be seen in section 2.5 that a factor of 2 increase in the MF length gives a performance improvement equivalent to 3 dB improvement in the Ec/N0'. Only 0.86 dB improvement is gained from using 2 bit (four level) quantization which also represents a factor of 2 increase in the matched filter elements. Alternately stated, an increase in the binary shift register length of only 22% is equivalent to changing the shift register from 2 level to 4 level.

2.4 Optimum Threshold (coherent)

An optimum threshold can be selected to give P_{fa} and P_d which minimize synchronization time. Coarse codephase acquisition time can be determined by using these transition probabilities P_{fa}(or P_{f1}) and P_d, the corresponding probabilities in verification mode, P_{f2} and P_{d2} and a flow graph of the acquisition process as shown in Fig. 11. The flow graph has transition times equal to the dwell time T_d (or T_{d1}) and the verification time T_v (or T_{d2}).

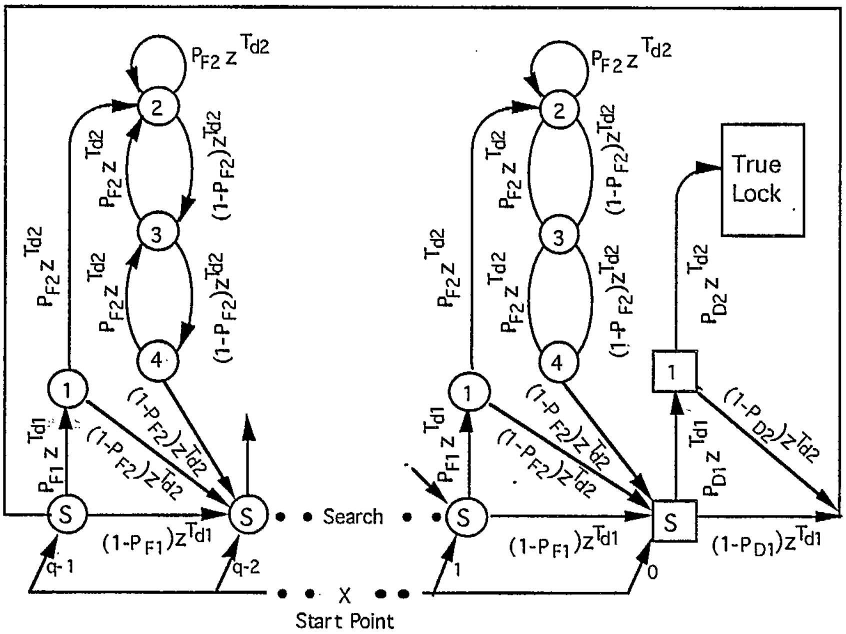

Fig. 11 Flow graph of the search process

The mean acquisition time, $E\{T_{acq}\}$, for spread spectrum acquisition has been developed in [Pan, Dodds & Kumar90]. To emphasize the significant terms, the expression for $E\{T_{acq}\}$ can be written as (14) where T^* represents the minor components. . The small value T^* includes a possible return to search mode and re-entry to verification. The time approaches T_V+T_d/2 as the verification time, T_V, increases and the probability of verification error approaches zero. The mean acquisition time is

$$E\{T_{acq}\} = q\left(\frac{1}{2}+\frac{1-P_d}{P_d}\right)T_d + P_{fa}\,(q-1)\left(\frac{1}{2}+\frac{1-P_d}{P_d}\right)T_{fa} + T^* \qquad (14)$$

which can be simplified to

$$E\{T_{acq}\} = \left(\frac{1}{P_d}-\frac{1}{2}\right)\left(qT_d + P_{fa}\,(q-1)\,T_{fa}\right) + T^* \qquad (15)$$

where q is the number of cells in the code period, T_d is the dwell time in search mode, T_V is the dwell time in verification mode, P_{fa} is the probability of mistaking a false codephase as the true one (false alarm), P_d is the probability of properly detecting the true codephase and T_{fa} is the average time spent in the false alarm state. The term T_{fa} may include lock time as well as verification time.

The acquisition process will start, on average, at a codephase which is displaced from the true codephase by one half the code period, (q/2). Upon reaching the true codephase, it has a small probability, $(1-P_d)$, of not recognizing it and continuing to search the next codephase. It is expected that the code cycle will be further searched $(1-P_d)/P_d$ times after the true codephase is reached. With the exception of the true codephase, at each searched cell there is the probability P_{fa} of entering the false alarm state.

Given a PDF for the decision variable, an optimum threshold can be set which will result in P_{fa} and P_d which give a minimum mean acquisition time. For optimum performance, the threshold should be adjusted so that the search process is delayed by false alarm and delayed by recycles in approximately equal proportion. The sum of these two delays is approximated by:

$$\text{Delay} = P_{fa}\, q\, T_{fa} + q\, T_d\, (1-P_d) \qquad (16)$$

In acquisition there will be roughly $q\, P_{fa}$ occurances of false alarm, each with duration T_{fa}. A recycle will cause a delay in the order of $q\, T_d$ and this will occur with missed detection probability $1-P_d$. In correlator systems, $T_{fa} \approx 5\, T_d$ typically. In a matched filter system, as shown in Fig. 9, verification uses a separate correlator and $T_{fa} = q\, T_d$ since the code sequence in the MF taps is not expected to be matched until one code cycle later. It may be possible to reduce this time by setting a new reference code segment in the matched filter.

If the two delay components are to be equal, then $P_{fa} << (1-P_d)$, and this is especially true for the matched filter. For the correlator, the threshold should be set such that $5\, P_{fa} \approx (1-P_d)$ and for the matched filter, the threshold should be set higher such that $q\, P_{fa} \approx (1-P_d)$. Fig. 12 shows the PDF of the decision variable in the H0 and H1 cases. Chip synchronization has been assumed (one sample per chip) and the correlation time and the matched filter length are both 100 T_c. Acquisition time in Fig.13 for a range of threshold values assuming $T_{d1} = 100\, T_c$ and $T_v = 5\, T_{d1}$ for the correlator and assuming $T_{d1} = T_c$ and $T_v = 1023\, T_c$ for the matched filter. The optimum thresholds are clearly different for the two techniques.

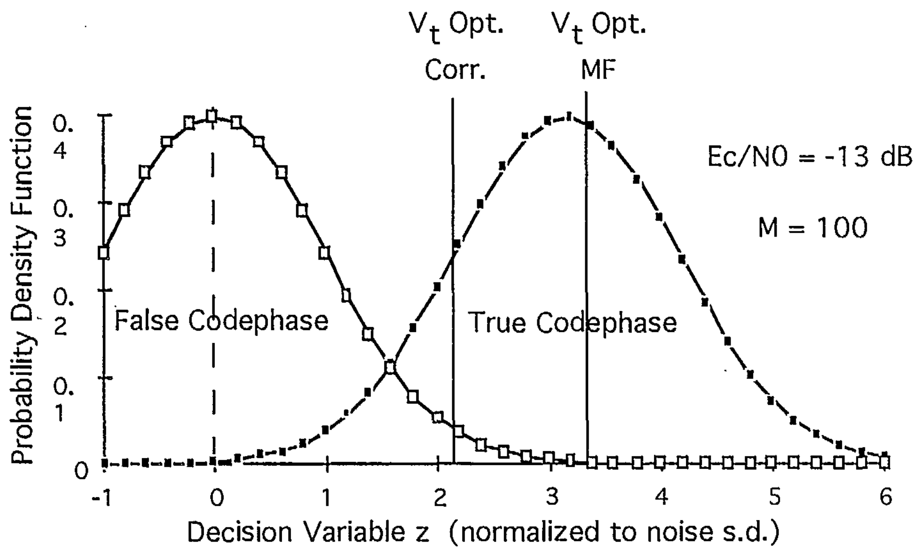

Fig. 12 Coherent detector PDF with optimum thresholds

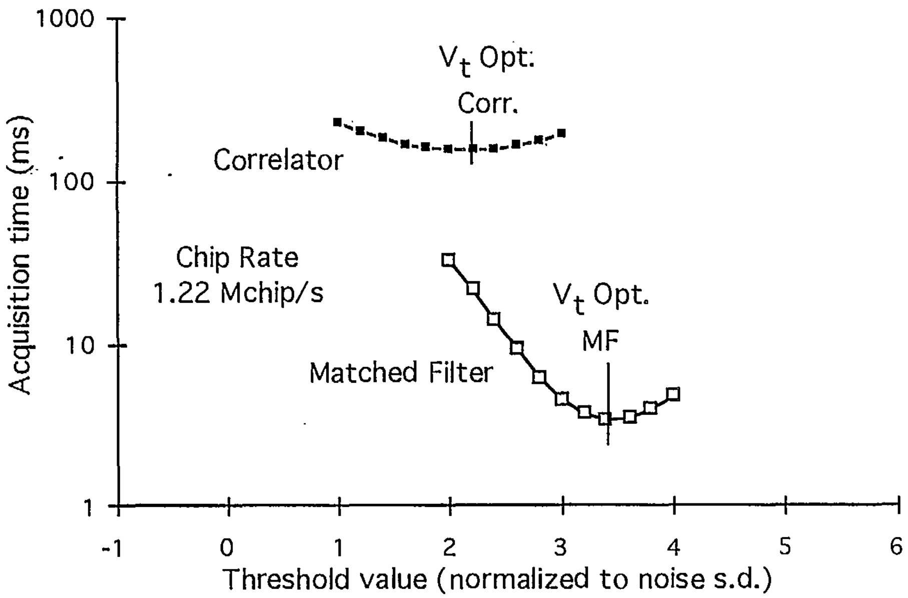

Fig. 13 Acquisition time vs threshold voltage

2.5 Optimum MF & correlator length (coherent)

Acquisition time performance has been investigated as a function of E_c/N_0 and of the PN code length. The performance has been computed as a function of matched filter length (in chips) or correlation time (in chips) and the minimum value over a range of threshold values has been recorded. For this section, Doppler offset frequency is assumed to be zero and coherent demodulation is assumed; a single MF or correlator is required since there is no quadrature information. Perfect chip alignment is assumed and it is assumed that there are no data transitions. It is assumed that no quantizing noise occurs.

In the calculation of $E\{T_{acq}\}$, the dwell time in verification for correlator serial search was assumed to be $T_v = 5\ T_d$ and for the matched filter it was assumed that a full code cycle would elapse for verification, $T_v = L\ T_c$. The probabilities in search mode were calculated from E_c/N_0', the correlation or matched filter length and the threshold. In verification mode, the threshold was set so that $P_{f2} = 0.031$ which meant that P_{d2} varied with E_c/N_0' and matched filter or correlator length. The important factor is that P_{d2} be close to unity which was the case since the dwell time was long.

Fig.14 shows calculated results for acquisition time vs. correlation length and code period for $E_c/N_0' = -10$ dB. Similar results are shown in Fig.15 for -20 dB. As the dwell time is increased, there is a certain value which optimally balances the increased time taken at each codephase step against the decreased time taken in false alarm and in recycles. As the code length increases form 1023 to 131,071, the optimum dwell time is seen to increase from 80 to 160 chip times at $E_c/N_0' = -10$ dB and from 400 to 1200 chip times at $E_c/N_0' = -20$ dB. The minimum acquisition time is approximately 1.5 LT_{d}opt. It can be seen in Appendix C that the optimum Td increases by an approximate factor of 2.5 for every 5 dB increase in noise.

Figs.16 and 17 show calculated results for acquisition time vs. MF length and code period. As the MF length is increased, the time taken in false alarm decreases to zero, leaving only 0.5 LT_c which is the time required to dwell at each codestep. Minimum acquisition time is reached at M = 300 chips for $E_c/N_0' = -10$ dB and at 3000 chips at $E_c/N_0' = -20$ dB. The optimum MF length appears to be independent of code length. It can be seen from Fig.19 and from Appendix C that the optimum value of M increases by a factor of 3.2 for every 5 dB increase in noise.

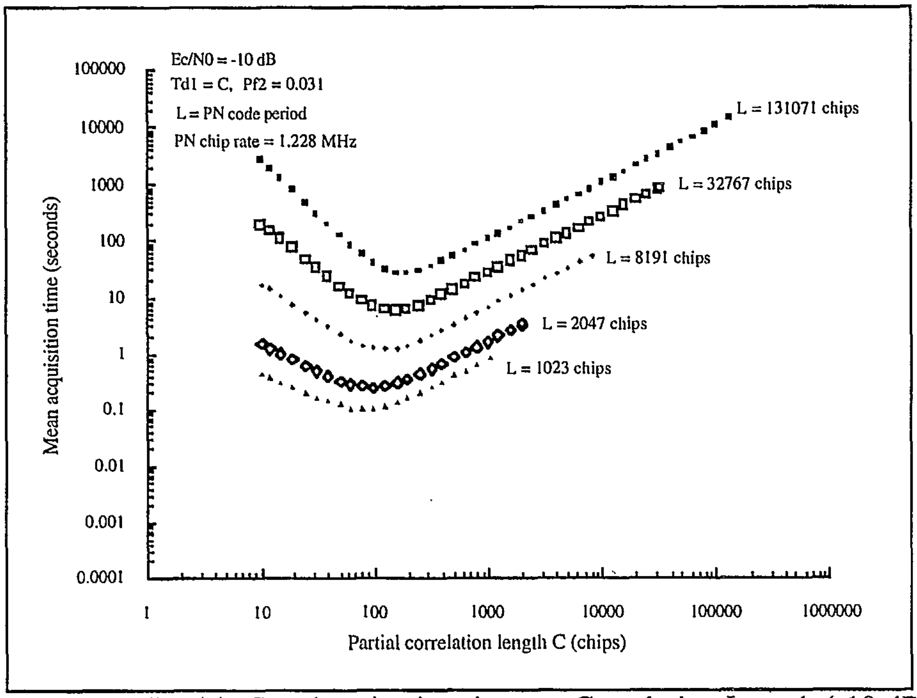

Fig. 14 Synchronization time vs. Correlation Length (-10 dB)

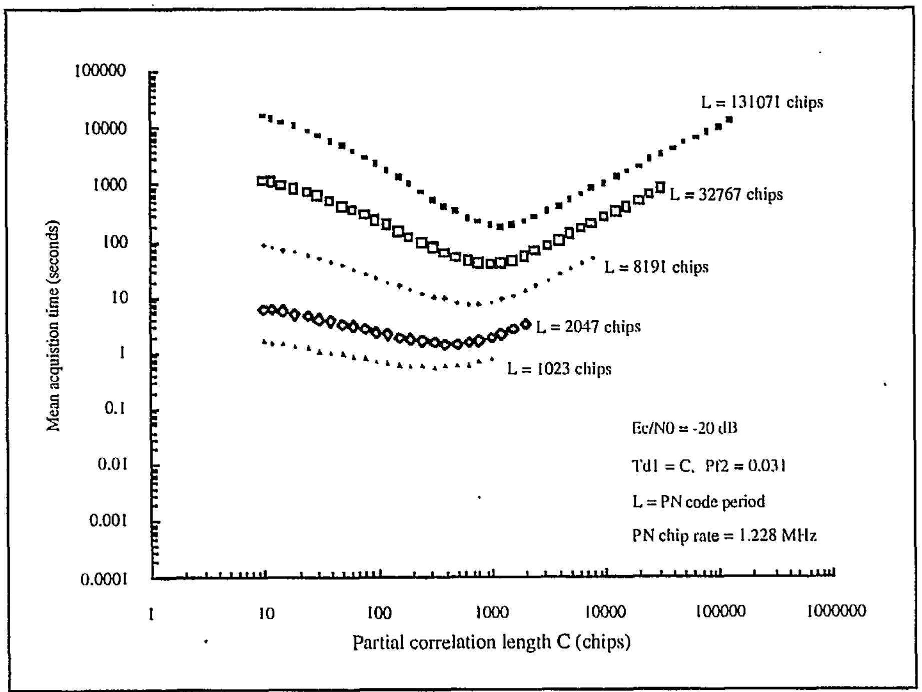

Fig. 15 Synchronization time vs. Correlation Length (-20 dB)

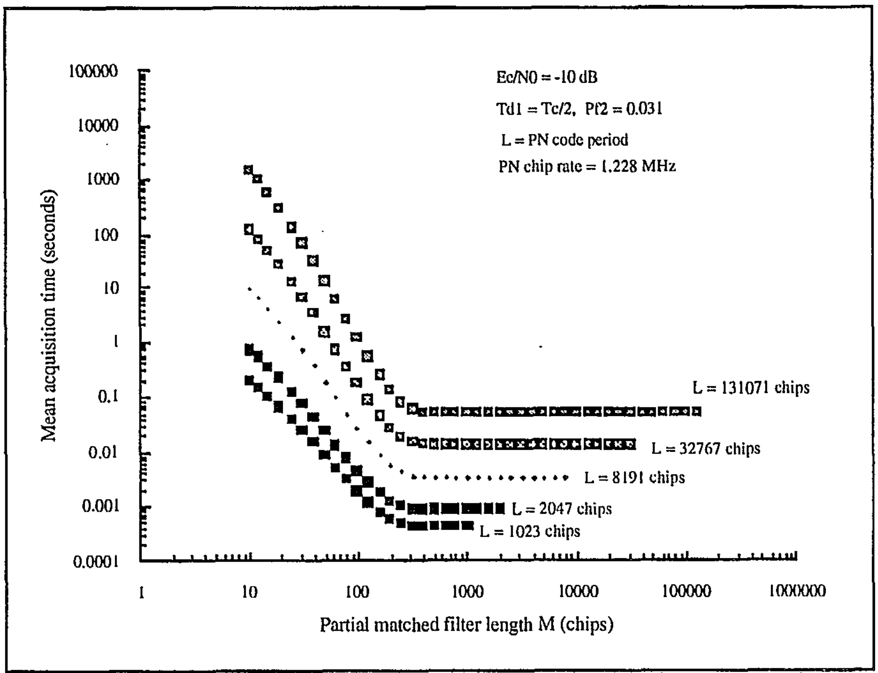

Fig. 16 Syncronization time vs Matched Filter Length (-10 dB)

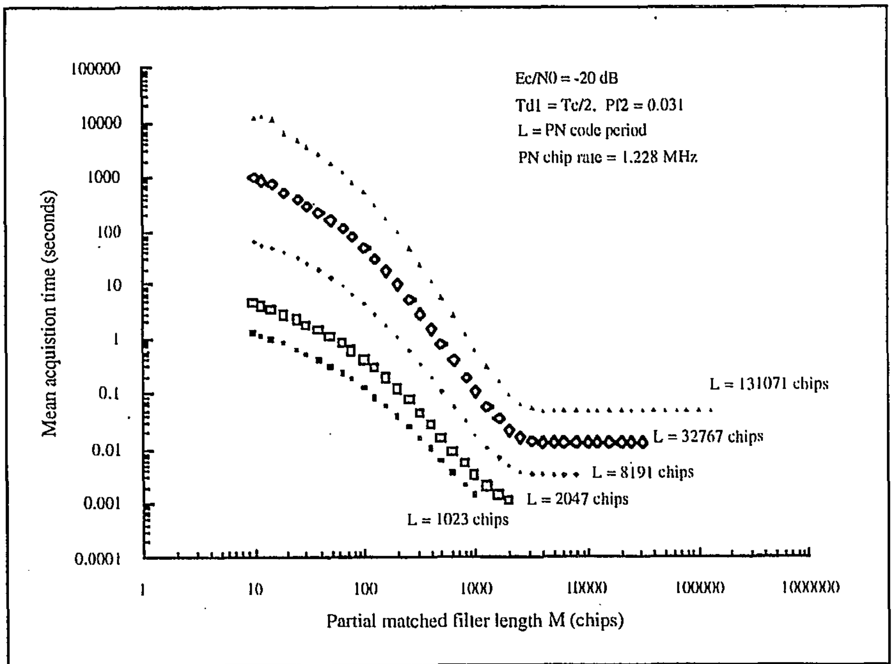

Fig. 17 Syncronization time vs Matched Filter Length (-20 dB)

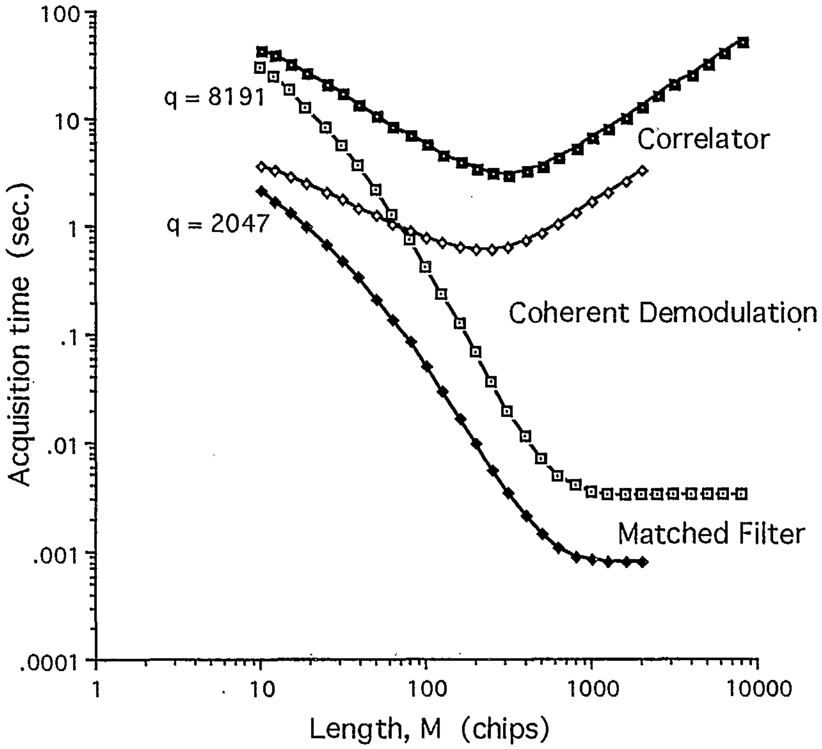

Fig. 18 Performance of MF and correlator search (E_c/N_0 = -15 dB)

Fig. 18 compares the performance of MF and correlator search assuming coherent demodulation. At short lengths, the process has many false alarms and both techniques have about the same acquisition time since, in this analysis, the false alarm time, T_{fa}, was assumed to be the same in both cases. As the length increases, the false alarm probability decreases, the total time spent in false alarm decreases and the acquisition time decreases. For the matched filter, the dwell time, T_d, remains unchanged, and the acquisition time simply decreases to the product of the dwell time and the number of codephase cells searched (L/2). For the correlator, the time spent in each dwell increases as the length increases and product of dwell time and the number of codephase cells dominates the acquisition time.

