SYNOPSIS

Among the various techniques employed for communication in the presence of noise and interference, the idea of using a common channel with large time-bandwidth (TB) product has been successfully exploited. Large TB products can be achieved by selecting long binary codes and encoding the message signal by them for transmission over the common channel. The channel is made available to other users simultaneously by allotting orthogonal codes to them, which places many restrictions on the types of codes which can be used for encoding the different messages. The spread-spectrum (SS) technique derives its advantages from the suppression of the noise and interference brought about by pulse compression at the receiver.

The objective of the present thesis has been to study the spread-spectrum-technique from its various aspects such as the effect of noise and interference, the design and development of analogue and digital matched-filter-correlators (MFC) and the problems of multiplexing and multiaccessing. The difficulties associated with the MFC for the long codes led to the development of the combined codes for obtaining large time-bandwidth product. The correlation properties and the detection scheme of the combined codes have been studied in detail. The application of the combined codes in data transmission and in multiaccess systems have also been considered.

It is known that larger the TB product is in a system, more immune it becomes to the noise and jamming of limited power and can cater for a larger number of simultaneous users. The performance of a system using large TB product has been examined in detail in the presence of the power limited noise, cochannel interference and impulsive noise. The probability of bit error P_{eg} in the presence of constant power Gaussian noise can be calculated from

$$P_{eg} = \frac{1}{2} - \Phi\left(\sqrt{2 R_g L}\right) \quad .. \quad (1)$$

where R_g represents the peak signal-power to average noise-power ratio (SNR) at the input of the receiver when the noise power is measured in a bandwidth equal to the bit rate, L is the length of the code which equals the bandwidth expansion factor used for achieving a large TB product and

$$\Phi(x) = \frac{1}{\sqrt{2\pi}} \int_0^x e^{-u^2/2} \, du .$$

In (1) putting L = 1, gives the probability of error in the case of direct bipolar-video-detection.

In the presence of cochannel interference and jamming noise the probability of bit error P_{ec} has been obtained as

$$P_{ec} = \frac{1}{2} - \Phi\left[\frac{2L}{\left(K-1 + \frac{J}{D}\right)\left(1 + \frac{\eta_0 B}{p}\right) + \frac{\eta_0 B}{p}}\right]^{1/2} \quad .. \quad (2)$$

where K is the number of simultaneous users, J is the jamming power, D is the desired power, η_0 is the one sided natural noise-power-spectral-density, B is the channel bandwidth and P is the total received power.

From (2) it is obvious that for a fixed code length i.e. for a fixed bandwidth expansion, and negligible natural noise power, the bit-error-probability increases when the number of simultaneous users K is increased. Similarly, when the jamming power, J, is increased, the probability of error increases. In order to decrease the probability of error, the TB product has to be increased and thus, requiring an increase in the code length.

It is interesting to note from (1) that theoretically the SS technique brings about an improvement in the output SNR over the input SNR by a factor of L. In an ideal communication system operating at its maximum capacity it has been shown that

$$(S/N)_O = \left[1 + (S/N)_I\right]^b - 1$$

$$\simeq b.\ (S/N)_I\ ,\ \text{when}\ \ (S/N)_I << 1 \qquad .. \quad (3)$$

where b = bandwidth expansion factor and the subscripts I,and O refer to the input and the output respectively. In our case, the input SNR, $(S/N)_I << 1$ and b = L. Thus, theoretically, with the SS technique, the SNR improves linearly with the bandwidth expansion in a manner similar to the ideal case when the input SNR << 1. When the input SNR >> 1, the improvement in the ideal system is exponential whereas in the SS system it remains linear.

In the presence of impulsive noise, the bit-error-probability P_{ei} has been calculated by using a simple technique as

$$P_{ei} = \frac{1}{(1 + L\sqrt{R_i})^3} \qquad .. \quad (4)$$

where R_i is the ratio of peak-signal-power to average-noise-power due to the impulsive noise at the input of the receiver. It is seen from (4) that as the impulse noise power increases, the probability of error increases for a fixed signal power. On the other hand for a fixed R_i, by increasing the TB product through the code length L, the P_{ei} decreases, though the decrease is not as much as in the case of the cochannel interference.