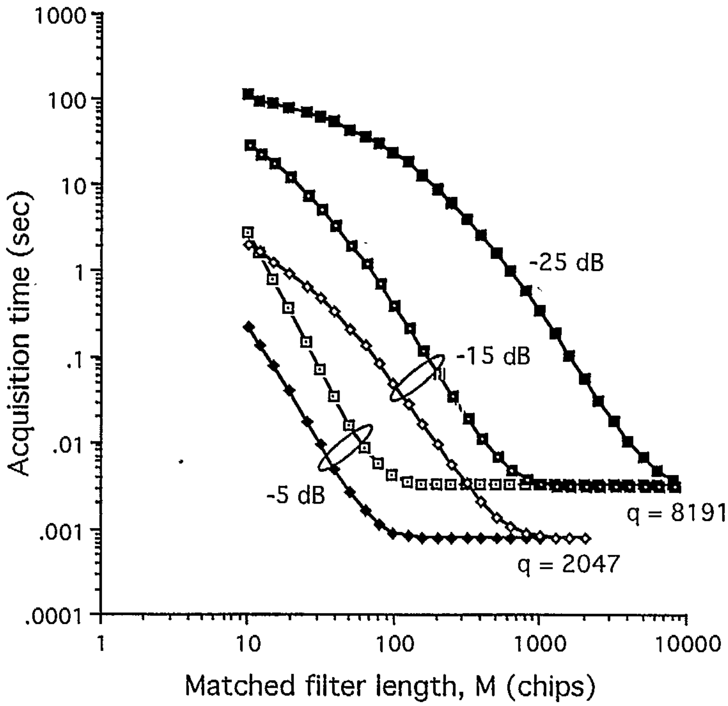

Fig. 19 Acquisition time vs. MF length and E_c/N_0

At high noise or low E_c/N_0, the acquisition time is slowed by false alarms. As the matched filter length is increased and the threshold accordingly adjusted, the false alarm probability and the time spent in false alarm is reduced. For equivalent performance, a factor of 10 increase in the matched filter length is required to compensate for a factor of 10 reduction in E_c/N_0. This is consistent with Equ. 12 in Sect. 2.2.

2.6. Other Schemes for Acquisition

Other schemes for spread spectrum acquisition include sequence estimation, pilot assisted acquisition and differential detection.

2.6.1 Direct subsequence estimation

Rapid acquisition by sequential estimation (*RASE*) was first proposed by Robert B. Ward in 1965 [Ward65]. The *RASE* system makes an estimate of the first 15 received chips, loads a local PN sequence generator with that estimate, and then starts operation of the PN code generator and the tracking circuits. If this verification fails, a new estimate is made and thus a new sequence of estimates may be made. The *RASE* scheme is much faster than the traditional serial search scheme in the case of good SNR and long PN code period. Unfortunately, the system is vulnerable to noise and interference because the estimate is performed on a chip-by-chip basis. Furthermore the code phase estimate is based directly on the received signal and therefore polarity reversals due to data modulation or non-coherent demodulation will corrupt the estimate. However it can tolerate some Doppler shift of the carrier frequency because the code length required for the estimate is very short.

Kilgus [Kilg73] in 1973 suggested a modified method applying a majority logic decoder which uses the recursive relation of the PN code to improve the initial n-chip estimate. Another modification of Ward's initial *RASE* system, rapid acquisition by recursion-aided sequential estimation (*RARASE*), was also investigated by Ward and Yiu in 1977 [WaYi77]. The *RARASE* scheme reduces the acquisition time by a factor of 7.5 for a $(2^{15} - 1)$ PN code length. These methods are also restricted to "baseband" or coherent systems and the estimate can also be corrupted by polarity reversals caused by data modulation. It should also be noted that these methods are not suitable for Gold codes.

2.6.2 Pilot tone method

Some CDMA systems make use of a pilot signal to facilitate code phase synchronization. These can be divided into two categories: pilot symbols and pilot tones. A pilot symbol is simply the spreading code used in the system which is not modulated by data at all. This allows use of the matched filter estimation schemes. Once synchronization of the pilot symbol is achieved, it provides timing information which can be used to synchronize spread spectrum data channels.

A pilot tone may also be used for code phase synchronization. One method utilizing pilot tones is to amplitude modulate a carrier so that modulation zero crossings correspond to the spreading code period boundaries. Note that the pilot tone technique provides an estimate which may require some additional serial code phase search. A novel technique which uses amplitude modulated pilot tones to aid PN code acquisition has been proposed [BrSaDo91]. The transmitter sends one or more AM pilot tones along with the spread spectrum signal. The zero-crossings of the demodulated AM tones mark the received pseudo-noise (PN) code period boundaries, and trigger the local PN code generator. The study indicates that the acquisition time may be faster than traditional serial-search scheme with the similar circuitry complexity. Pilot tones may suffer frequency selective fading on multipath channels, which would degrade acquisition performance. This scheme could be applied to the digital cellular systems by using CDMA system time as a universal reference. Aided PN code acquisition using the pilot signal technique is of interest in systems where interception is not of concern.

2.6.3 Differential detection

If data modulation is present, a transition based estimate can be used. This proposed scheme uses a transition sequence detected by matched filter (*TS/MF*). The *TS/MF* scheme is implemented by inserting the transition detector (between A and B) to the *MF* scheme. In the transition detector as shown in Fig. 20, the received spread spectrum signal is first delayed by exactly one chip interval, then the delayed version is multiplied with the received signal. At high SNR, the product indicates the location of transitions in the received signal. A positive output indicates no transition while a negative output indicates a transition in the received signal. The transition detector makes the acquisition process almost insensitive to polarity reversals of the input data signal and small frequency offsets in demodulation.

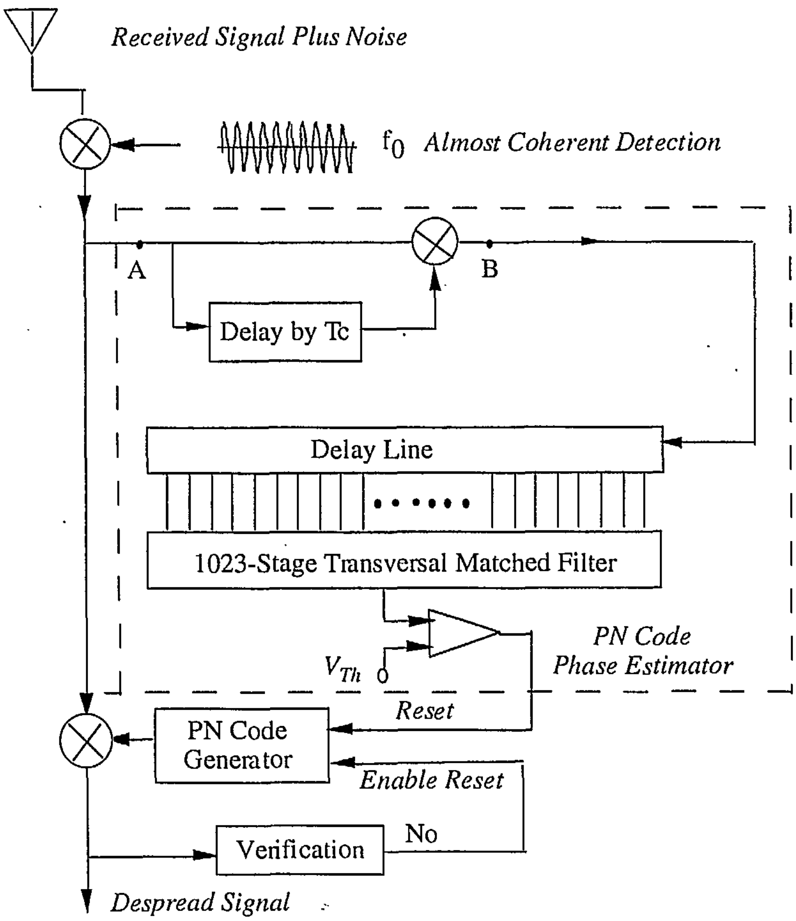

Fig. 20 TS/MF acquisition scheme

After passing through the transition detector, the received signal is then tested in the matched filter. The output from the matched filter is compared with a preset threshold (V_{Th});

this is one of the most important parameters affecting system performance. When the MF maximum output is sensed, the local PN code generator is reset to its initial state and the code phase estimate is complete. Unfortunately, the transition detector squares the noise power so that the noise rejection capability of *TS/MF* scheme is much less than that of *MF* scheme.

A variation of this scheme uses the matched filter delay line for cyclic averaging. If the delay is exactly equal to the period of the PN code then a fraction (90%) of the delayed signal can be added to the input signal to gain some noise reduction by averaging. This is called cyclic accumulated transition sequence detected by a matched filter (*CATS/MF*) . After passing through a transition detector circuit, it is then accumulated by the 1023-stage cyclic accumulator and then detected with the 1023-tap matched filter. Equivalent results could be obtained by extending the length of the matched filter.

2.7. Scheme comparison

The code phase search schemes may be classified as illustrated in Fig. 21. Methods can be divided into correlator serial search, sequence estimation and direct or transition based matched filter schemes.

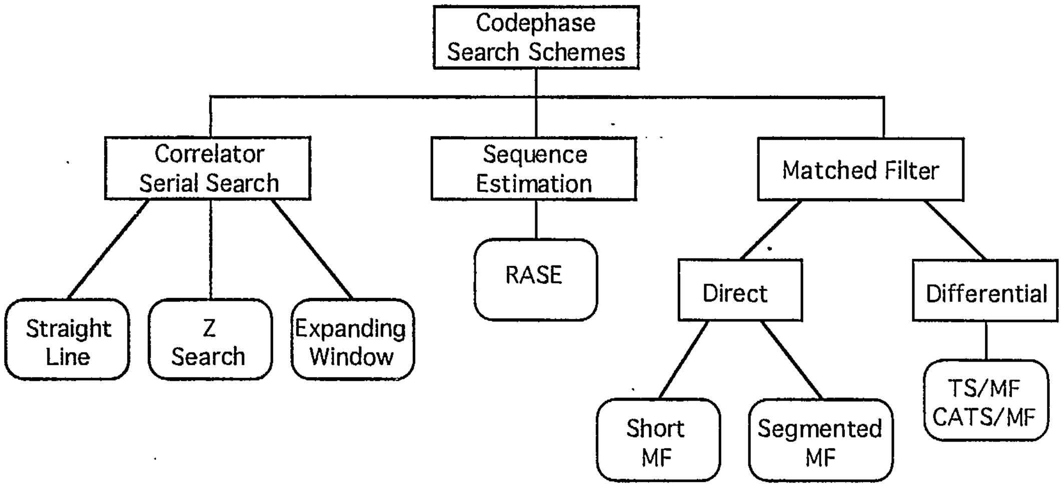

Fig. 21 Code phase acquisition schemes

The expected performance for the discussed schemes in previous section are summarized in Table. 1. Among these acquisition techniques, the serial search scheme is the most mature and popular technique because of its relative simplicity and good performance in the presence of strong interference and noise. The major disadvantage of the serial search is its long acquisition time.

The *RASE* and related schemes which are based on the "code state" estimation are faster than the traditional serial search scheme for high SNR and long PN signal periods. But it is restricted only to "baseband" or coherent systems and the estimate can be corrupted by polarity reversals of the source data. The *RASE* technique has the drawback of being highly vulnerable to noise and interference signals due to the chip-by-chip based estimation.

Short matched filter schemes require coherent demodulation and can tolerate only minor Doppler shift of the carrier frequency and no data modulation transitions. In this analysis, zero Doppler and perfect chip synchronization has been assumed. By introducing the transition detector circuit, the *TS/MF* and *CATS/MF* schemes allow the acquisition process insensitive to polarity reversals of the input data signal and small frequency offsets in the demodulation. Because the transition detector squares the noise power, the noise rejection capability is poor compared to the *MF* scheme. A comparison of acquisition times for PN sequence length 1023 and for correlator and matched filter lengths 1023 is shown in Fig. 22. In this simulation work, the thresholds were not accurately optimized, however, the performance ranking of the schemes is clear.

Not shown is the performance of the pilot tone technique which is expected to be somewhat faster serial search [BrSaDo91]. Since the pilot signal is narrowband, the acquisition performance would be degraded in multipath fading environments. Pilot signals take extra power and will appear as additional background noise if they are sent inside the spread spectrum bandwidth.

Table 1. Comparison of main acquisition approaches

Scheme	Speed	Complexity	Noise tolerance	Doppler tolerance
Serial Search	slow	low	good	Yes
RASE	very fast	low	very poor	Some
MF	fast	high	good	No/Yes
TS/MF	fast	high	fair	Yes
CATS/MF	fast	highest	fair-good	Yes
Pilot signal	moderate	low	fair	Yes

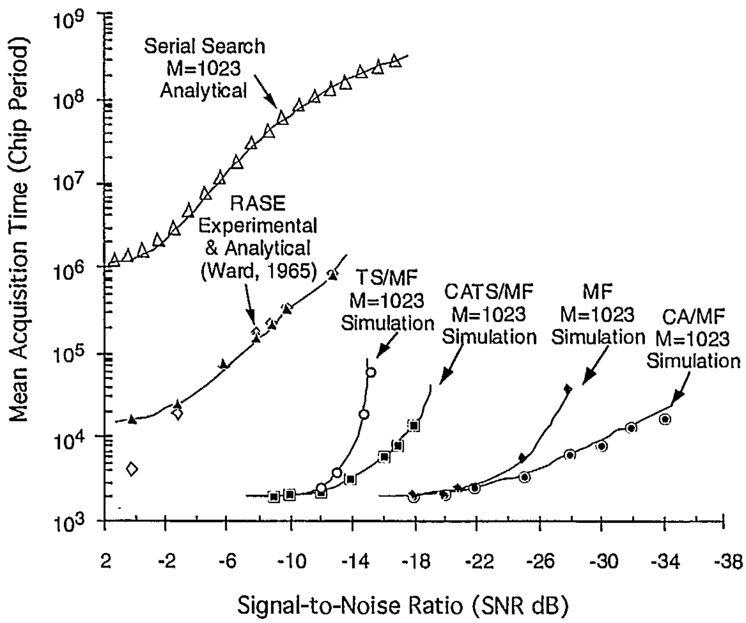

Fig. 22 Performance of MF schemes (X. Zhang M Sc. Thesis)

3. Almost-Coherent Acquisition

In the past decade, several almost-coherent spread spectrum code synchronization structures have been reported in the literature [PoWe84], [SiWe86], [MiGeDa85]. Because the incoming carrier phase is unknown, in-phase and quadrature (I-Q) schemes have been used with non-coherent combining and since the incoming carrier frequency may have some Doppler offset. Short correlators or matched filters are used so that there is only small phase rotation over the filter length. The receiver processes for almost-coherent acquisition are illustrated in Fig. 23.

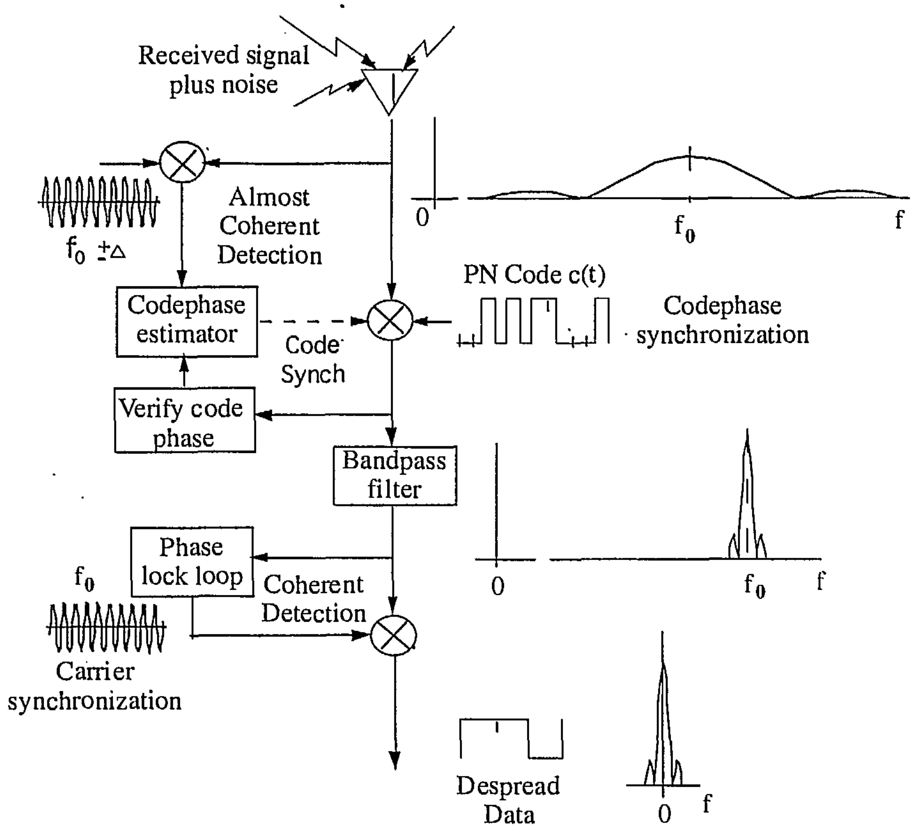

Fig. 23 Almost coherent codephase acquisition

3.1 I-Q correlator (almost coherent)

The performance of non-coherent I-Q correlators in spread spectrum code synchronization has been reported by Rick and Milstein at ICC'94 [RiMi94]. In-phase and quadrature (I-Q) almost coherent demodulation is followed by I-Q correlators and square law combining. For speed, a bank of I-Q correlators is used, one for each possible codephase in the uncertainty region. Analysis has been performed with assumed chip synchronization and for correlation time T_d ranging from 125 to 500 chip periods. Work on non-coherent I-Q correlators has also been presented by [Cheng88]. The I-Q correlator appears to have no advantage over the BPF in a serial search scheme.

3.2 I-Q matched filter (almost-coherent)

The performance evaluation of a non-coherent matched filter in spread spectrum code synchronization was reported by Polydoros and Weber [PoWe84]. An in-phase and quadrature (I-Q) MF scheme was presented which uses almost coherent demodulation and where the code despreading occurs at baseband. This implementation is suitable for SAW or CCD transversal matched filters which retain a received signal history of length MT_c (a short section of the PN code sequence). It is suggested that M>100 and this will provide sufficiently high SNR at the matched filter output for each decision. An advantage of this technique is that out of code phase cells are rejected at the rate at which they arrive. The filter is operated on a fractional chip interval ΔT_c where Δ = 1/2 or 1/4, etc. Their analysis assumes that no data transitions are present during the acquisition period. Excessive frequency offset caused by Doppler shifts can result in serious degradation when the offset approaches $0.25/MT_c$. The system performance rapidly deteriorates at a predetection SNR below a critical point (-21 dB). A verification mode is described which also makes use of the matched filter. Two years later, an improved system was proposed and the analysis included the effect of data modulation and a single tone jammer [SiWe86].

L. B. Milstein et al. presented a rapid acquisition scheme using a bank of surface acoustic wave (SAW) convolvers in 1985 [MiGeDa85]. Each SAW device is a matched filter, matched to a short section of the PN code sequence. The scheme simultaneously processes different subsequences of a longer spreading sequence in a bank of N transversal SAW convolvers. Each of N convolvers spans the duration 2T seconds, where $T = MT_c$, and has one of the subsequences of length M as reference input. The total phase uncertainty to be searched is spanned by the MN phase positions of the N convolvers. The convolver outputs are sampled at chip time intervals (T_c) and then passed through the square law detection. The largest of the resulting MN samples is chosen as the correct phase of the incoming waveform. Since one of the MN samples is the correct sample, it is therefore possible to initially acquire in $2T = 2MT_c$ seconds. The advantage of the scheme is that in 2T seconds, MN phase positions are examined. By comparison, the scheme in [PoWe84] requires MNTc seconds to acquire where MN is the number of chips in the uncertainty region. In a standard serial search technique, examining MN phase positions requires at least MNT_d seconds, where T_d is the dwell time at each tested code phase.

As an extension of the schemes described in [PoWe84], which uses one I-Q MF only, and in [MiGeDa85] which employs a bank of SAW convolvers, E. A. Sourour and S. C. Gupta proposed a DS-SS acquisition scheme that utilizes a bank of N parallel I-Q non-coherent matched filters [SoGu90, SoGu92]. Its mean acquisition time performance is analyzed in both nonselective and frequency selective Rayleigh fading environments. The effects of data modulation and code Doppler were not considered. When compared with a single MF serial scheme in a nonfading channel, the multiple MF parallel system is shown to have faster acquisition equivalent to an additional 2.5 dB SNR tolerance. In the Rayleigh fading channel, the multiple MF parallel system has shorter dwell time and can elude some of the fading. The improvement over the single MF system is as much as 4 dB.

In both of the above schemes, the length of the matched filters or the SAW convolvers must be less than a data bit duration. They could be classed as short MF schemes. The lowest practical frequency range for a SAW convolver (about 30 MHz) may be not suitable for some applications.

3.3 Segmented matched filter

Short matched filter schemes require that the demodulation be almost coherent over the time period of the delay line and can therefore tolerate only minor Doppler shifts in the carrier frequency. To address larger carrier frequency shift, a "segmented matched filter" scheme as shown in Fig. 24 is proposed. The shift-registers are functionally segmented and the partial sums are added according to "Doppler masks" which have binary coefficients (-1, +1) or ternary coefficients (-1, 0 , +1). It is assumed that the length of the segmented matched filter is less than the interval between the data transitions.

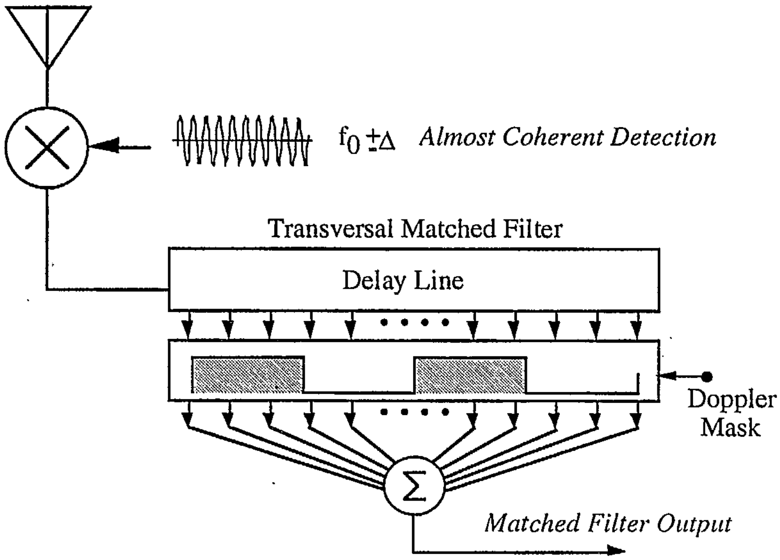

Fig. 24 Segmented matched filter

The output of MF is the sum of each segment weighted by this Doppler coefficient. Several summations can be made simultaneously each with a mask corresponding to 1 kHz, 2 kHz, 3 kHz, ··· Doppler offset frequencies. The codephase is then searched simultaneously at several potential Doppler frequencies. Several associate threshold circuits are used to detect the correct codephase. The largest MF output at the correct codephase gives an estimate of the Doppler offset frequency. This could be used to adjust the local carrier oscillator for data demodulation.

Fig. 25 shows the effects of "nearly coherent" carrier demodulation. Each signal is represented by 501 samples from a Matlab simulation. Except as shown, all magnitudes vary between +1 and -1. Fig. 25(a) represents the sine of the carrier phase offset resulting from motion of the receiver or transmitter (Doppler shift) or from a frequency error. The spreading sequence is shown in Fig. 25(b). It is 100 chips in length. Fig. 25(c) shows the resulting signal after the attempted demodulation. Note that the last half of the signal is inverted with respect to the spreading code. This corresponds to the negative half cycle of the carrier phase offset. If this signal is multiplied with the correct spreading sequence phase, the signal of Fig. 25(d) is obtained. Fig. 25(d) may also be interpreted as a spatial wave to represent the tap output voltages of a transversal matched filter and if the tap output voltages are summed, the result is zero. When a mask (Fig. 25(e)) that is in alignment with the carrier phase offset is applied to the signal, sections of the demodulated signal are inverted to be in code phase synchronization with the spreading sequence, as shown in Fig. 25(f). The despread masked signal shown in Fig.25(g) has an average value of $2/\pi = 0.637$. This may be compared to a coherently demodulated and despread signal which has an average value of 1.00.

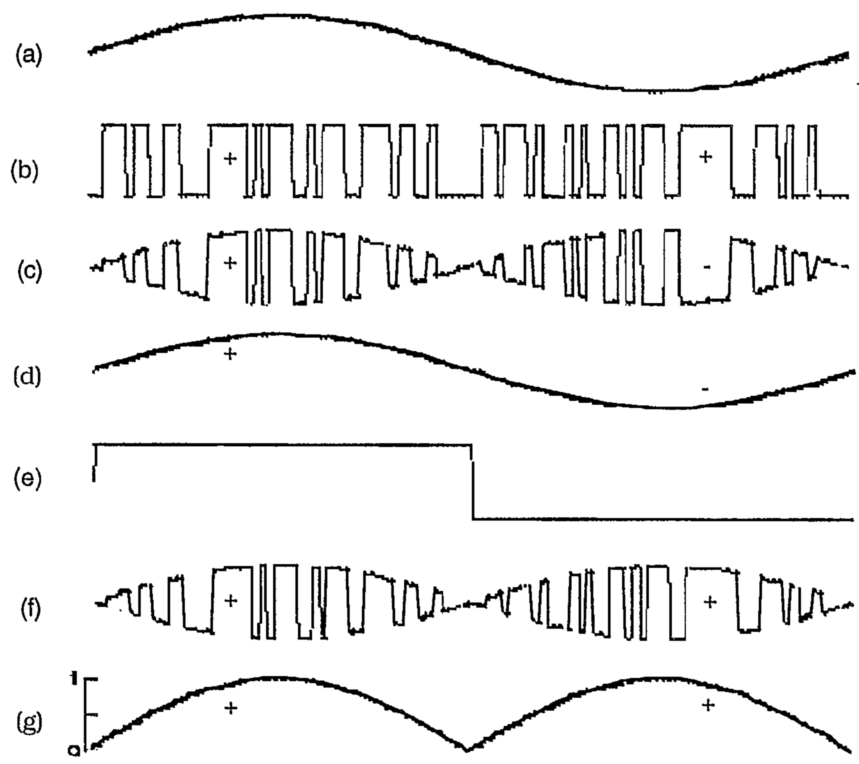

Fig. 25 Illustration of effects of "nearly coherent" carrier demodulation and conventional vs. segmented matched filter components]

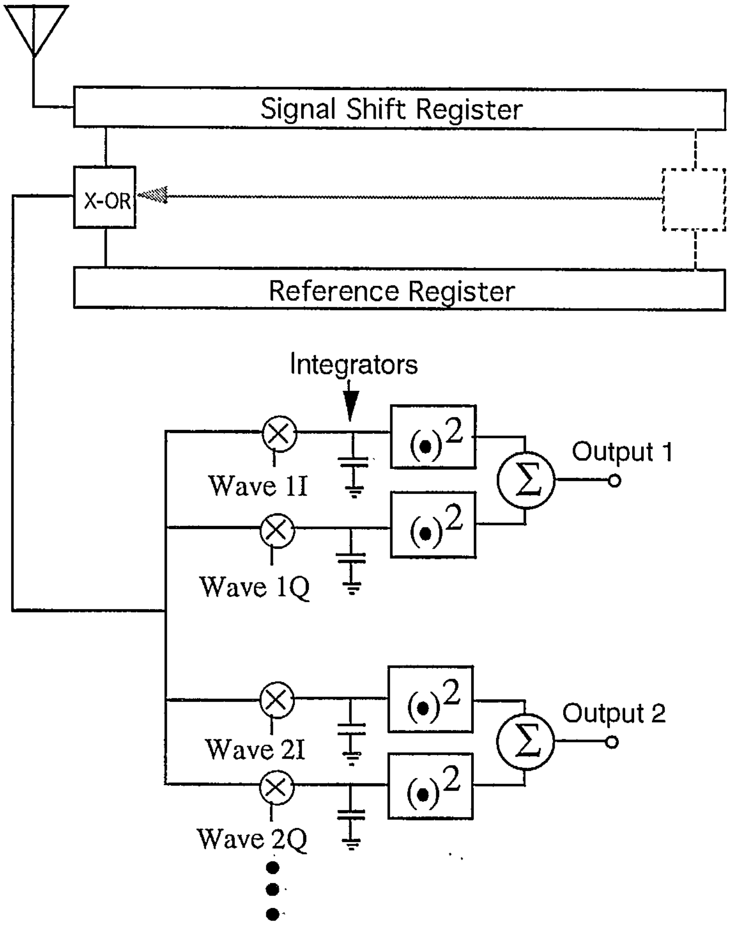

Fig. 26 Implementation of Doppler Compensation

Demodulation is performed with an intended offset frequency which is much lower than the chip rate but somewhat greater than 1/MTc so that there are several "twist" cycles over the length of the matched filter. The I-Q detection is formed by serially "playing back" to sum the tap outputs of the matched filter. The serial summing signal is multiplied by I and Q Doppler

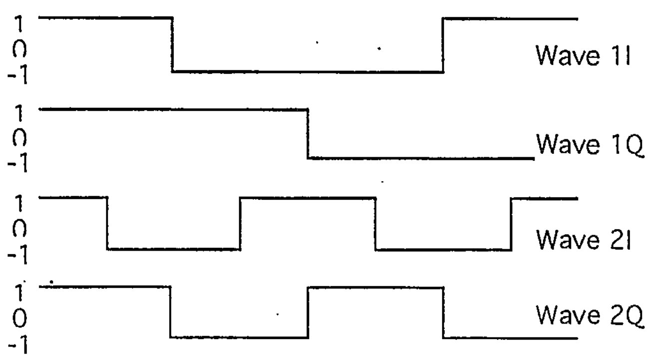

Fig. 27 Doppler mask waveforms (1 kHz & 2 kHz).

masks to generate I and Q outputs for a range of incremental frequencies. This effectively demodulates toward baseband where the inphase and quadrature outputs are non-coherently combined to provide outputs corresponding to 1 kHz, 2 kHz $\cdots$ offset frequencies.

The output for 2 kHz has been calculated as a function of codephase and Doppler offset frequency and the result is shown in Fig. 28. Ten different phases of the offset frequency were used. Calculations were made without added noise. Fig. 29 shows the 2 kHz output at the correct codephase while Fig. 30 shows four outputs over a range of input offset frequencies.

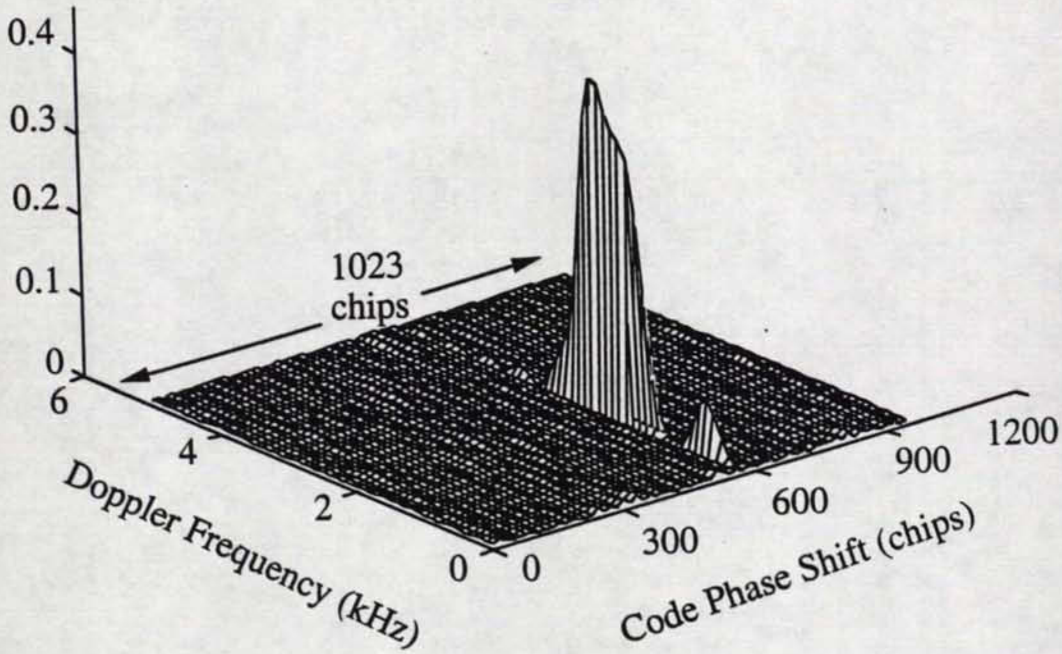

Fig. 28 MF Output 2 vs. Doppler frequency and codephase (M = 1023)

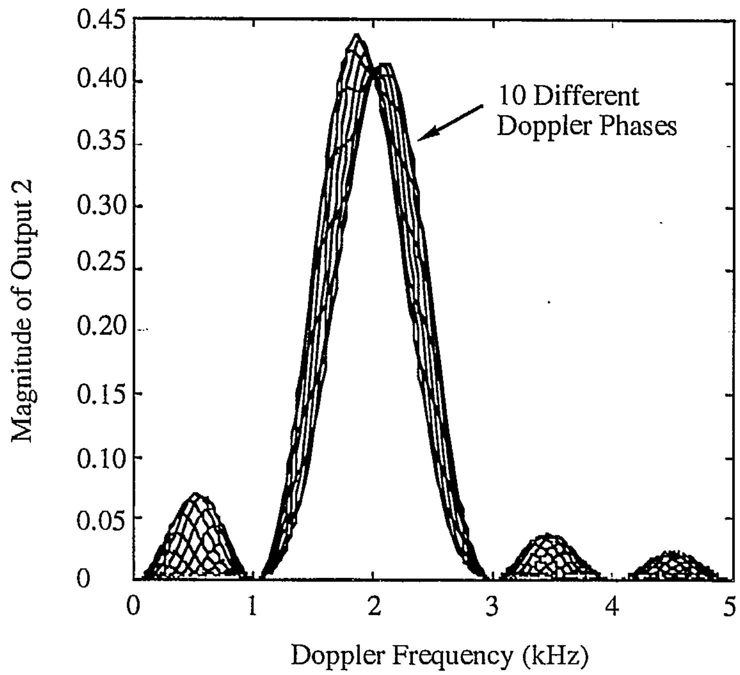

Fig. 28 MF Output 2 vs. Doppler frequency

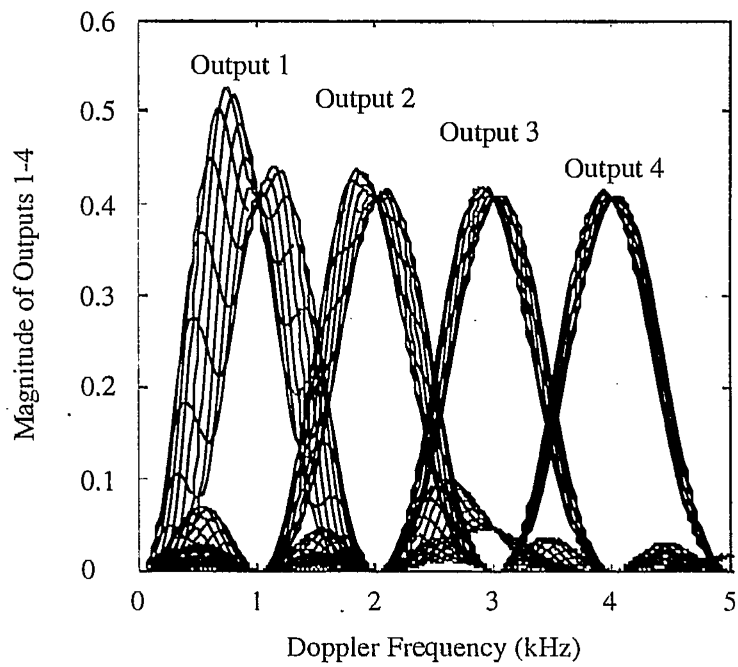

Fig. 29 MF outputs vs. Doppler frequency

4. Tracking

Once correct code phase has been acquired, the tracking loop takes over in supporting the acquisition process. It continually adjusts the locally generated PN code phase and frequency using a feedback loop to maximize correlation. Therefore, the optimum design of tracking loop is also an essential component in the DS-SS receiver design.

There are two common techniques for tracking the coded incoming signal. One is the delay-locked loop, the other is the tau-dither loop. Both of these configurations can be operated in a coherent or non-coherent mode depending on the system application.

4.1 Delay-locked loop

In the delay-locked loop as shown in Fig. 29, the incoming signal is cross-correlated with advanced and retarded versions of the local PN code in two separate correlators. The delay between these two local reference signals is one or two chips. The results of these cross-correlation operations are detected by square-law envelope circuits and then subtracted to produce a discriminator characteristic. This difference output is applied to a loop filter and used to control the receiver's voltage controlled oscillator (VCO) for closing the loop. The receiver's code will track the incoming code at a point halfway between the two correlation maxims.

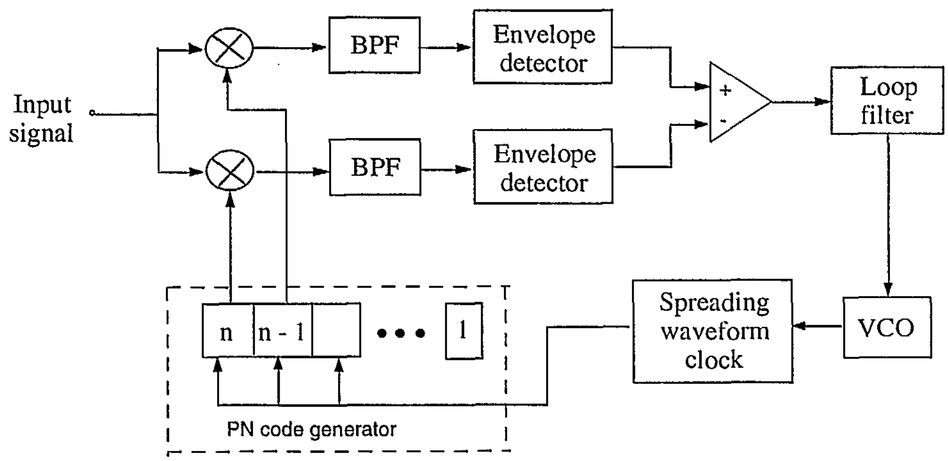

Fig. 29 Delay-locked loop

4.2 Tau-dither loop

Tau-dither tracking in spread spectrum systems also makes use of the triangular code correlation function. The typical tau-dither tracking loop is illustrated in Fig. 30. In tau-dither loop a phase modulator is used to drive the receiver's code reference back and forth alternately between shifted and unshifted positions. The received signal is alternately correlated with the advanced and retarded versions of the locally generated PN code. As the clock phase is shifted back and forth, the correlator output is amplitude modulated at the phase shifting rate. This amplitude modulated signal is then detected by the envelope detector and alternately inverted by the binary signal to develop a difference signal which drives the voltage controlled oscillator (VCO) through the loop filter.

One major advantage of the tau-dither loop over the delay-locked loop is that only a single input correlator is required, thus eliminating problems of gain imbalance and other mismatches that are present in a two-channel loop.

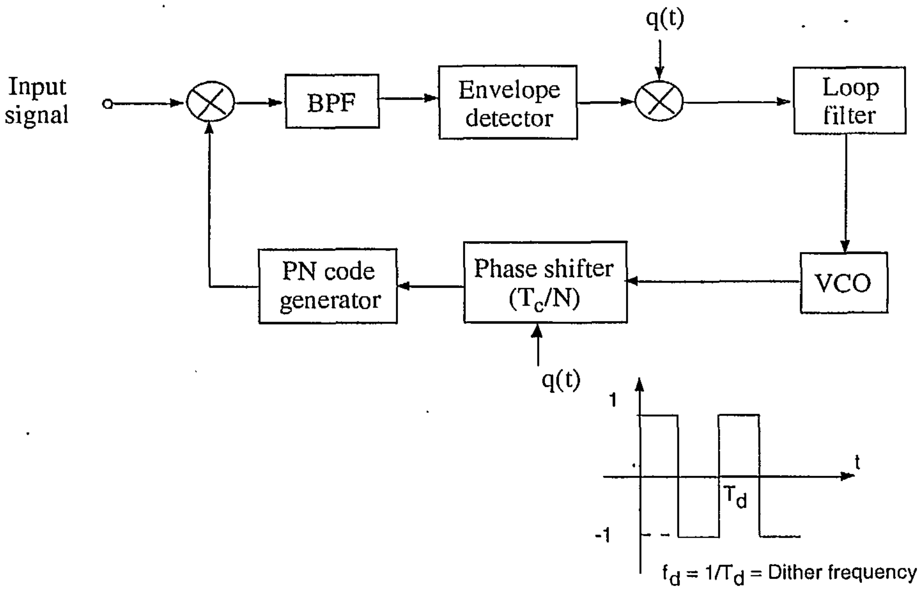

Fig. 30 Tau-dither loop

5. CDMA for Personal Communications Satellite

There has been recent interest in satellite based personal communication systems and one of the transmission methods under consideration is code division multiple access (CDMA) using direct sequence spread spectrum (DS/SS) modulation

For a geostationary system, four to six satellites are usually considered; for an intermediate circular orbit (ICO), 10 to 20 satellites are appropriate; and for a low earth orbit (LEO) system, 40 to 70 satellites have been suggested. Each satellite would have several spotbeams and the isolation between spotbeams will in general not be as great as the isolation between adjacent cells in a terrestrial person communication system. In addition, there will often be more than one satellite illuminating the same area, partly to ensure coverage and partly to provide diversity. This is particularly true for satellites in non-geostationary orbits. A large number of codes must therefore be used to distinguish between different satellites as well as different spotbeams.

5.1 PN code length and number of users

There are a number of factors which affect the length of the spreading sequence in a CDMA system. The PN sequence period is sometimes chosen to match the symbol period. The length of the sequence thus defines the spreading gain which, in turn, determines the number of simultaneous users that the system can support. In larger systems the choice of spreading sequence length is dictated by other considerations and the PN sequence period usually exceeds the information bit period.

As a reference example, we consider the EIA IS-95 terrestrial system initially proposed by Qualcomm. Each base station emits a pilot channel which is simply modulated by the basic spreading sequence and the mobile station uses this channel first for initial frequency and code synchronization and then for synchronization tracking. The amplitude of this pilot channel is also used for determining when to handover between base stations. In this system, the same PN code is used by each base station, however, each base station is distinguished by a different code offset (time delay). This means the mobile can search for a single code during the synchronization phase. The strongest signal identified corresponds to the code phase of the best cell site. A code of length 2^{15} is used and each base station is offset by a multiple of 2^6 or 64 chips, thus, there are 512 such offsets. At a chip rate of 1.22 Mchip/sec, each offset is equivalent to a transmission delay of 50 ms which corresponds to a distance of 15 kilometers. Assuming base stations are located closer than 15 kilometers apart, this provides isolation between the different base stations. Details of synchronization in EIA/IS-95 CDMA system are given in Appendix A.

Assuming up to 100 effective co-users per cell and a required $E_b/N_0' > 7$ dB a suggested spreading factor is 512 (spreading gain = 27 dB). If the required user data rate is 10 kb/s, the system chip rate is approximately 5 Mchip/s. This is a factor of 4 faster than the chip rates used in simulations and analyses presented elsewhere in this report and thus the acquisition times would be correspondingly shorter.

In a cellular system, a CDMA cell is surrounded by other CDMA cells and there is considerable interference from users in these adjacent cells. This additional co-user interference decreases the number of users which can be supported by one cell. and the number of actual users is less For a design with 100 effective co-users, the actual number of users per cell may be in the order of 50. This number is consistent with the original 20 MHz AMPS system in which there were up to 45 active users and 3 signaling channels per cell.