In order to decrease the probability of error in all the three cases, the code length L has to be increased as can be seen from (1), (2) and (4). Though long codes with good correlation properties are available, their corresponding receiver implementation requires a large amount of hardware. Combined codes have been suggested by the author here as an alternative to single long codes for the purpose of increasing the TB product and yet simplify the receiver as explained later. These codes are formed by replacing each bit of a code of length L_o, called the outer code, by another code of length L_i, called the inner code, to result in a combined code of length $L_i \times L_o$. Proper selection of the inner and the outer codes will yield a combined code which will meet the requirements in a given situation in a manner similar to a single code of length $L_i \times L_o$. The probability of error P_e can be calculated by using (1), (2) and (4) by simply replacing L by $L_i \times L_o$. With the same P_e as in the case of a single code of equivalent length, the combined code has the great advantage of a simpler receiver implementation.

In communication systems using the SS technique, the receiver is a matched-filter-correlator which may be made of delay lines. In the case of the receiver for the combined codes, a two stage MFC is used. The first matched filter is for the inner code and thus, requires a delay line of length L_i. The first matched filter gives at its output the pulses corresponding to the outer code. These pulses are stretched by a pulse stretcher in order to make them compatible with the second matched filter matched to the outer code. The second matched filter requires a delay line of length L_o. Thus, we shall require two delay lines of lengths L_i and L_o for the two MFC's and a delay line of length L_i for the pulse stretcher in order to make the detector of the combined code. The total number of delay units required will be $2L_i + L_o$, whereas in the case of an equivalent single code of length $L_i \times L_o$ a total of $L_i \times L_o$ delay units are required. As with the single code of length L, the combined code of length $L = L_i \times L_o$ will improve the output SNR of the MFC by a factor of L.

The SS technique finds application in antijamming and multiplexed and multiaccess systems. In multiaccess and multiplexing application, a code is to be assigned to every subscriber in the network and therefore, a large number of code-words are necessary. The interference in such a situation is caused by other subscribers and to minimise it, the code words are required to be uncorrelated. In an antijamming application, the interference is of unknown nature and can be considered to be uncorrelated with the code used in the system. We first discuss the correlation properties of the combined codes for multiaccess and multiplex applications.

It has been shown that the cross correlation R_{xy}^{c} between two combined code words depends on the correlation R_{mp}^{i} between the inner code words and the correlation R_{nq}^{o} between the outer code words and is given by

$$R_{xy}^{c} = R_{mp}^{i} \times R_{nq}^{o} \qquad .. \quad (5)$$

Thus, it is seen that the correlation properties of the component codes are quite important in deciding the correlation properties of the combined codes. If the orthogonal codes with M_i and M_o inner and outer code-words respectively are chosen as the component codes, the resultant combined code is also an orthogonal code having $M_i \times M_o$ code words.

If the component codes are transorthogonal, however, then the resultant combined codes are not transorthogonal. The cross-correlation peaks may be as high as the auto-correlation peaks of the component codes. Out of a total of $M_i \times M_o$ combined code words, M_o (if $M_o < M_i$) words will, however, have a comparatively smaller correlation.

For instance it has been shown that with a Reed Muller code of word-length 8 as the inner code and another orthogonal code of length 4 as the outer code a combined code of 32 code-words, each of length 32, results and the code words are orthogonal to each other. On the other hand if 15 length and 7 length maximal PN codes, which are transorthogonal, are taken as the inner and outer codes respectively then the combined code will have 105 code words but the unnormalised cross correlation values will be -15, -7 and +1 .

In some multiplexed communications applications, the different transmitters may be operating completely independently of others in which case the complete correlation function between any two code words with all possible time-shifts between them has to be studied. The Gold's sequences are quite suitable in such cases as they give $2^{l}+1$ sequences each of length $2^{l}-1$ and with a cross-correlation bound given by

$$\left| R_{xy}(\tau) \right| \leq \begin{cases} 2^{(l+1)/2} + 1 & \text{when } l \text{ is odd} \\ 2^{(l+2)/2} + 1 & \text{when } l \text{ is even} \end{cases} \qquad \ldots (6)$$

where l is an integer such that $l \neq 4m$, m being an integer. The ratio of the autocorrelation-peak to the cross-correlation-peak is approximately $2^{l/2}$ when l is large and increases fast as l is increased. Thus, the longer sequences are found to be more suitable from the view point of low cross-correlation-peaks.

The combined codes can be made using the Gold's codes as both the inner and the outer codes and the resulting simplicity in the receiver implementation is realised. The combined Gold sequences may have large cross-correlation peaks — larger than the cross-correlation peaks of an equivalent single long Gold code. Combined codes can also be formed by taking distinct PN sequences of a particular length as the inner and the outer codes for the application just mentioned above. However, the number of distinct PN sequences of a particular length are limited in number and their cross-correlation-peaks have been shown to be quite high.