5.2 CDMA synchronization aspects for LEOS systems

In contrast to terrestrial mobile systems where the delay to adjacent base stations can vary up to 50 ms, satellite systems have relative signal to adjacent satellites that can range from 10 to 20 ms. This makes it difficult to use offsets of the same PN code as implemented in the IS-95 system.

An alternative approach would be to assign each base station a separate spreading code. Under these conditions, presumably a shorter spreading code could be used, for example, codes of length between 2^{10} and 2^{13} may be suitable. However, for initial synchronization, the mobile would have to search through a number of codes to determine which was the strongest received code. In practice, one might expect that 10-20 codes may be sufficient, however, if one provided the same isolation as in the IS-95 system then there would be 512 such codes to search.

An example cell pattern for a LEOS personal communication system is illustrated in Fig. 31. Each satellite has several spot beams which would emanate from separate antennas with i) different carrier frequencies, ii) different spreading codes or iii) different phases of the same spreading code. All three multiple access methods are possible and it would be logical to select option iii) since all signals arrive with approximately the same delay and this represents the most simple implementation. Multipath propagation from different antennas to the same receiver in a spot beam overlap area would be resolved by despreading only one of the received signals using the correct delay of the PN sequence assigned to that satellite. If 50 us differential delay is assumed between the spot beam signals, then 6 or 7 spot beams would require in excess of 350 us. This could be accommodated with a code length of 2^{11} at 5 Mchip/s (LT_c = 400 us). For a spot beam cell with 50 active users, users could be assigned a codephases which are each offset by 4 chips (0.8 ms). This would occupy 40 ms of the PN code cycle and would leave 360 ms guard time to the adjacent spot beam codephases.

In the overlap coverage area of two or three satellites, it is not appropriate to use different phases of the same code as this would require an interval of (20 ms)(5 Mchip/s) = 100,000 chips between each of 10-20 satellites. In the extreme case, the receiver would have to search through 2,000,000 codephases over a range of Doppler frequency offsets and the acquisition time would be much too long for mobile telephone service. Methods of using i) different frequencies or ii) different codes are appropriate and would require the shorter length codes (2047 chips) to be searched over a variety of carrier frequencies or spreading codes. It should be noted that the CDMA approach provides natural and gradual frequency reuse and that the separate frequency approach would be wasteful of bandwidth.

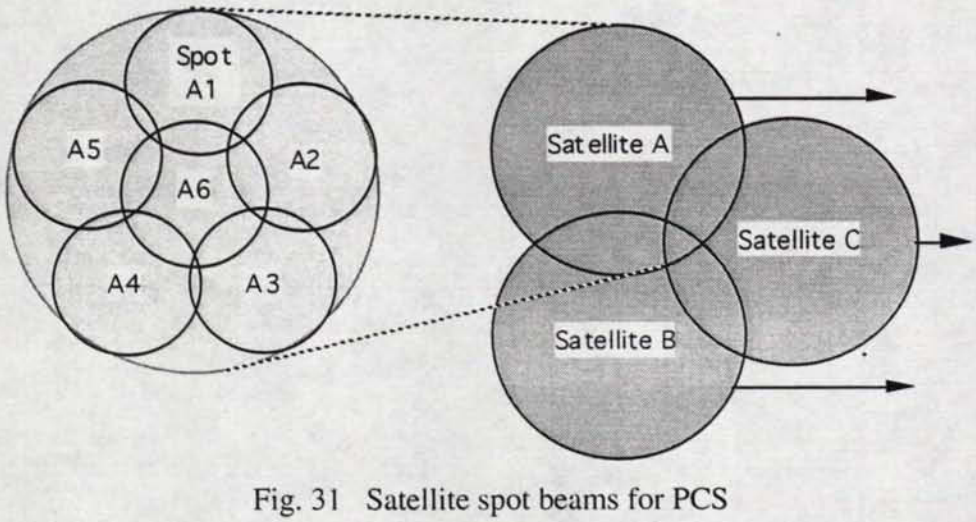

Fig. 31 Satellite spot beams for PCS

Rapid synchronization is essential for an effective personal communication system. The receiver must achieve synchronization with the transmitter RF carrier, the spreading pseudonoise codephase and also the timing of the data bits before communication can take place. Synchronization should take place in less than one second and with certainty in less than 5 seconds. Some of the current acquisition methods would not be suitable for the spreading scheme proposed here for PCS satellites. The following discussion will assume 100 effective co-users and $E_c/N_0' = -20$ dB.

The direct subsequence estimation methods are unsuitable since they require a high SNR, in the order of 0 dB, so that the shift register can be loaded with a reliable value. A CDMA system has a low SNR even with a modest number of co-users.

Correlator serial search with coherent demodulation can have an excessive synchronization time if no a priori information is available. For search with random starting location, a code length of 2^{11} a chip rate of 1.22 Mchip/s and dwell time of $450T_c$, the search time is shown in Fig 14 to be approximately 1.5 seconds at -20 dB E_c/N_0'.

Non-coherent demodulation must be used on the reverse channel (or uplink) in a spread spectrum system. When part of the integration occurs in the predetection BPF and part occurs in the postdetection integrator, the portion after the detector is relatively less effective. As developed in Section 2.2, the performance relative to the coherent correlator, is degraded by the factor $\sqrt{M2}$ and assuming this to be $\sqrt{455}$ as in [Hopins77] then T_d would have to be increased by a factor of 21 to have the same flow graph probabilities. The search time then increases to 32 seconds at 1.22 Mchip/s. The predetection filter would have bandwidth B = 50 kHz which allows for at least ±10 kHz Doppler shift or frequency error. The bandpass filter can be replaced by the I-Q correlators noted in Sect. 3.1, however, there appears to be no advantage in doing so.

Matched filter search must use I-Q matched filters if the received signal is not assured to be in phase. In the case of almost-coherent demodulation, only small frequency offsets can be tolerated so that there is less than 1/4 cycle of "twist" or rotation over the length of the matched filter. For a system operating at 1.22 Mchip/sec at $E_c/N_0' = -20$ dB the optimum matched filter length is M=2047 which gives an acquisition time of 1 ms for coherent demodulation. For I-Q filters, the noise power is effectively doubled (see Equ. 21) which reduces the effective MF length by 2 and increases the acquisition time to 4 ms. Unfortunately the Doppler shift tolerance of the filter with length M = 2047 would only be ±0.25(1.22 MHz)/M = ±150 Hz which is insufficient for a LEOS system with Doppler shifts up to 10 kHz.

A solution to this problem is to use successive attempts to synchronize each with a different local oscillator shift in the receiver. To cover ±10 kHz, a total of 120 overlapping frequency bins is suggested. Sequential searching in up to 120 "Doppler bins" does not add greatly to the hardware complexity, however, it will be required to dwell 8 ms in an average of 60 bins. This will increase the search time from 1 ms to 480 ms.

A second solution to this Doppler problem is to use a shorter I-Q MF length which will adversely affect P_{fa} and will increase the acquisition time. For E_c/N_0 = -20 dB , sequence length L = 2^{11} and matched filter length M = 30 (effective length 15), the Doppler tolerance is ±10 kHz but the acquisition time increases from 4 ms to 4 s.

It is also possible to use an intermediate length I-Q MF with length, for example, M = 128 and to use a method of simultaneous Doppler bin search. A method, which was introduced in Sect. 3.3, offers a single matched filter which has several outputs over a range of incremental frequencies. For length M = 256 (same complexity as I-Q 128) and chip rate 1.22 Mchip/s, the Doppler tolerance is 1.2 kHz and a total of 16 overlapping frequency bins is suggested. The correct output has half the signal power and twice the noise power when compared with a coherent MF. The effective length is thus reduced to 64 and the expected acquisition time is 900 ms. A table for comparison of acquisition time is given below.

Coherent Correlator	narrowband	1.5 s
Non-coherent correlator	wideband	32 s
Coherent MF	narrowband	1 ms
2047 chip I-Q MF	120 bin serial Doppler	480 ms
30 chip I-Q MF	wideband	4 s
128 chip I-Q MF	8 bin parallel Doppler	900 ms
256 chip I only MF	16 bin parallel Doppler	900 ms

Table 2 Acquisition time comparison (L=2047, R=1.22 Mchip/s, E_c/N_0 = -20 dB)

6. Conclusions and Recommendations

6.1 Recommendation for signal structure

Results from this study suggest a signal structure which uses different CDMA codes to differentiate each satellite (as in the GPS system) and different phases of one code (as in the IS-95 standard) to differentiate spot beams and to differentiate different users within each spot beam.

For a user data rate of 10 kb/s, 100 effective co-users and a despread $E_b/N_0 = +10$ dB, a spreading gain of 1000 is required and the chip rate becomes 10 Mchip/s. For a code period of 400 us to provide 50 us isolation between spot beams, a code period of 4095 chips is required. To facilitate use of (almost) coherent demodulation and a matched filter in the receiver, the data symbol period should be an integer multiple or fraction of the code period and the symbol boundaries should be fixed to a known phase in the code sequence. A factor of 2 in co-user noise can be used to allow for spotbeam and satellite overlap and this yields a capacity of 50 active users. This capacity is consistent with the original 20 MHz AMPS system in which there were 45 users per cell and 3 signaling channels.

In each spotbeam, a pilot sequence is suggested in which a portion (e.g.. the first 1/4) is transmitted without data modulation, the next 1/4 is used for signaling to users within that spotbeam and the remaining 1/2 could be used for low rate data services. Reception of the pilot/signaling sequence is critical to successful operation and this could be transmitted at 2 times or 4 times the power of the co-user signals. This would improve the effective E_c/N_0' from -20 dB to -14 dB which would then decrease the synchronization time and improve the reliability of the signaling and data channels. The penalty would be a loss of 1 - 3 user channels.

6.2 Recommendation for receiver structure

There are 5 non-coherent acquisition methods listed in the last section which are appropriate for a LEOS spread spectrum PCS. The following discussion will assume $E_c/N_0' =$ -20 dB and the nominal chip rate of 1.22 Mchip/s which was adopted for the study. This gives a pessimistic evaluation. Operation with 5 Mchip/s rate would decrease the acquisition time by a factor of 4 and using a pilot with +6 dB power relative to the users could also decrease acquisition time by a factor of 5 - 10.

The traditional serial search non-coherent correlator can be brought into the 5 second acquisition time specification by using 8 parallel "hunters" for the correct codephase. This would reduce the time from 32 s to 4 sec. Integrated circuits available from Qualcomm have 4 simultaneous hunters. This technique is not sensitive to data transitions and naturally accommodates Doppler offset. A pilot channel is not required.

The wideband I-Q matched filter with length M=30 covers the required Doppler frequency range at 1.22 Mchip/s and has an acquisition time within the 5 second requirement. A chip from Stanford Telecom (described in Appendix B) can perform the required MF function. This technique does not accommodate data transitions so boundaries of the data bits should coincide with boundaries of the code cycle. Some frequency offset is acceptable. A pilot channel is not required but would be convenient.

Methods which simultaneously search codephase and offset frequency can give somewhat improved performance but at the price of increased complexity. Work is already in progress on the I only MF as part of an M. Sc. project. It may be possible to develop an efficient (single chip) analog processing structure for the 128 I-Q MF listed in the previous section and this is proposed as a topic for further investigation. Another topic for further investigation is the possible use of the shift and add properties of M sequences in differential detection MF synchronization.

Appendix A

APPENDIX A Synchronization Summary for TIA/EIA/IS-95

Personal Cellular Networks

Similar to the computer and telecommunication networks, the personal cellular network (PCN) based on CDMA system has been structured and layered by the Telecommunications Industry Association. The layering structure is divided into the physical, link, and control process layers as shown in Fig. 1. This gives a simplified logical view of the CDMA protocol structure for the PCN. The CDMA systems is actually composed of four channels: Pilot/Sync Channel, Paging /Access Channel, Forward Traffic Channel and Reverse Traffic Channel. The protocol divides the system function into conceptual layers.

- Layer 1 is the physical layer of the digital radio channel which describes the functions of bit transmission such as modulation, coding, framing and channelization via radio waves.
- Multiplex sublayer is between layer 1 and layer 2. It contains the multiplexing functions that allow sharing of the digital radio channel for user data and signaling processes.
- Layer 2 includes Signaling Layer, Paging/Access Channel Layer, Sync Channel Layer. The protocol is associated with the reliable delivery of signaling between the base station and the mobile station, such as message retransmission and duplicate detection.
- Control Process Layer 3 is the protocol associated with call processing, radio channel control, and mobile station control, including: call setup, handoff, power control, and mobile station lockout.

These layers are combined to provide the basic mobile station services of call setup and tear down, power control, handoff, maintenance, authentication, and registration. Many service options at end-user application are provided in the interim standard IS-95. These optional services may have an entirely different set of upper layers for a different user application, but they can be viewed as plugging into sockets provided by the multiplex sublayer. For example, Multiplex Option 1 allows both primary and secondary traffic simultaneously to be active and plug into the multiplex sublayer. These service options can be implemented in common software by the manufacturer for the user selection.

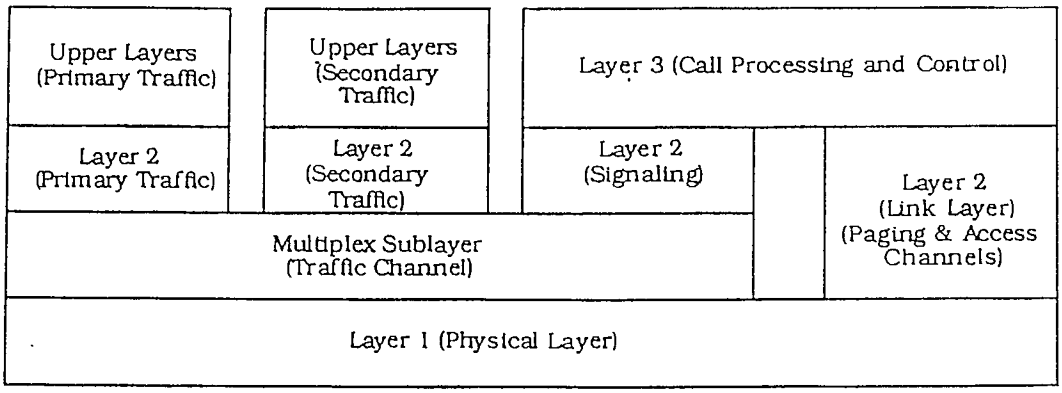

Fig. A-1 Mobile Station and base station layers [from TIA/EIA/IS-95]

Pilot Carrier Synchronization (forward transmission):

One of the most important applications for the CDMA technique is the direct- sequence CDMA used for digital cellular radio communications and advanced wireless technologies. The CDMA technique allows a spread spectrum radio network on top of existing users occupying the same radio frequency (RF) band. The CDMA approach can solve the near-term capacity concerns of frequency band and the long-term need for an economic, efficient and truly portable communications. In the CDMA cellular telephone system, each cell site transmits a pilot carrier signal. This pilot carrier is used by the mobile station to obtain initial system synchronization and to provide robust time, frequency and phase tracking of the signal from the cell site. The pilot carriers are transmitted by each cell site using the same code but different spread spectrum code phase offsets. The mobile station distinguishes them by their particular phase offset. Using the same code by all pilots allows the mobile station to find system timing synchronization by a single search through all code phases. The strongest signal identified corresponds to the code phase of the best cell site.

Each cell also transmits a setup or sync channel. This channel uses the same PN sequence and phase offset as the pilot channel and can be demodulated whenever the pilot channel is being tracked. This sync channel carries cell site identification, pilot transmit power, and the cell site pilot PN carrier phase offset. With this information, the mobile station is capable of establishing System Time and knows the proper transmit power to initiate calls.

CDMA System Time:

The start of CDMA System Time is January 6, 1980 00:00:00: UTC, which coincides with the start of GPS time. The long code and the zero offset pilot PN sequences for the I and Q channels have their initial states aligned to the System Time. The alignment of the initial states of the long code and the pilot PN sequence does not occur again for more than 37 centuries. The System Time at various points in the transmission and the reception processes is the absolute time referenced at the base station antenna offset by the one-way delay (5 μs/mile or 6 chips/mile) or round-trip delay of the transmission. Time measurements are referenced to the transmit and receive antennas of the base station and the RF connector of the mobile station.

Timing Reference Source

Each station shall use a time base reference which is time-aligned to CDMA System Time. All critical CDMA transmissions, including pilot PN sequences, frames, and Walsh functions shall be derived from the time base reference. Reliable external means should be provided at each base station to synchronize each base station's time base reference to CDMA System Time.

All base stations should radiate the pilot PN sequence within ±3 μs of CDMA System Time. The maximum tolerance is ±10 μs. All CDMA Channels radiated by a base station shall be within 1 ms of each other.

Pilot PN Code and Long Code

Pilot PN Sequence and Initial State

After the direct sequence spreading, the signals are spread by the zero-offset I and Q pilot PN sequences. The pilot PN sequences are a pair of modified m-sequences with a period of 2^{15} chips and are based on the following characteristic polynomials, respectively:

For the in-phase (I) sequence:

$$P_I(x) = x^{15} + x^{13} + x^9 + x^8 + x^7 + x^5 + 1$$

For the quadrature-phase (Q) sequence:

$$P_Q(x) = x^{15} + x^{12} + x^{11} + x^{10} + x^6 + x^5 + x^4 + x^3 + 1$$

The I and Q pilot PN sequences based on $P_I(x)$ and $P_Q(x)$ can be generated by the maximum length linear feedback shift register structure. In order to obtain the I and Q pilot PN sequences of period 2^{15}, a '0' is inserted in the generated sequence after 14 consecutive '0' outputs which occurs only once in each period. Therefore, the pilot PN sequences have one run of 15 consecutive '0' outputs instead of 14. The chip rate of pilot PN sequence is 1.2288 MHz, therefore the pilot PN sequences repeat every 26.667 ($2^{15}/1228800$ seconds). There are exactly 75 repetitions in every 2 seconds.

The data spread by the Q pilot PN sequence shall be delayed by half a PN chip time (406.901 ns) with respect to the data spread by the I pilot PN sequence.

The initial state of the pilot PN sequence for both I and Q is the first '1' output following 15 consecutive '0' outputs from the pilot PN sequence generator.

Pilot Search

A pilot is associated with the Forward Traffic Channels in the same Forward CDMA Channel. All pilots in a pilot set have the dame CDMA frequency assignment.

Soft handoffs and CDMA to CDMA hard handoffs using the same frequency assignment are typically initiated by the mobile station. The mobile station searches for pilots to detect the presence of CDMA Channels and to measure their strengths. When the mobile station detects a pilot of sufficient strength that is not associated with any of the Forward Traffic Channels assigned to it, it sends a Pilot Strength Measurement Message to the base station. The base station can then assign a Forward Traffic Channel associated with that pilot to the mobile station and direct the mobile station to perform a handoff.

The requirements for pilot search: The base station specifies the search window (range of PN offsets) for each of the following pilot sets: Active Set, Candidate Set, Neighbor Set and Remaining Set. In the search window, the mobile station is to search for usable multipath components (i.e., multipath components that the mobile station can use for demodulation of the associated Forward Traffic Channel) of the pilots in the set.

Search performance criteria are defined in IS-98 "Recommended Minimum Performance Standards for Dual-Mode Wideband Spread Spectrum Cellular Mobile Stations".

The Long Code Generator and Its Initial State

Before transmission, the Traffic Channel and the Access Channel shall be direct sequence spread by the long code. This long code is a maximal-length PN sequence with period of $2^{42} - 1$ chips and is generated from the following characteristic polynomial with linear recursion:

$$p(x) = x^{42} + x^{35} + x^{33} + x^{31} + x^{27} + x^{26} + x^{25} + x^{22} + x^{21} + x^{19} + x^{18} + x^{17} + x^{16} + x^{10} + x^{7} + x^{6} + x^{5} + x^{3} + x^{2} + x^{1} + 1$$

Each PN chip of the long code shall be generated by the modulo-2 inner product of a 42-bit mask and the 42-bit state vector of the sequence generator as shown in Fig. 2. The mask used for the long code varies depending on the channel type on which the mobile station is transmitting. For example, the mask should be set to the Access Channel Long Code Mask when transmitting on the Access Channel, while the Public or Private Long Code Mask is used for the transmission on the Reverse Traffic Channel.

The initial state of the long code is the first '1' following 41 consecutive '0' outputs from the output of the long code generator with the binary mask of '1' in the MSB followed by 41 '0's. Referring to the shift register in Fig. 2, the 42nd shift register contains '1' and the rest shift registers are equal to '0'.

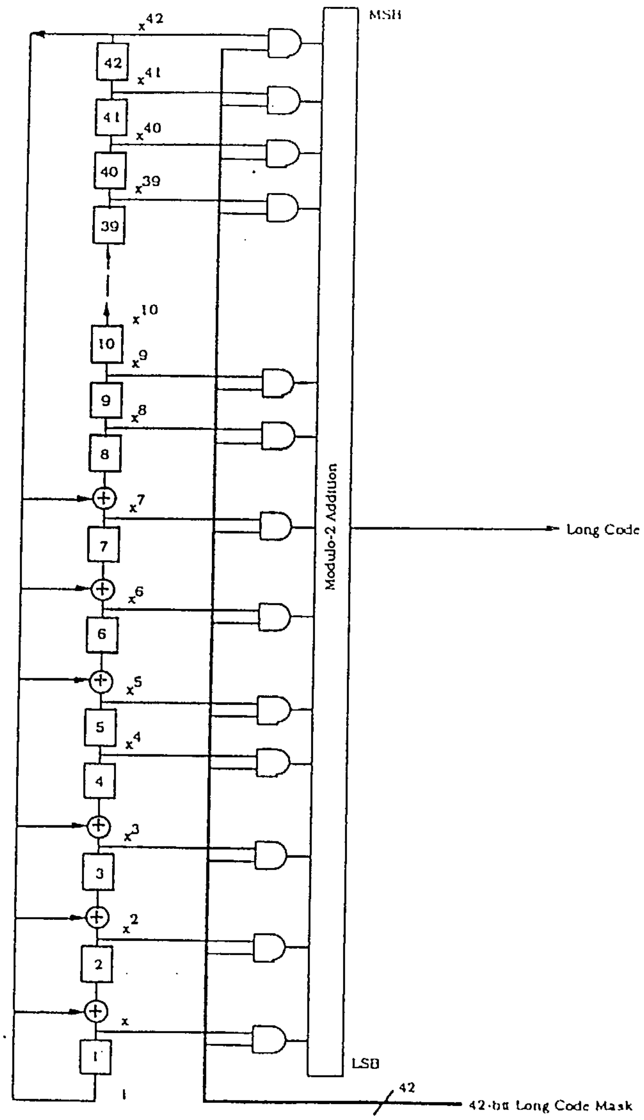

Fig. A-2 Long code generator [from TIA/EIA/IS-95]

Channel Classifications in CDMA System

The CDMA cellular system has been designed with a very flexible signaling and control structure. This is to allow extendibility so that additional features and capabilities can be readily added in the future. When a mobile station is not involved in a call, signaling functions must be provided with the base station. For this purpose, the CDMA system has the Pilot, Sync, Paging, and Access Channels.

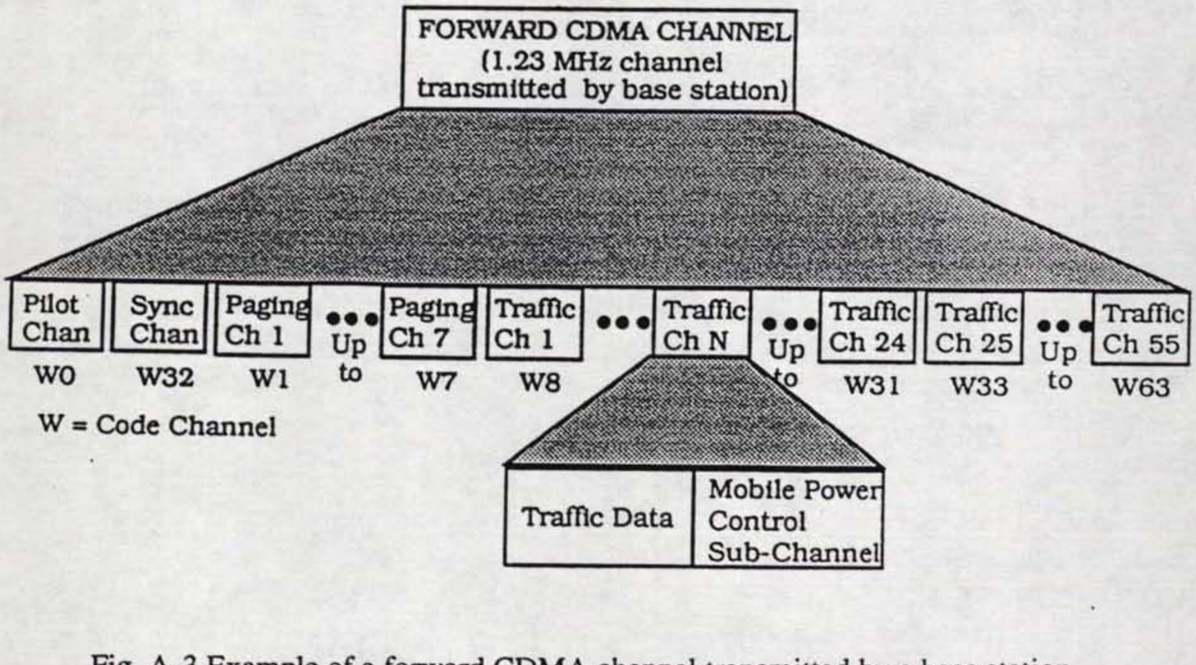

Fig. A-3 Example of a forward CDMA channel transmitted by a base station
[from TIA/EIA/IS-95]

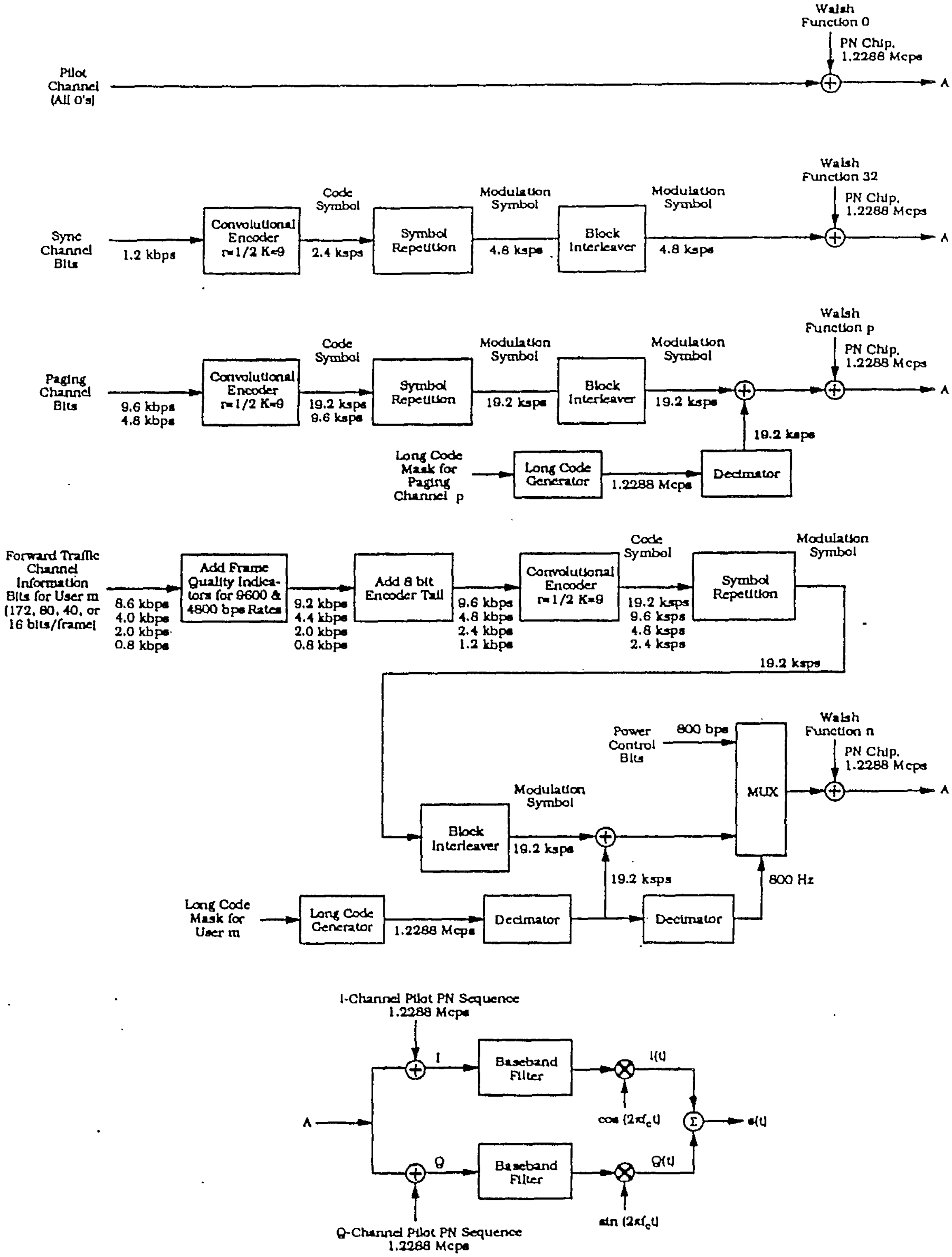

Fig. A-4 Forward CDMA channel structure [from TIA/EIA/IS-95]

Pilot Channel:

The pilot channel is an unmodulated, direct-sequence spread spectrum signal transmitted continuously by the base station on each active Forward CDMA Channel. The pilot channel is used for synchronization by a mobile station operating within the coverage area of the base station. The Pilot Channel allows a mobile station to acquire the timing of the Forward CDMA channel, provides a phase reference for coherent demodulation, and provides a means for signal strength comparisons between base stations for determining when to handoff.

Pilot PN Sequence Offset: Each base station shall use a time offset of the pilot PN sequence to identify a Forward CDMA Channel. Time offsets may be reused within a CDMA cellular system.

Distinct Pilot Channels shall be identified by an offset index (0 through 511 inclusive). This offset index specifies the offset value from the zero offset pilot PN sequence. The start point of the zero offset pilot PN sequence shall be at the beginning of every even second in time, referenced to base station transmission time. The precise start instant of the zero offset pilot PN sequence for either the I or Q sequence is the midpoint between the last '0' of the 15 consecutive '0' outputs and the succeeding '1' of the pilot sequence.

Five hundred twelve unique values are possible for the pilot PN sequence offset. The offset (in the units of 64 chips) for a given pilot PN sequence from the zero shift pilot PN sequence equals the index value multiplied by 64. The pilot PN sequence offset is illustrated in Fig. 3. The same pilot PN sequence offset shall be used on all CDMA frequency assignments for a given base station.

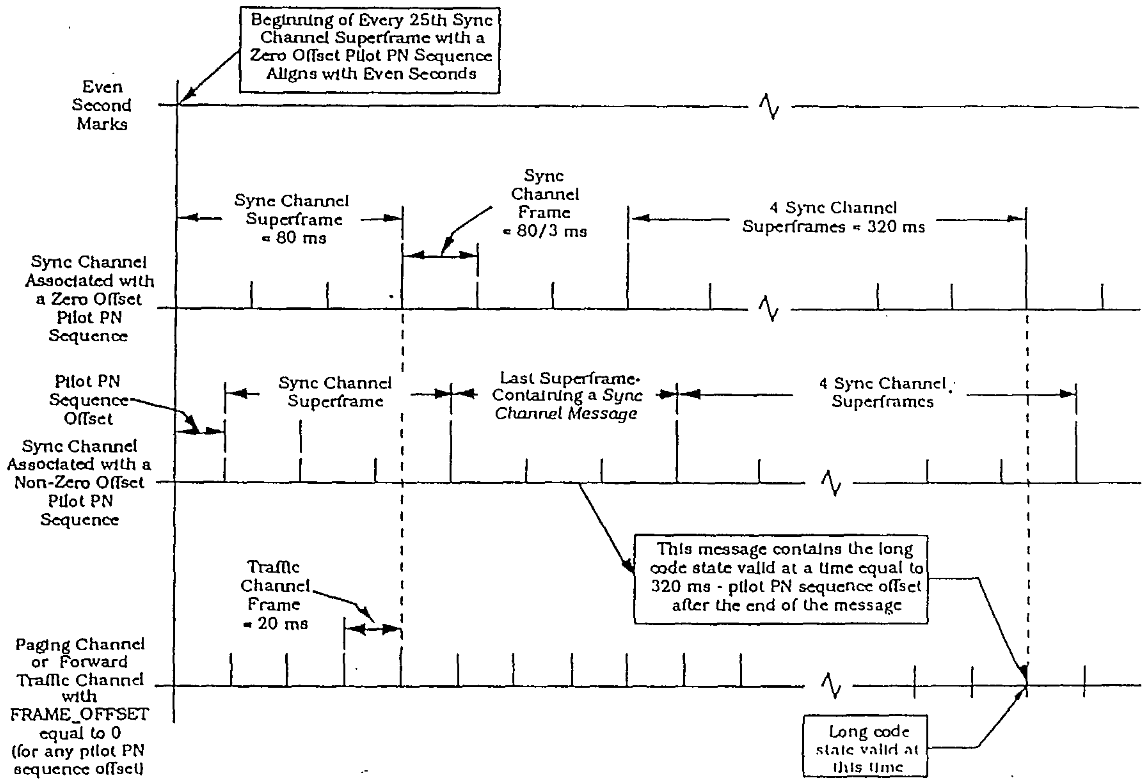

Fig. A-5 Forward CDMA channel pilot PN sequence offset [from TIA/EIA/IS-95]

Sync Channel

This channel is code channel 32 in the Forward CDMA Channel which transports the synchronization message to the mobile station.

The Sync Channel is an encoded, interleaved, spread, and modulated spread spectrum signal that is used by mobile stations operating within the coverage area of the base station to acquire initial time synchronization. The Sync Channel provides the mobile station with system configuration and timing information. The base station shall transmit at most one Sync Channel for each supported CDMA Channel and continually send the Sync Channel Message on each Sync Channel that the base station transmits.

The bit rate for the Sync Channel is 1200 bps. A Sync Channel frame is 26.667 ms (80 ms/3) in duration. The I and Q channel pilot PN sequences for the Sync Channel use the same pilot PN sequence offset as the Pilot Channel for a given base station. Once the mobile station achieves pilot PN sequence synchronization by acquiring the Pilot Channel, the synchronization for the Sync Channel is immediately known. This is because the Sync Channel (and all other channels) are spread with the same pilot PN sequence, and because the frame and interleave timing on the Sync Channel are aligned with the pilot PN sequence.

Paging Channel

The Paging Channel is a code channel in a Forward CDMA Channel used for transmission of control information and pages from a base station to a mobile station.

The paging channel is an encoded, interleaved, spread, and modulated spread spectrum signal that is used by mobile stations operating within the coverage area of the base station. The base station uses Paging Channel to transmit system overhead information and mobile station specific messages. The base station transmits from one to seven Paging Channels on each supported CDMA Channel.

The paging Channel shall transmit information at a fixed data rate of 9600 or 4800 bps. The 2400 and 1200 bps data rates are not supported on the Paging Channel. The Paging Channel frame is 20 ms in duration. Paging Channel slots are 80 ms in duration.

The I and Q channel pilot PN sequences for the Paging Channel use the same pilot PN sequence offset as the Pilot Channel for a given base station.

Access Channel

This is a Reverse CDMA Channel used by mobile stations for communicating to the base station. The Access Channel is used for short signaling message exchanges such as call originations, responses to pages, and registrations. The Access Channel is a slotted random access channel. Each Access Channel is associated with a Paging Channel. Up to 32 Access Channels can be associated with a Paging Channel.

For the Reverse Traffic Channel, this spreading operation involves modulo-2 addition of the data burst randomizer output stream and the long code. For the Accesses Channel, this spreading operation involves modulo-2 addition of the 64-ary orthogonal modulator output stream and the long code.

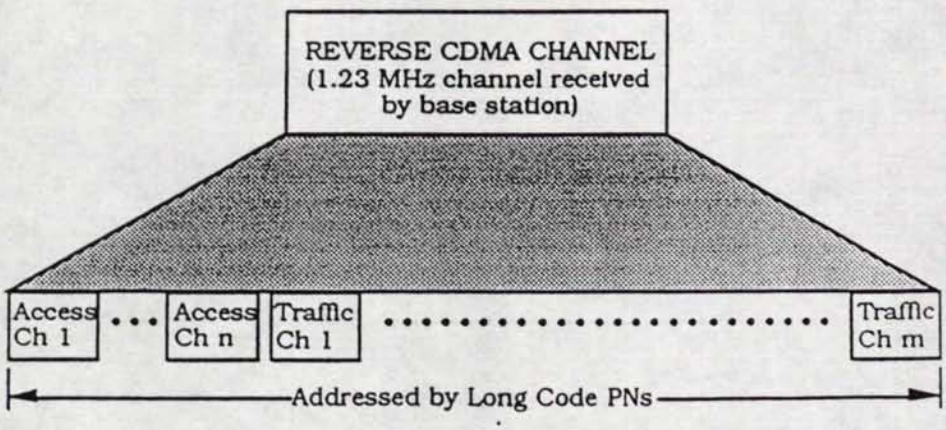

Fig. A-5 Example of logical reverse CDMA channels received at a base station
[from TIA/EIA/IS-95]

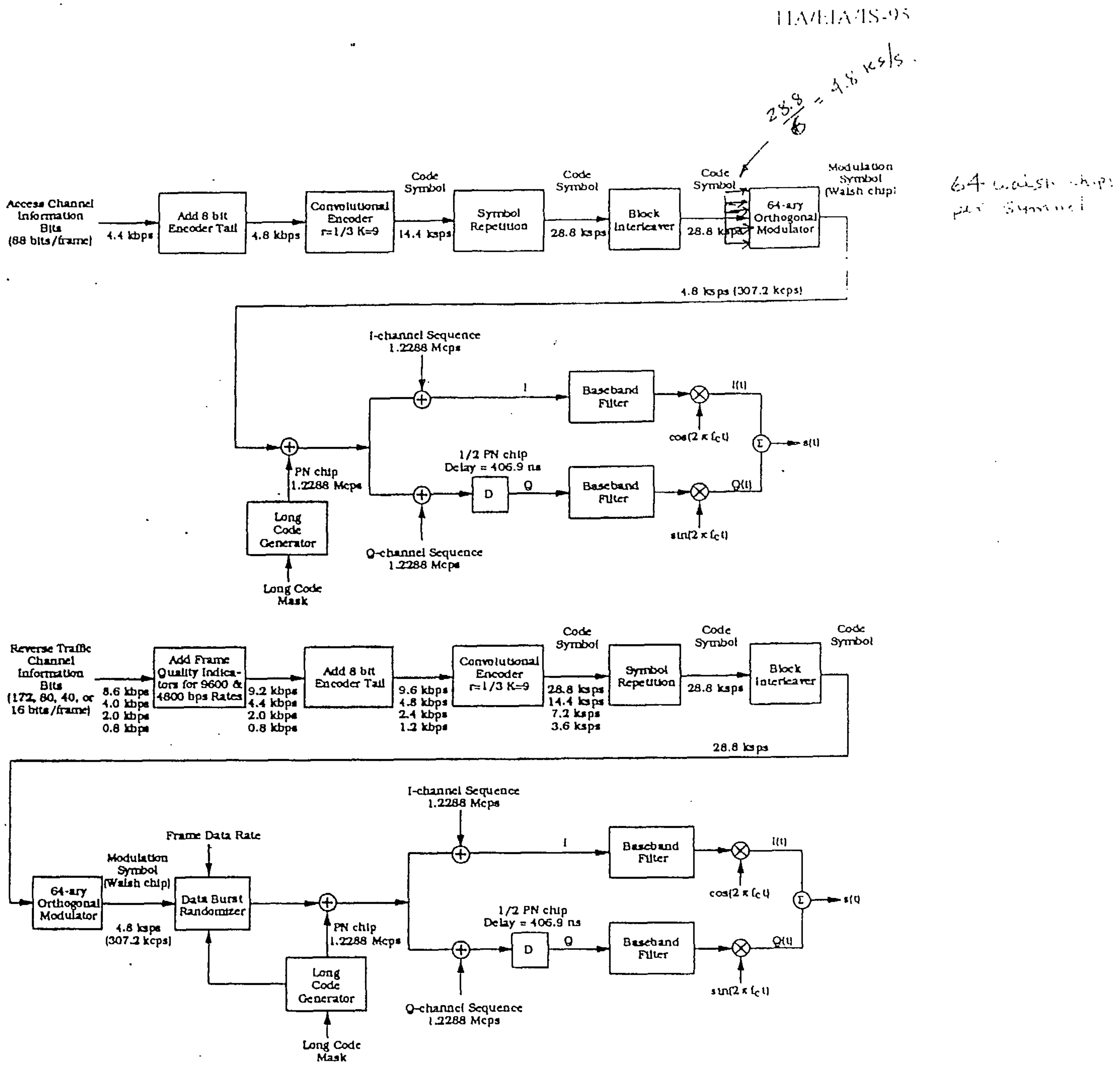

Fig. A-6 Reverse CDMA channel structure [from TIA/EIA/IS-95]

Appendix B

PRELIMINARY PRODUCT BRIEF:

DIGITAL, FAST ACQUISITION,

SPREAD SPECTRUM BURST PROCESSOR

STEL-2000

To prospective users of the STEL-2000:

Stanford Telecom is pleased to provide this Preliminary Product Brief on the STEL-2000 Digital Spread Spectrum Fast Acquisition Burst Processor. This is intended to keep you informed of this forthcoming product prior to the release of the full data sheet.

This brief describes the main features and benefits of the new STEL-2000, currently being tested in our laboratories. The preliminary results obtained so far indicate that the device is operating as expected. ***Official product release is anticipated during the second quarter of 1993, at which time the full data sheet will also be made available. Stanford Telecom reserves the right to change the specifications of this product prior to the release of both the product and the final data sheet.***

Customers may choose to design this device as-is into board level products and end systems. Note, however, that the STEL-2000 has been designed with customization in mind so, that derivative products may be targeted to specific applications. In either case, Stanford Telecom is available to support customer-specific design activities.

Please feel free to contact us to discuss your applications and how we can service your requirements. Our telephone number is (408) 541-9031, fax (408) 541-9030.

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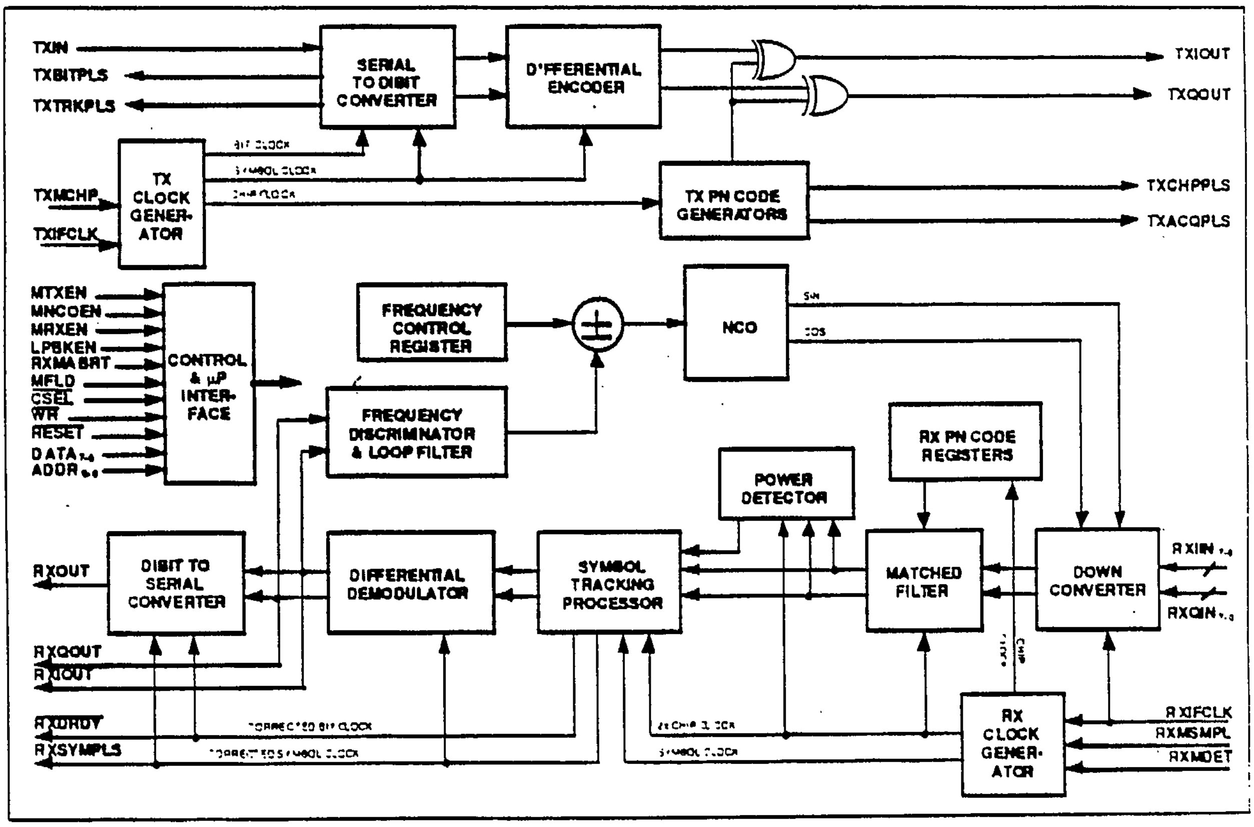
- Complete Digital CMOS Direct Sequence Spread Spectrum Burst Modem
- Operates at up to 10 MChips/Sec. in Transmit and Receive Modes
- Acquisition Within One Symbol Using Digital PN Matched Filter
- Independent PN Sequences up to 64 Bits Long for Acquisition/ Preamble and Data Symbols
- 160-Pin PQFP Package
- Full or Half Duplex Operation

BENEFITS

- High Performance and Reliability, Low Power, Low Manufacturing Costs
- Ideal for Wireless Local Area Networks (WLANs) up to 2 Mbps
- Very Low Acquisition Overhead When Operating in Burst Mode
- Long Code Acquisition/Preamble Symbol Permits High Acquisition Probability Even with Short Code Data Symbols
- Small Footprint, Surface Mount
- Permits Dual Frequency (Frequency Division Duplex) or Single Frequency (Time Division Duplex) Operation

BLOCK DIAGRAM



INTRODUCTION

The STEL-2000 is a single ASIC which performs all the digital processing required to implement a Fast Acquisition Direct Sequence Spread Spectrum Full Duplex BPSK or QPSK modem. The device implements all of the functions performed by the STEL-2130 Digital Downconverter, the STEL-3340 Digital Matched Filter and the STEL-2120 Differential Demodulator ASICs, as well as adding several new features not previously incorporated into ASICs which would otherwise require significant additional logic to implement. The new functions include a differential BPSK/QPSK encoder and PN spreader for the transmit function, so that only a modulator is required to complete the transmitter. Programmable arbitrary PN code generation for both the transmitter and receiver is also provided. The PN codes can be up to 64 bits long.

The STEL-2000 operates at up to 10 Mchips/sec. (Mcps) in both transmit and receive modes, and the code length used determines the maximum data rate. Operating in QPSK mode the device transmits two bits of data per symbol, and therefore the maximum data rate for a PN code of length N will 20/N Mbps. Conversely, this relationship can be used to determine the maximum possible PN code length for a given data rate.

The STEL-2000 uses a Numerically Controlled Oscillator (NCO) and complex multiplier to perform full single sideband downconversion directly from I.F. to baseband using dual Analog to Digital Converters (ADCs). At lower chip rates (below approximately 4 Mcps) a single ADC may be used, as described in the Applications section. The I.F. frequency is not limited by the capabilities of the STEL-2000 but by the track-and-hold capabilities of the ADC selected. Signals at I.F. frequencies up to about 100 MHz can be processed by currently available 8-bit ADCs, but the implementation cost, as well as the performance, can be improved by using I.F frequencies in the 30 MHz region or lower.

The STEL-2000 is designed to operate in either burst or continuous mode; in the latter case the data is simply treated as a burst of infinite length. The use of a Digital PN Code Matched Filter for code detection and despreading permits signal and symbol timing acquisition in just one symbol. Consequently, the device automatically precedes data bursts by a single Acquisition/Preamble symbol, and different PN codes can be used for spreading the preamble and data symbols. In this way a long PN code can be used for the Acquisition/Preamble symbol to maximize the probability of burst detection even when circumstances such as data rate dictate a shorter code for the data symbols. To improve the performance in the presence of high noise and interference levels in such cases the symbol timing recovery circuit also incorporates a flywheel circuit to maximize the probability of correct symbol timing. This circuit will insert a symbol clock at the correct time if the correlation peak in the matched filter fails to exceed the set threshold at the expected time during a given symbol. A missed detect counter is incorporated and this can be used to abort a burst in the presence of abnormally high interference. A timing gate circuit also minimizes the probability of false detection of symbol clocks due to noise or interference.

To minimize power consumption, individual sections of the device can be turned off when not in use. For example, the NCO and receiver circuitry can be turned off during transmission when operating in Time Division Duplex (TDD) mode. The fast acquisition characteristics of the device also make it ideal for use in applications where bursts are transmitted infrequently, so that the device can be programmed to be in full "sleep" mode for most of the time, reducing power practically to zero during this time.

FUNCTION BLOCKS - DESCRIPTION

Transmit and Receive Clock Generator

The timing in the transmitter and receiver sections of the STEL-2000 are controlled by the Transmit and Receive Clock Generator Blocks. These blocks are programmable dividers providing signals at the chipping and symbol rates (as well as multiples and submultiples of these frequencies) according to the information programmed in the control registers. The transmit and receive clocks can be mutually asynchronous since the transmitter and receiver sections are completely independent.

Serial to Dibit and Dibit to Serial Converters

When the transmitter and receiver are operating in QPSK mode, the data to be transmitted and the received data is processed in pairs of bits (dibits), one bit for the I channel and one for the Q channel The dibits are transmitted and received as single QPSK symbols. The single-bit wide I/O data is converted to and from this format in the Serial to Dibit Converter Block in the transmitter and the Dibit to Serial Converter in the receiver. The receiver will also generate two clock pulses per symbol when operating in QPSK mode, to allow the data bits to be clocked out sequentially at the **RXOUT** pin. The received data is also available as the I and Q dibits directly before dibit to serial conversion at the **RXIOUT** and **RXQOUT** pins.

Differential Encoder

The data to be transmitted is differentially encoded before being spread with the PN code. Since the demodulator operates differentially, no separate differential decoder is required. The encoding scheme depends on whether the modulation format is to be DBPSK or DQPSK. With DBPSK modulation the encoding algorithm is the straightforward differential scheme:

Output bit(n) = Input bit(n) $\oplus$ Output bit(n−1)

where $\oplus$ represents the logical EXOR function. However, with DQPSK modulation the algorithm is more complex since there are now sixteen possible new states, depending on the four possible previous output states and four possible new input states, as shown in the following table:

$IN(I, Q)_m$	OUT $(I, Q)_{m-1}$: 0 0	0 1	1 1	1 0
0 0	0 0	0 1	1 1	1 0
0 1	0 1	1 1	1 0	0 0
1 1	1 1	1 0	0 0	0 1
1 0	1 0	0 0	0 1	1 1
	OUT $(I, Q)_m$			

Transmitter PN Code Generation

When the STEL-2000 is used for burst signal operation, each burst requires a preamble symbol for acquisition, followed by data symbols for information transmittal. Accordingly, two separate and independent PN codes may be employed, one for spreading the acquisition/preamble symbol, and one for the information data symbols. The code lengths are completely independent of each other and can be up to 64 bits long. The PN codes are stored as code coefficients. The number of Chips per Acquisition/Preamble Symbol, the Acquisition/Preamble Symbol coefficient values, the number of Chips per Data Symbol and the Data Symbol coefficient values are all stored in memory.

STEL-2000 begins the burst by transmitting a single symbol using the Acquisition/Preamble PN code. After the Acquisition/Preamble Symbol has been transmitted the device switches into data symbol transmission mode and transmits data symbols using the Data Symbol PN code. The length of the burst is stored in memory.

The spreading codes are EXORed with the data bits (in BPSK mode) or bit pairs (in QPSK mode) to transmit one complete code sequence for every preamble and data symbol at all times. The spread I and Q channel signals are brought out as the **TXIOUT** and **TXQOUT** signals for use in an external modulator. Only **TXIOUT** is used for the BPSK mode. The carrier should be modulated as shown to be compatible with the STEL-2000 receiver:

I, Q Bits	Signal Quadrant
0 0	First
1 0	Second
1 1	Third
0 1	Fourth

Frequency Control Register and NCO

The STEL-2000 incorporates a Numerically Controlled Oscillator (NCO) to synthesize a local oscillator signal for the downconverter. The NCO is clocked with the master receiver clock signal, **RXIFCLK**. The NCO has 32-bit frequency resolution and generates quadrature outputs. The frequency is controlled by the data stored in the 32-bit Frequency Control Register. The output of the Loop Filter is added to or subtracted from this data to form the actual frequency control information.

Digital Downconverter

The STEL-2000 incorporates a Quadrature (Single Sideband) Downconverter which allows the received signal to be sampled and digitized directly at IF and converted digitally to baseband. The downconverter includes a complex multiplier in which the 8-bit receiver input signal (either real or complex) is multiplied by the sine and cosine signals from the NCO. All operations in the receiver are controlled by the master receiver clock signal, **RXIFCLK**. In conjunction with the built-in frequency tracking loop, this permits the received signal to be accurately downconverted to baseband. The input signals can be accepted in either Two's Complement or Offset Binary formats.

The outputs of the complex multiplier are:

$$I_{OUT} = I_{IN} \cdot \cos(\omega t) - Q_{IN} \cdot \sin(\omega t)$$

$$Q_{OUT} = I_{IN} \cdot \sin(\omega t) + Q_{IN} \cdot \cos(\omega t)$$

$$\text{where } \omega = 2\pi f_{NCO}$$

These are fed into the I and Q channel Integrate and Dump Filters. These filters allow the samples from the complex multiplier to be integrated over a number of sample periods. The dump rate of these filters can be controlled either by the **RXMSMPL** input signal or by an internally generated sample clock. The internal clock is generated from **RXIFCLK** in a divider whose division ratio is set by the data stored in memory. Since the accumulation process increases the magnitude of the signal values, a selectable viewport is provided at the output of the Integrate and Dump Filters to allow the appropriate output bits to be selected for the 3-bit inputs of the PN Matched Filter.

Receiver PN Code Register and PN Matched Filter

The STEL-2000 is designed for burst signal operation in which each burst requires a preamble symbol for acquisition followed by data symbols for information transmission. Two separate and independent PN codes may be employed, one for spreading the acquisition preamble symbol and one for spreading the information data symbols. The code lengths are completely independent of each other and can be up to 64 bits long. In the receiver section the signal is despread by means of the PN Matched Filter. The STEL-2000 contains a fully programmable 64-tap complex matched filter using ternary coefficients which can be set to ±1 or zero. By setting the end taps of the filter to zero the effective length of the filter can be reduced for use with PN codes shorter than 64 bits. Each coefficient is stored as a 2-bit number, so that the N-bit PN codes are stored as N 2-bit PN coefficients.

At the start of each burst, the receiver section of the STEL-2000 is automatically configured into Burst Acquisition reception mode and the Acquisition/Preamble Coefficients are used in the matched filter. Once this symbol is successfully detected the system will automatically switch into Data Symbol reception mode and the Data Symbol Coefficients will be used in the matched filter.

The PN Matched Filter computes the cross-correlation between the I and Q channel signals every chip period. To enable the STEL-2000 to operate asynchronously with respect to the chipping rate of the incoming signal, the matched filter is designed to operate with two signal samples per chip. A front end processor (FEP) in both the I and Q channels allows the incoming data to be averaged over each chip period by adding each incoming sample to the previous sample. The algorithm for this process is as shown in the following equation:

$$FEP_{OUT} = FEP_{IN}\,(1 + z^{-1})$$

The FEP can be disabled, but for normal operation it should be enabled.

The 3-bit signals from each tap in the matched filter are multiplied by the corresponding coefficient in two multiplier arrays. Each array consists of 64 multipliers which multiply the 3-bit signals by zero or ±1, according to the value of the coefficient. The products from the I and Q multiplier arrays are added together in the I and Q Adders to form the sums of the products, which represent the complex cross-correlation factor. The resulting algorithm for the correlator is:

$$\text{Output}_{(I,Q)} = \sum_{n=0}^{n=63} \text{Data}_{n(I,Q)} * \text{Coefficient}_{n(I,Q)}$$

The additions result in 10-bit signals at the outputs of the adders and two selectable viewports are provided to allow the appropriate output bits to be selected for the 8-bit inputs of the Power Detector and Demodulator Blocks.

Power Detector

The complex output of the Matched Filter is fed into a Power Detector which computes the magnitude of the vector of the I and Q channel correlation sums.

$$MAG = \sqrt{I^2 + Q^2}$$

This is accomplished by means of the approximation algorithm:

$$MAG = Max(Abs(I), Abs(Q)) + \tfrac{1}{2} Min(Abs(I), Abs(Q))$$

This 10-bit value represents the power level of the correlated signal during each chip period and is used in the Symbol Tracking Processor.

Symbol Tracking Processor

The output of the Power Detector Block represents the signal power during each chip period, which theoretically has a high peak value once per symbol. The peak will occur during the chip period when the code sequence of the received signal in the matched filter is the same as the reference PN code in the matched filter. At this time the I and Q channel outputs of the matched filter are, theoretically, the optimally de-spread I and Q symbols. The symbol power value is compared against a 10-bit threshold value to detect this chip period in each symbol period. A symbol clock pulse is generated each time the power value exceeds the threshold value, indicating a symbol detect. Since the Acquisition/Preamble Symbol and Data Symbols can have different PN codes with different peak correlation values (which depend on the code lengths). The STEL-2000 is equipped with two separate threshold registers to store the Acquisition/Preamble Threshold value and the Data symbol Threshold value. The device will automatically use the appropriate value depending on whether it is in the preamble or data mode.

The STEL-2000 is equipped with a flywheel circuit to enhance the operation of the symbol tracking function when operating under extremely adverse signal to noise ratio conditions. This is designed to ignore false detects at inappropriate times in each symbol period as well as to insert symbol clock pulses at the appropriate times if the symbol detection is missed. The flywheel circuit operates by its *a priori* knowledge of when the next detect pulse is expected. This event will occur one symbol period after the last correctly detected one and a ±1 chip window is used to gate the detect pulse. Any detects generated outside this window are ignored. A symbol detect pulse will be inserted into the symbol clock stream if the power level does not exceed the threshold inside the window, indicating a missed detect, and resulting in a missed symbol. This will be done one symbol after the last valid detect as determined by the Number of Chips/Symbol information.

The cross-correlation characteristics of a noisy signal with its own PN code may result in the "smearing" of the peak power value over the adjacent chip periods. This can result in two or three consecutive power values (the On-time, One-chip early and One-chip late values) exceeding the threshold. A maximum power selector circuit is incorporated to choose the highest of any three consecutive power levels each time this occurs, enhancing the probability that the optimum symbol timing will be chosen in such cases. This function can be disabled if desired.

The STEL-2000 also includes a circuit to keep track of missed detects. An excessively high rate of missed detects is an indication of poor signal quality and can justify aborting the reception of a burst of data, which can be up to 256 bits long. A counter is used to monitor the number of missed detects in each burst, expressed as a value. The system can be configured to automatically abort a burst and return to the Acquisition Ready status if this number exceeds the Missed Detects per Burst Threshold value. Under normal operating conditions, the STEL-2000 automatically returns to the Acquisition Ready status when the number of symbols processed is equal to the length of the burst. This function can be disabled, permitting longer bursts and continuous data to be processed.

Differential Demodulator

The I and Q symbol information generated in the Matched Filter is differentially demodulated by comparing the current symbol vector with the previous symbol vector. Two registers are used to generate the inter-symbol delay to allow the comparison of adjacent symbols. An optional 45° signal rotation is also provided to optimize the constellation boundaries in the comparison process for QPSK signals. This is not done for $\pi/4$ QPSK signals where this rotation is an intrinsic part of the signal itself. Provision is made to allow the previous symbol to be rotated by 0° or ±45°. Note that the 45° rotation introduces a scaling factor of $1/\sqrt{2}$ to the signal level in the system.

Carrier Discriminator and Loop Filter

The frequency discriminator is generated from the dot and cross products of the I and Q signals. The dot and cross products are the real and imaginary results of the complex multiplication of the current and previous symbols. The frequency discriminator uses the dot and cross product information to generate the AFC signal for the frequency acquisition and tracking loop. The algorithm used depends on the signal type.

for BPSK mode, the following algorithm is used to compute the carrier discriminator function:

CD=Cross x Sign[Dot]

For QPSK mode, the following algorithm is used to compute the carrier discriminator function:

CD=(Cross x Sign[Dot])-(Dot x Sign[Cross])

The computation of the carrier discriminator results in a 17-bit signal and a selectable viewport is provided to allow the appropriate output bits to be selected for the 8-bit input of the Loop Filter Block.

The Loop Filter transfer function can be set up to be either first or second order with coefficient values adjusted in powers of 2, from 2^0 to 2^{21}. The overall transfer function is:

$$\text{Transfer Fn.} = K1 + \tfrac{1}{4} K2 \cdot \frac{z^{-1}}{1 - z^{-1}}$$

The $^1/_4$ factor is introduced because the signal in the integrator path of the loop is divided by four after the integrator by truncating the 2 LSBs of the signal. This signal is then added to the signal in the direct path such that the LSBs of the signals are aligned. The coefficients K1 and K2 are stored in memory. As an additional feature, both the first and second order paths can be disabled, giving the user full control of the loop filter characteristics.

APPLICATIONS INFORMATION - THEORY OF OPERATION

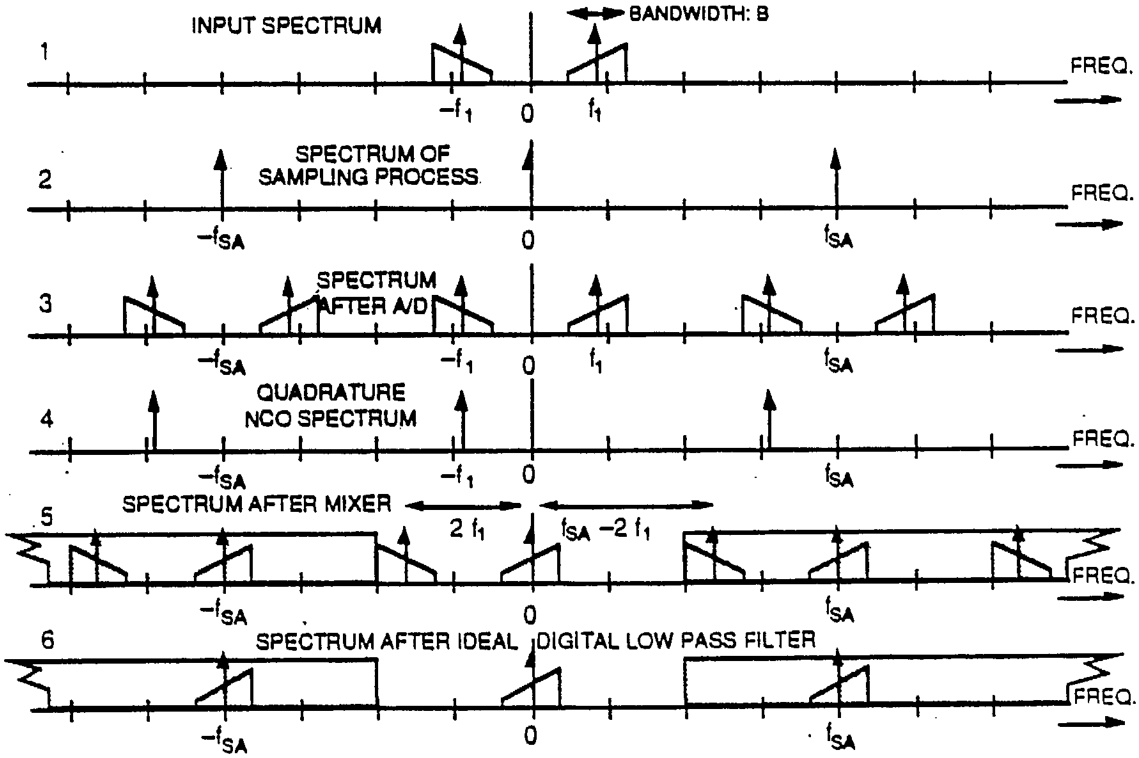
Digital Downconversion

The STEL-2000 can be used in two different sampling modes depending on the application. In data rate applications up to approximately 10% of the sample clock rate it can be used with a single ADC, referred to as the Direct I.F Sampling Mode. For high speed data applications it is necessary to use the STEL-2000 in the full Quadrature Sampling Mode, i.e., using a conventional quadrature-signal source and using the on-chip NCO in quadrature mode.

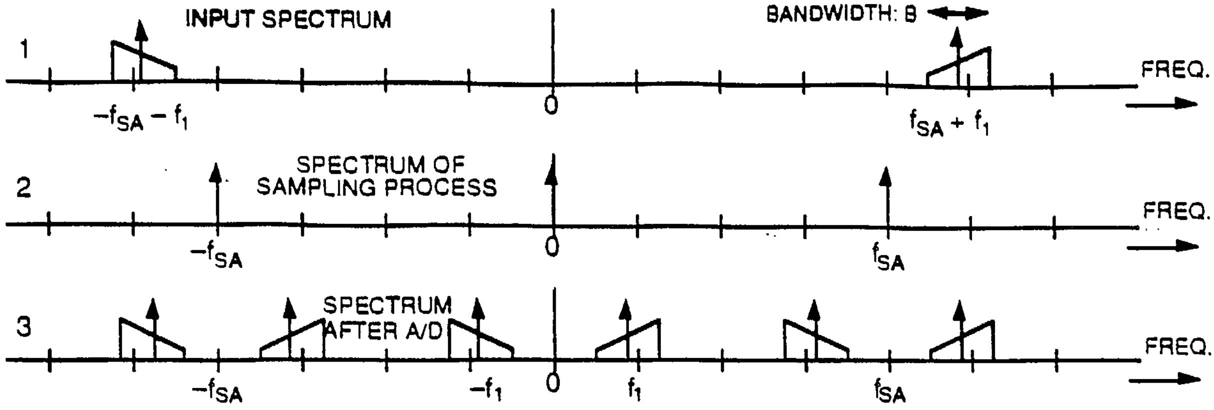
Using the STEL-2000 with a Single ADC in the Direct I.F. Sampling Mode

In this mode the incoming signal is fed directly into the complex multiplier and converted to baseband and only two of the four multipliers in the complex multiplier are required. The sine and cosine inputs to the unused multipliers, as well as the signal inputs from the Q Channel, are shut down, thus reducing the block power consumption by a factor of two. In the Direct I.F. Sampling Mode the device does not make a true single-sideband conversion from I.F. to baseband, unlike the full Quadrature Sampling Mode, using two ADCs. The consequence of this is shown spectrally in the diagram below. The input signal is a real signal and has the spectrum shown in line 1. This signal is then sampled with a signal at the frequency f_{SA} (line 2), resulting in the spectrum shown in line 3. When this signal is modulated with the complex signal of the quadrature NCO (line 4), the spectrum after the mixer is shown in line 5. The sections shown inside the shaded areas are the aliases of the baseband signal beyond the Nyquist frequency and are not of concern. The signals inside the primary Nyquist region consist of the desired signal and a spectrally reversed image signal which is separated from the former by f_1, the I.F. frequency before sampling. At this point this component must be removed by filtering, and an Integrate and Dump filter is provided in the STEL-2000 for this purpose. This type of first order filter provides more than 25 dB of attenuation of the unwanted signal if the bandwidth B of the signal is less than 10% of the sampling frequency. In this example, the input signal was shown at a low I.F. frequency. ($f_1 < {}^1\!/_2\, f_{SA}$, i.e. the signal was inside the primary Nyquist region.) However, provided that $B < {}^1\!/_2\, f_{SA}$ this is not necessary as long as the input is completely contained in any of the non-primary Nyquist regions ($N.{}^1\!/_2\, f_{SA}$ to $(N+1).{}^1\!/_2\, f_{SA}$). This is shown in the second diagram, where it can be seen that in line 3 of the diagram the high frequency input has the same spectrum after sampling as the low frequency input had in the original case; consequently the processes following this are identical to those in the original case.

SPECTRA OF SIGNALS IN DIRECT I.F. SAMPLING MODE



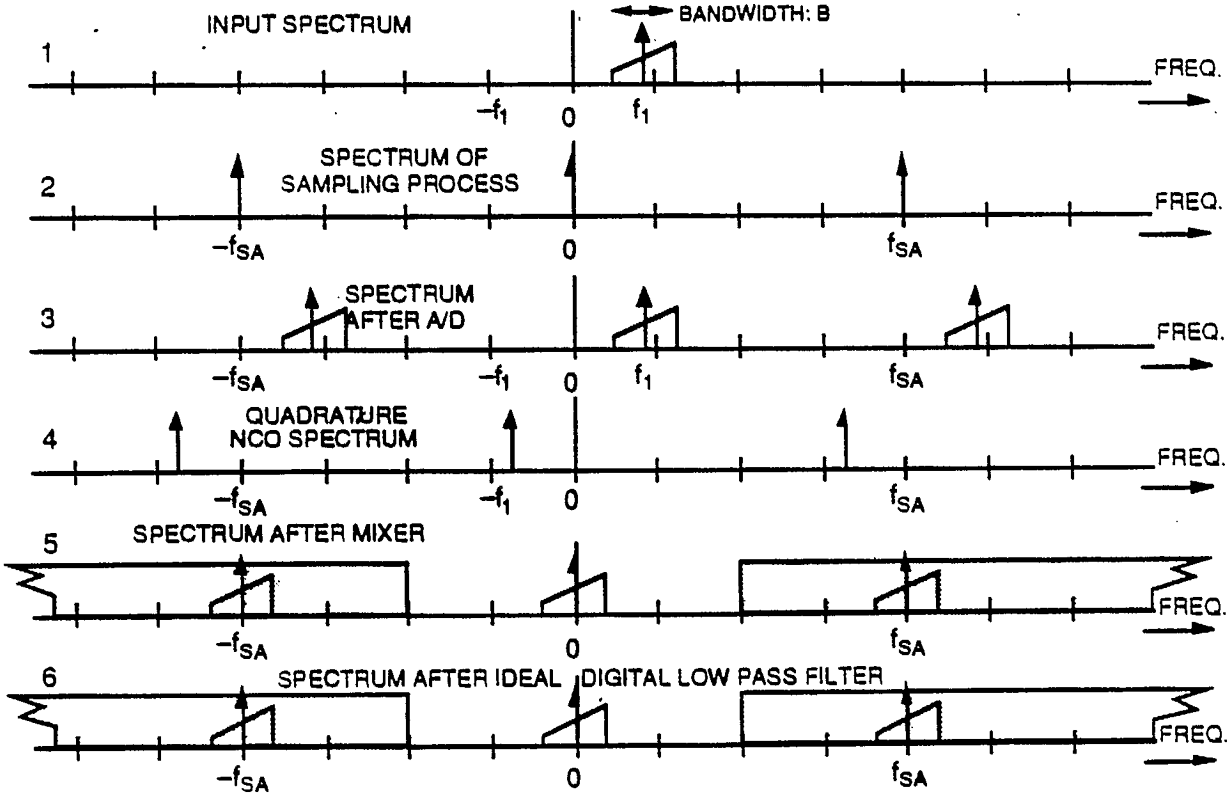
SPECTRA OF SIGNALS IN DIRECT I.F. SAMPLING MODE WHEN THE I.F. FREQUENCY IS HIGHER THAN THE SAMPLING FREQUENCY



USING THE STEL-2000 WITH TWO ADCs

Signals with higher symbol rates (more than 10% of f_{SA}) can be processed with the STEL-2000 by using the Quadrature Sampling Mode. In this mode the device performs a true single-sideband conversion to baseband; the image sideband associated with the Direct I.F. Sampling Mode does not appear, as shown in the diagram below. Since both the input and NCO signals are now complex the spectrally reversed image found in the Direct I.F. Sampling Mode no longer exists and only the Nyquist limit constrains the maximum value of B.

SPECTRA OF SIGNALS IN QUADRATURE SAMPLING MODE



DIFFERENTIAL DEMODULATION

The basic operation consists of generating the complex conjugate product between two input samples, one or two samples apart. The k-th input sample, $s_{in}(k)$, is defined as:

$$s_{in}(k) \equiv I(k) + j\,Q(k)$$

where I(k) and Q(k) are the 8-bit inputs to the I/Q processor block. In the following equations the polar form of $s_{in}(k)$ is used and is defined as:

$$s_{in}(k) \equiv A(k).e^{j\varnothing(k)}$$

with

$$A(k) \equiv \sqrt{I^2(k)+Q^2(k)}$$

$$\varnothing(k) \equiv \arctan\left(\frac{Q(k)}{I(k)}\right)$$

The demodulated output data on the Dot and Cross outputs are:

$$s_{out}(k) = s_{in}(k) . [s_{in}(k-1) . \omega_{fixed}]$$

$$\equiv Dot(k) + j\,Cross(k)$$

where

ω_{fixed} represents a ±45° phase rotation (1 ± j1), needed for QPSK only

$Dot(k) = Re[s_{out}(k)]$

$Cross(k) = Im[s_{out}(k)]$

BPSK Demodulation

For DBPSK the phase difference between successive samples is the sum of 1) the data modulation, $\Delta\varnothing_{mod}$, which is 0° or 180° for DBPSK, plus 2) the phase rotation, $\Delta\varnothing_{rot}$, due, for example, to a frequency offset between the received signal and the local downconverter:

$$s_{out}(k) = A(k).A(k-1).e^{j\varnothing(k)}.e^{-j\varnothing(k-1)}$$

$$= A(k).A(k-1).e^{j(\Delta\varnothing_{mod}(k)-\Delta\varnothing_{rot}(k))}$$

For DBPSK, only the real part of $s_{out}(k)$ is used:

$$Dot(k) = \pm A(k).A(k-1).\cos(\Delta\varnothing_{mod}(k)-\Delta\varnothing_{rot}(k))$$

$$= \pm A(k).A(k-1).\cos(\Delta\varnothing_{rot}(k))$$ with the sign determined by the transmitted data.

$$\approx \pm A^2(k)$$ if the amplitude of the signal is constant during consecutive symbols, and if the phase rotation due to the frequency offset is small.

QPSK Demodulation

For DQPSK modulation the phase shifts due to modulation are 0°, 90°, 180°, or 270°. The STEL-2000 is to be configured to provide an extra 45° phase shift so that the possible phase differences become 45°, 135°, 225°, or 315°, which makes the decision boundaries coincide with the sign of Dot and Cross. For QPSK both the Dot and Cross outputs are used:

$$Dot(k) = \pm A(k).A(k-1).\cos(\Delta\varnothing_{mod}(k)-\Delta\varnothing_{rot}(k)+45°) \approx \pm A^2(k)$$

$$Cross(k) = \pm A(k).A(k-1).\sin(\Delta\varnothing_{mod}(k)-\Delta\varnothing_{rot}(k)+45°) \approx \pm A^2(k)$$

where we again assumed that the phase rotation due to the frequency offset is negligible.

For π/4 QPSK the modulator inserts the 45° between consecutive samples, so the STEL-2000 should be configured with $\varnothing_{Fixed} = 0°$.

FREQUENCY ERROR GENERATION

The frequency error is generated by calculating the sine of the phase difference between the present and prior symbol after correcting for the estimated increments due to data modulation. In the STEL-2000 the frequency error is calculated through a decision directed cross-product algorithm and is used in the Loop Filter to drive the NCO frequency. The decision directed cross-product algorithm calculates the following on the basis of the input:

$$s_{in}(k) = I(k) + j\,Q(k) \text{ for BPSK:}$$

$$\begin{aligned} s_{AFC/BPSK}(k) &= \mathrm{SIGN}[\mathrm{Dot}(k)]\,.\,\mathrm{Cross}(k) \\ &= \mathrm{SIGN}[\mathrm{Dot}(k)]\;\; A(k).A(k-1).\sin(\varnothing(k)-\varnothing(k-1)) \\ &= \mathrm{SIGN}[\mathrm{Dot}(k)]\,.\,A^2(k).\cos(\Delta\varnothing_{mod}(k).\sin(\Delta\varnothing_{rot}(k)) \end{aligned}$$

$$\text{i.e., } s_{AFC/BPSK}(k) = A^2(k).\sin(\Delta\varnothing_{rot}(k)$$

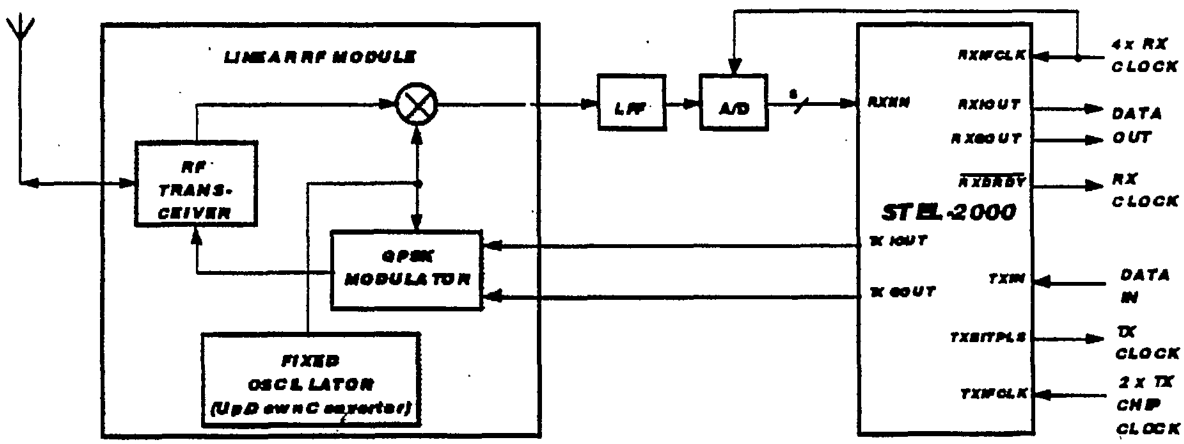
The last equation indicates an error signal related to the change of phase between successive samples. As the time between successive samples is fixed, this is equivalent to a known frequency error.

For QPSK:

$$s_{AFC/QPSK}(k) = \mathrm{SIGN}[\mathrm{Dot}(k)]\,.\,\mathrm{Cross}(k) - \mathrm{SIGN}[\mathrm{Cross}(k)]\,.\,\mathrm{Dot}(k)$$

$$\text{i.e., } s_{AFC/QPSK}(k) = A^2(k).\sin(\Delta\varnothing_{rot}(k))$$

TYPICAL APPLICATION
SPREAD SPECTRUM BPSK/QPSK TRANSCEIVER



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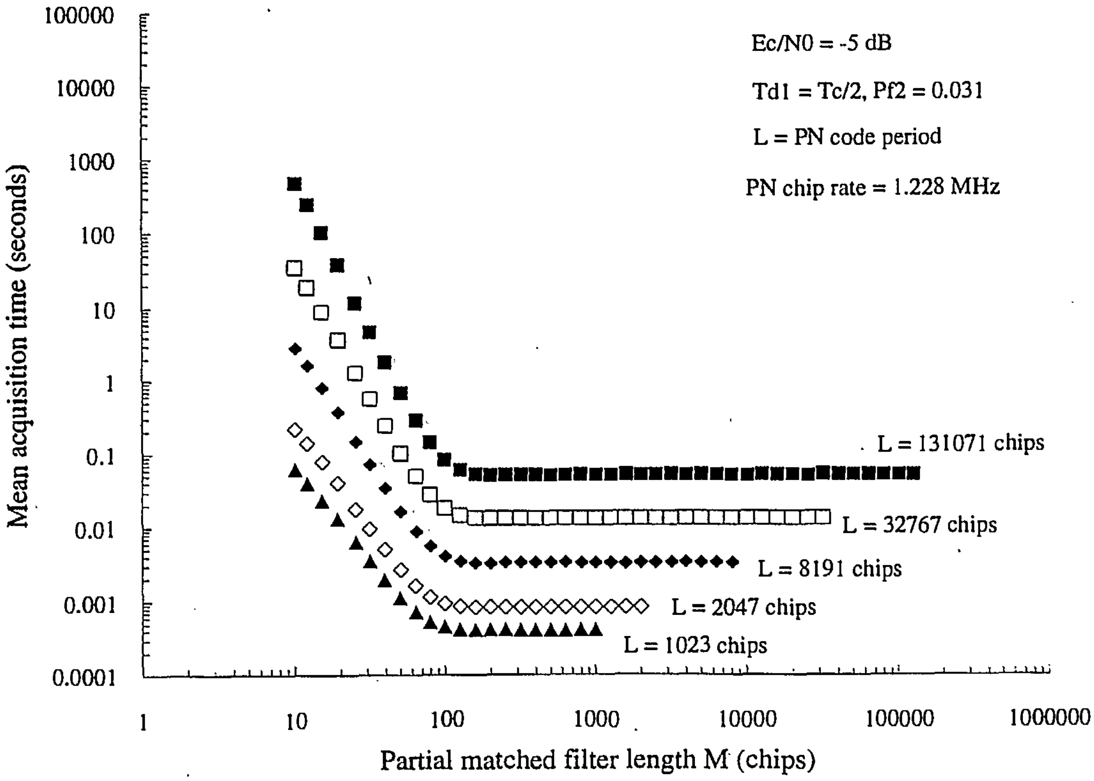
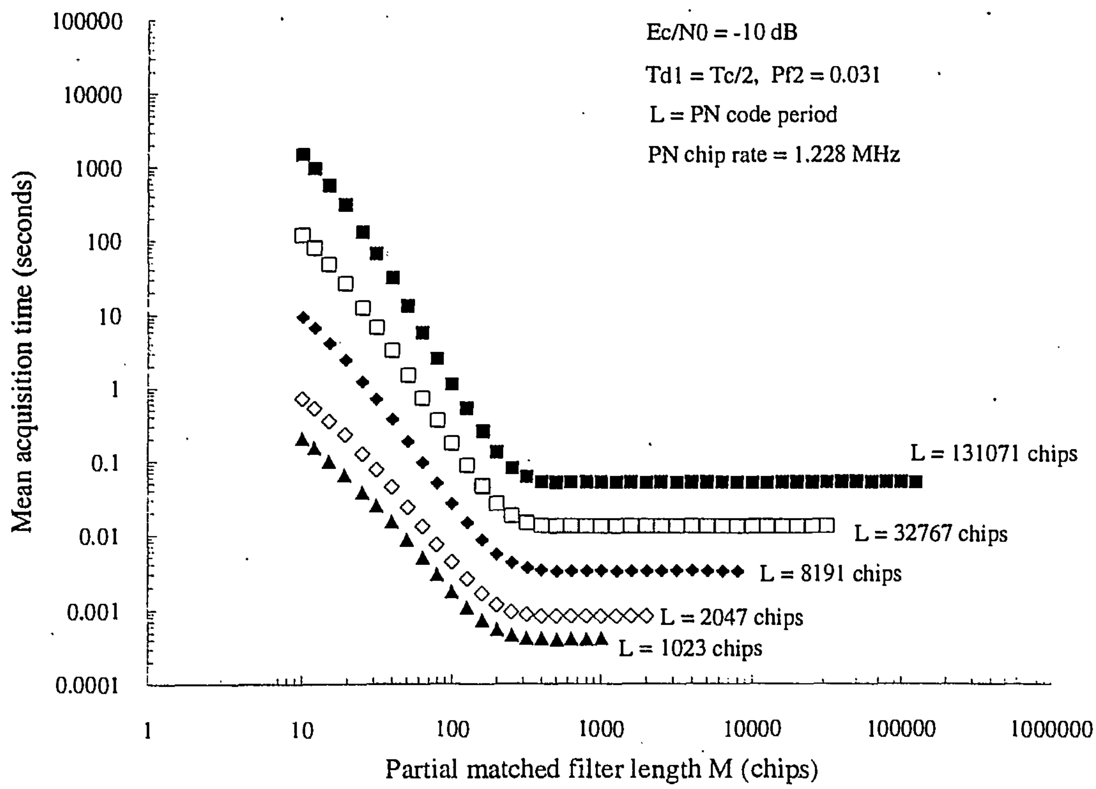
Appendix C


Ec/N0 = -5 dB
Td1 = Tc/2, Pf2 = 0.031
L = PN code period
PN chip rate = 1.228 MHz
L = 131071 chips
L = 32767 chips
L = 8191 chips
L = 2047 chips
L = 1023 chips
Mean acquisition time (seconds)
Partial matched filter length M (chips)
100000
10000
1000
100
10
1
0.1
0.01
0.001
0.0001
1
10
100
1000
10000
100000
1000000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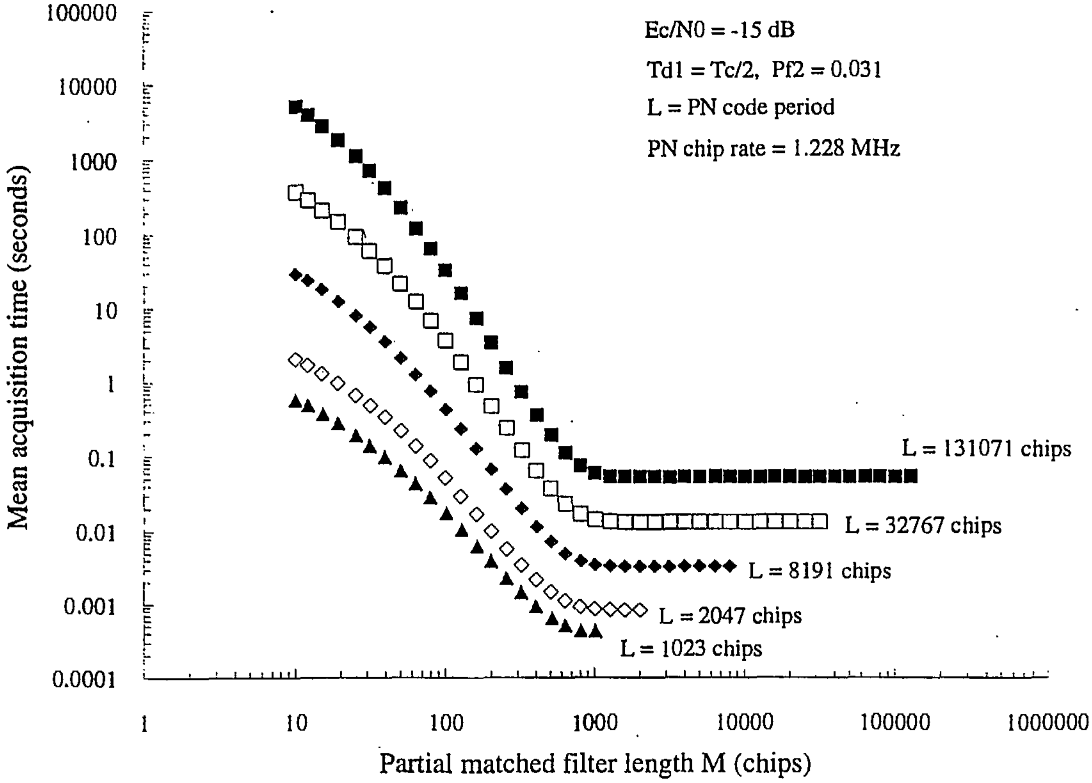
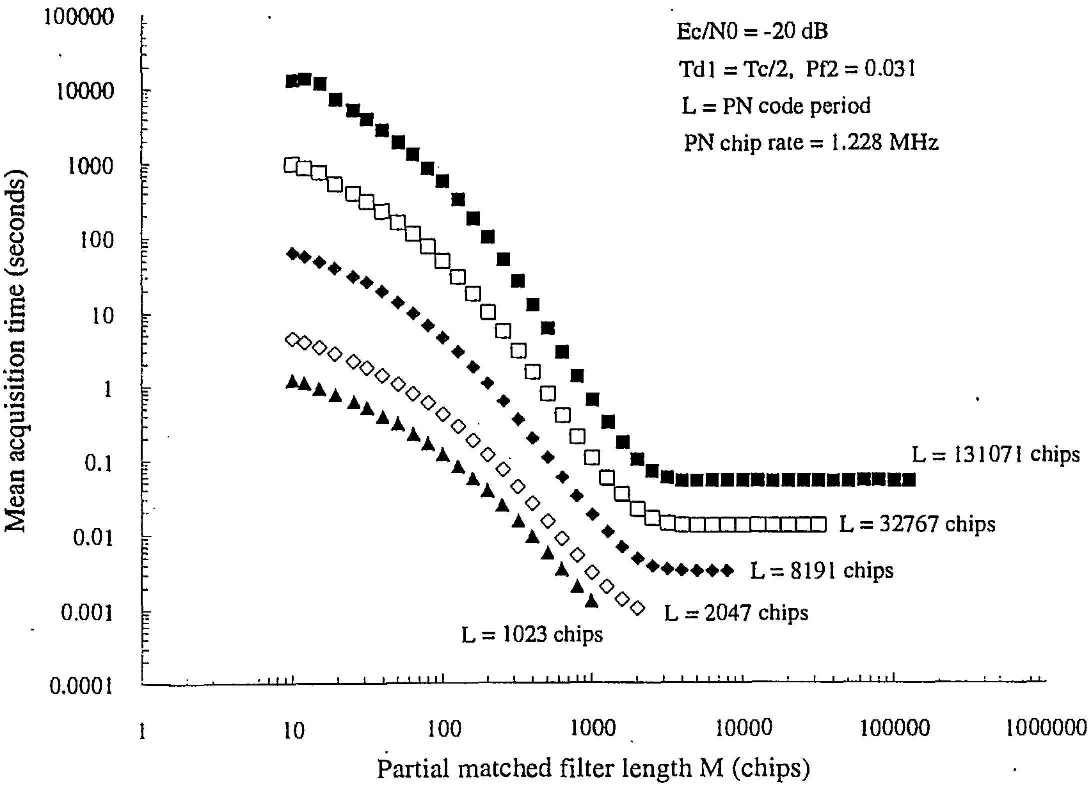
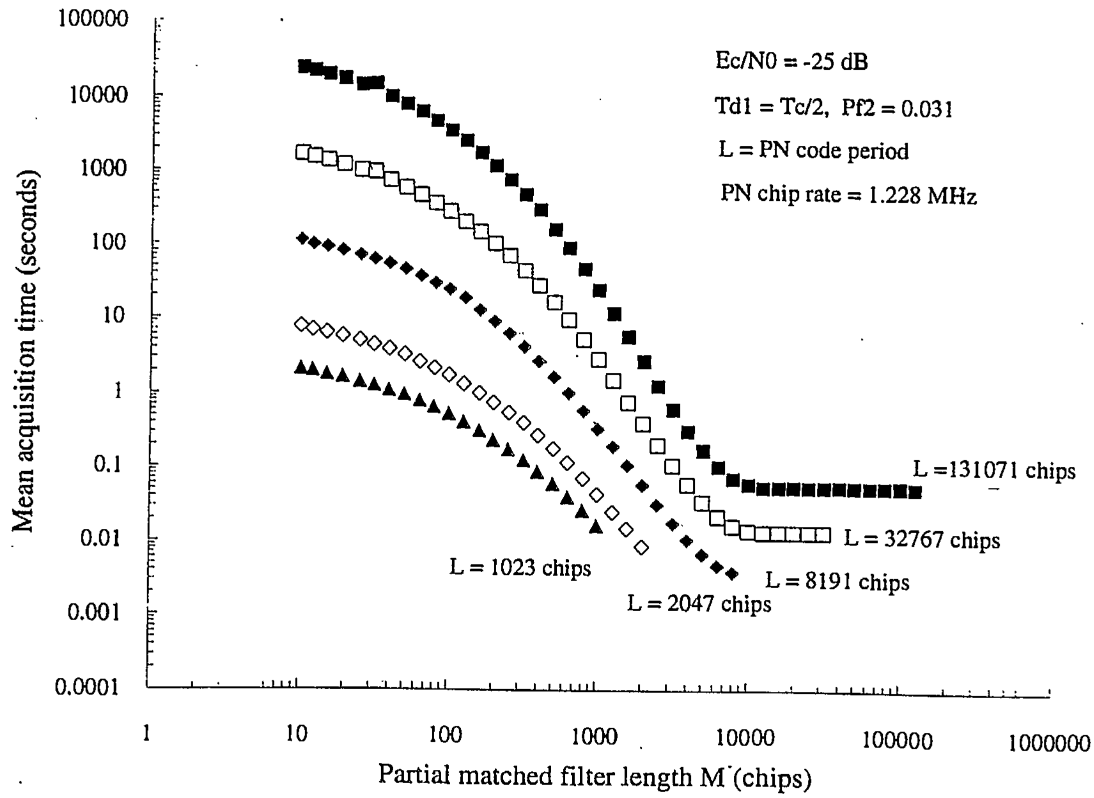


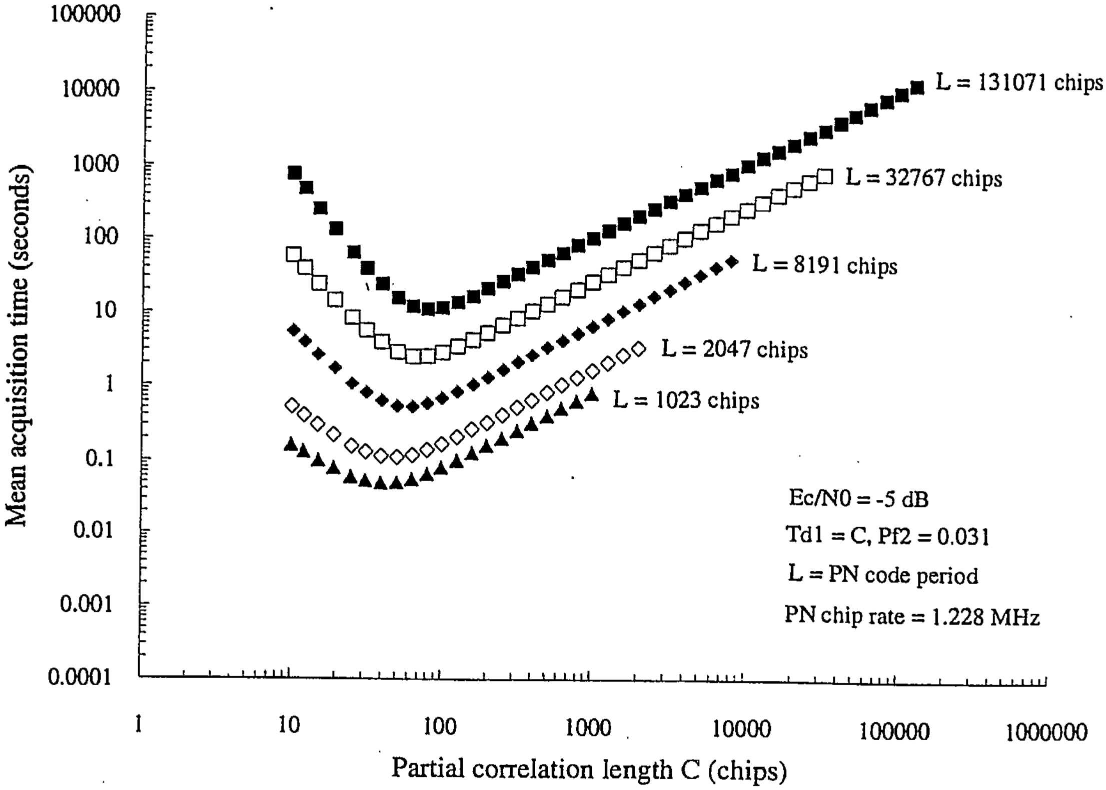
Ec/N0 = -15 dB
Td1 = Tc/2, Pf2 = 0.031
L = PN code period
PN chip rate = 1.228 MHz
L = 131071 chips
L = 32767 chips
L = 8191 chips
L = 2047 chips
L = 1023 chips
Mean acquisition time (seconds)
Partial matched filter length M (chips)



Ec/N0 = -20 dB
Td1 = Tc/2, Pf2 = 0.031
L = PN code period
PN chip rate = 1.228 MHz
L = 131071 chips
L = 32767 chips
L = 8191 chips
L = 2047 chips
L = 1023 chips
Mean acquisition time (seconds)
Partial matched filter length M (chips)



Ec/N0 = -25 dB
Td1 = Tc/2, Pf2 = 0.031
L = PN code period
PN chip rate = 1.228 MHz
L =131071 chips
L = 32767 chips
L = 8191 chips
L = 2047 chips
L = 1023 chips
Mean acquisition time (seconds)
Partial matched filter length M (chips)
100000
10000
1000
100
10
1
0.1
0.01
0.001
0.0001
1
10
100
1000
10000
100000
1000000


Mean acquisition time (seconds)
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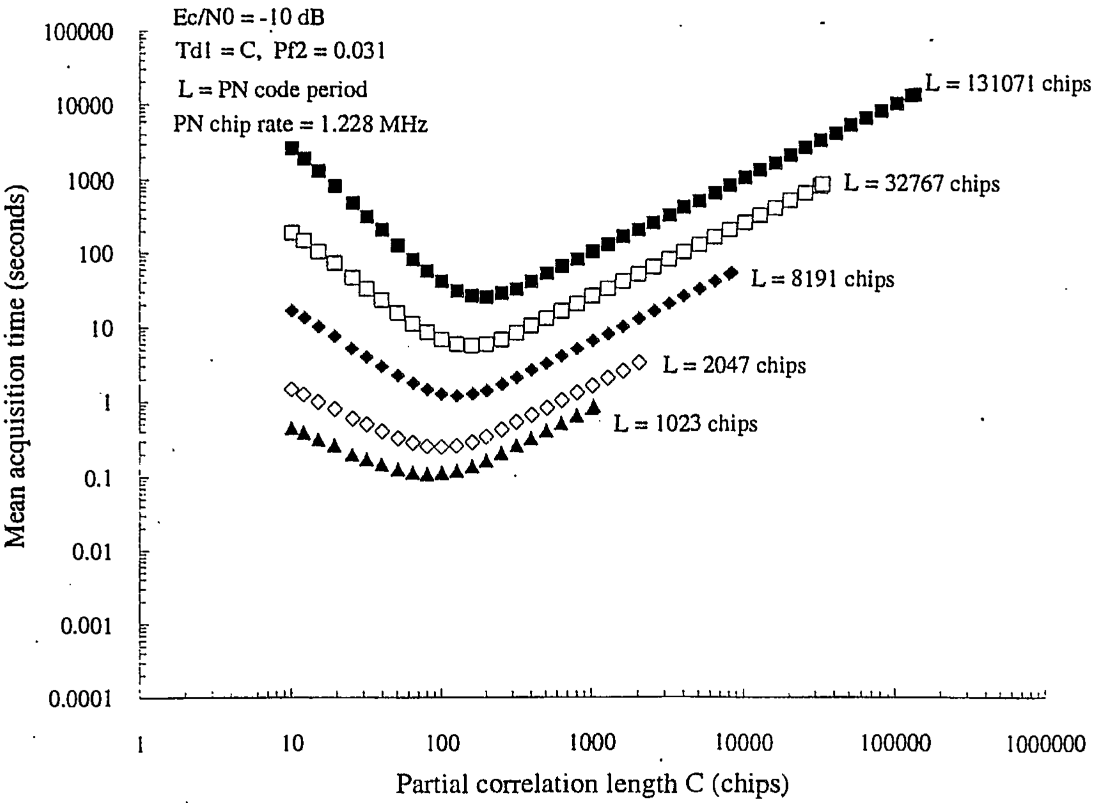
10000
1000
100
10
1
0.1
0.01
0.001
0.0001
1
10
100
1000
10000
100000
1000000
Partial correlation length C (chips)
L = 131071 chips
L = 32767 chips
L = 8191 chips
L = 2047 chips
L = 1023 chips
Ec/N0 = -5 dB
Td1 = C, Pf2 = 0.031
L = PN code period
PN chip rate = 1.228 MHz


Ec/N0 = -10 dB
Td1 = C, Pf2 = 0.031
L = PN code period
PN chip rate = 1.228 MHz
Mean acquisition time (seconds)
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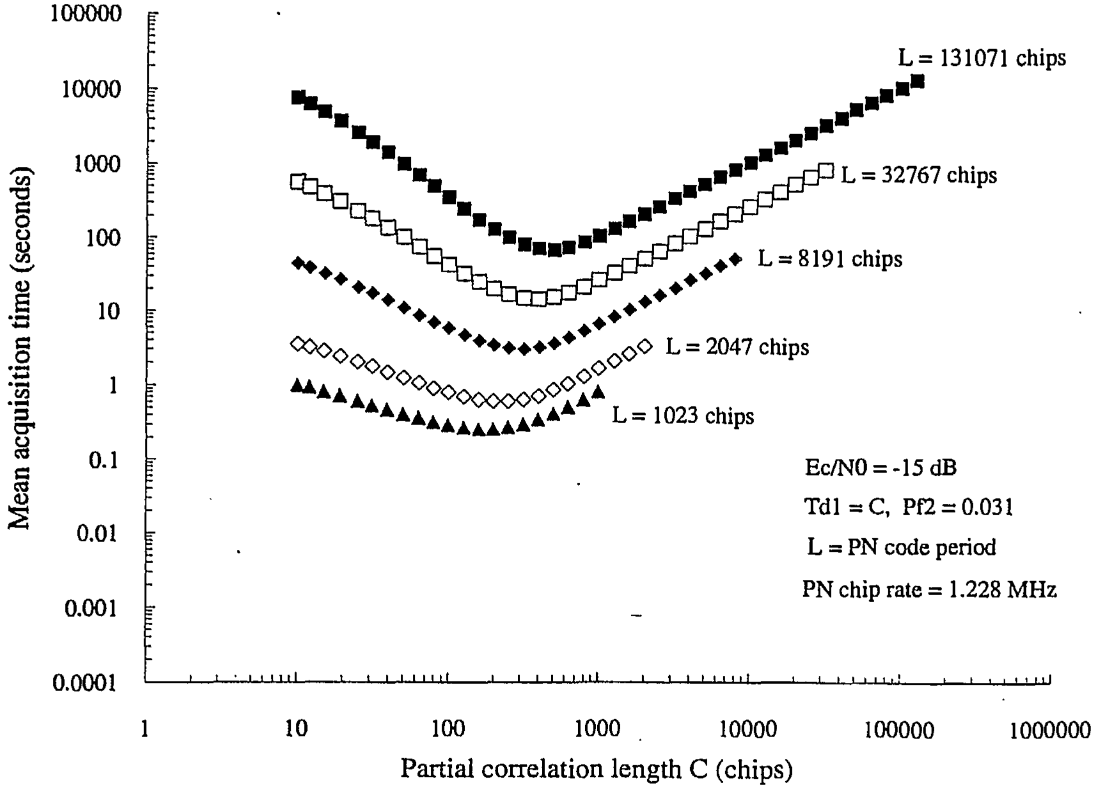
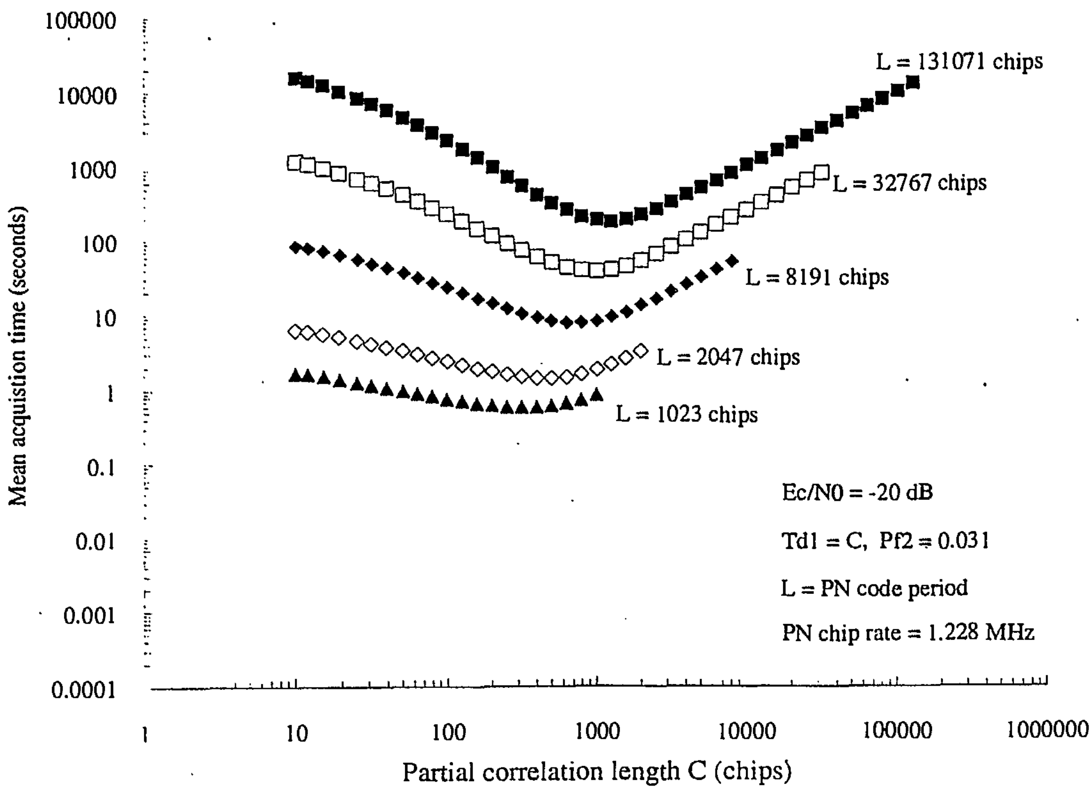
100
10
1
0.1
0.01
0.001
0.0001
1
10
100
1000
10000
100000
1000000
Partial correlation length C (chips)
L = 131071 chips
L = 32767 chips
L = 8191 chips
L = 2047 chips
L = 1023 chips



L = 131071 chips
L = 32767 chips
L = 8191 chips
L = 2047 chips
L = 1023 chips
Ec/N0 = -15 dB
Td1 = C, Pf2 = 0.031
L = PN code period
PN chip rate = 1.228 MHz
Mean acquisition time (seconds)
Partial correlation length C (chips)



L = 131071 chips
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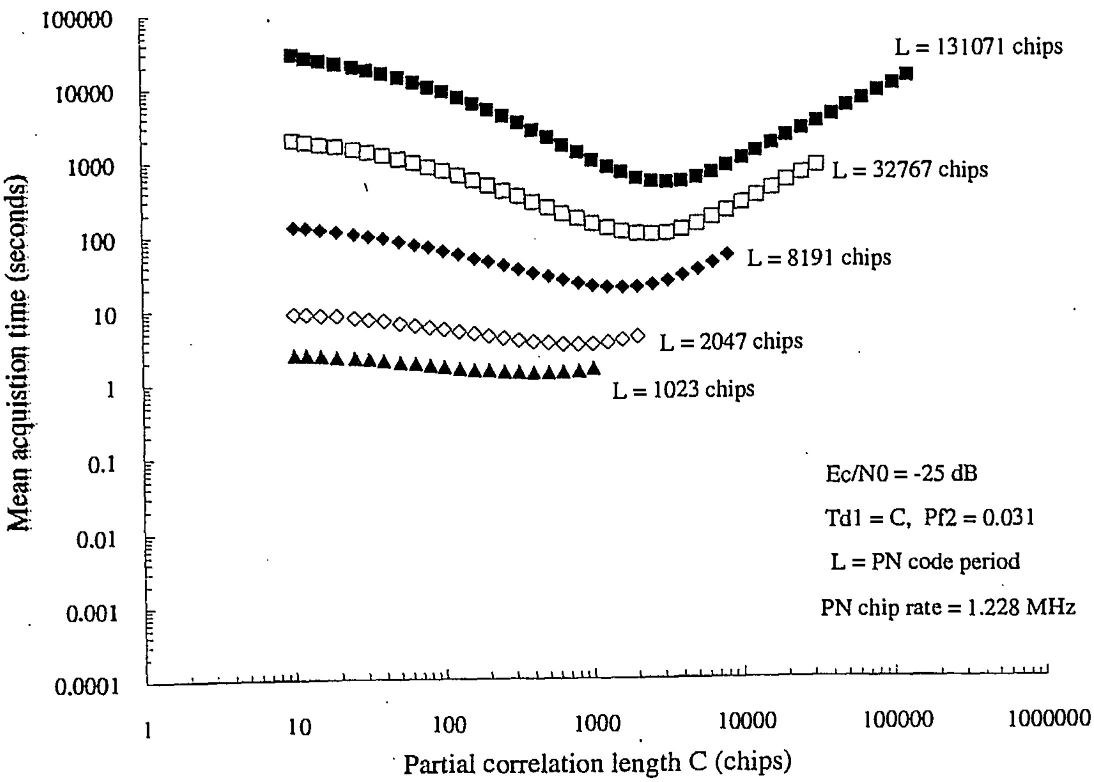
L = 8191 chips
L = 2047 chips
L = 1023 chips
Ec/N0 = -20 dB
Td1 = C, Pf2 = 0.031
L = PN code period
PN chip rate = 1.228 MHz
Mean acquistion time (seconds)
Partial correlation length C (chips)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10
1
0.1
0.01
0.001
0.0001
Mean acquistion time (seconds)
1
10
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1000
10000
100000
1000000
Partial correlation length C (chips)
L = 131071 chips
L = 32767 chips
L = 8191 chips
L = 2047 chips
L = 1023 chips
Ec/N0 = -25 dB
Td1 = C, Pf2 = 0.031
L = PN code period
PN chip rate = 1.228 MHz

Appendix D

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