In antijamming applications also the requirement of a large TB product can be achieved by using combined codes with different component codes including the Barker codes. The advantages of the Barker codes is that they have low auto-correlation function (ACF) sidelobes which is necessary when there is no synchronisation between the transmitter and the receiver as in a radar ranging problem or in asynchronous communication system. The Barker codes are not suitable in multiplexing applications as they are unique for a particular length.

The matched-filter-correlator forms the heart of the receiver in a SS system. As such a detailed study of both the analogue as well as the digital MF correlators have been made and they have been implemented also. The analogue MF correlator has been implemented with the help of tapped analogue RC-active delay-line which has been developed in the laboratory specifically for the purpose, but can be used in other applications also.

The building block of the active RC delay line is a low-pass type of delay filter which has been designed to give a large delay which is practically constant over the desired passband. At the same time, the filter has to meet the pulse shape requirements of small rise time and low overshoots.

Two different configurations of the delay filter with similar characteristics have been developed in the laboratory, where, in one configuration an inverting amplifier has been used and in the other configuration a noninverting amplifier. The analysis of the delay filters have been given in detail from which it is possible to design a filter for a particular application.

The transfer function G(s) of the delay filter with the inverting amplifier is as follows

$$G\,(s) = \frac{H}{s^2 + b_1^{'}s + b_2^{'}} \times \frac{a_o s^2 + a_1 s + a_2}{b_o s^3 + b_1 s^2 + b_2 s + b_3} \qquad ..\ (7)$$

The first factor in (7) corresponds to a two section RC-low-pass filter and the second factor corresponds to a third order active-RC filter. The constants H, $b_1^{'}$, $b_2^{'}$, ... etc. depend on the circuit parameters. The transfer function of (7) has five poles and two zeros. The poles are adjusted by controlling the amount of feedback to give the desired response. The zeros are actually nondominant and thus, play no important role. The filter with noninverting amplifier has six poles and once again has nondominant zeros. Its transfer function G(s) is given as

$$G(s) = \frac{a_0 s^4 + a_1 s^3 + a_2 s^2 + a_3 s + a_4}{b_0 s^4 + b_1 s^3 + b_2 s^2 + b_3 s + b_4} \times \frac{a_0' s^2 + a_1' s + a_2'}{b_0' s^2 + b_1' s + b_2'} \quad \text{.. (8)}$$

The first factor in (8) corresponds to a fourth order RC-active delay filter and the second factor to a second order RC-active filter. The constants a_0 ..., b_0, ... a_0', ... b_0' etc. depend on the circuit parameters. The frequency response, phase response and the pulse response of both the filters have been investigated. These filters have been made with discrete components but can be implemented in the thick/thin film form as an integrated circuit. These active-RC-delay-lines are found to be superior to their LC counterpart in many ways. A summary of the performance of the two delay filters have been given in Table-I.

Table-I : Performance of the inverting and the noninverting delay filters.

	Noninverting type	Inverting type
3 db Bandwidth	2.5 KHz	3.3 KHz
Gain in the passband	0.73	-2
Delay, t_{d1}	192 Microsecs.	166 Microsecs.
Rise time, t_{r1}	152 Microsecs.	120 Microsecs.
Overshoot	14.38%	19%
Undershoot	15.75%	14%

The figure of merit, defined as the ratio of time-delay per section t_{d1} to rise-time per section t_{r1}, of the RC-active-delay-line developed has been found to be 1.38 and is attainable for frequencies in the range of 500 KHz with proper scaling of the component values. The number of delay sections z per unit, and the number of units Z in the delay line are related to the figure of merit in the following manner

$$\frac{t_{d1}}{t_{r1}} = \frac{2\ Z^{1/3}}{z^{2/3}} \qquad .. \quad (9)$$

From this relation, where the final output rise-time has been taken to be half the input pulse width, the number of sections per delay unit for a 15 length line is found to be 7. In matched filter applications we find that a lesser number of sections per delay unit can be used with a little loss in the performance. Thus, in the laboratory a 15 length delay line has been made with only two sections per delay unit.

It is to be noted that as the length of the delay line is increased, the number of delay sections per delay unit also increases to keep the distortion in the last output pulse within tolerable limits. This, therefore, results in unmanageable size if the length of the code is long requiring long delay lines. Particularly, from this view point, it can be seen that if combined codes are used, then the length of the individual delay line for the inner and the outer codes are small and an analogue implementation is in the realm of easy implementation.

The digital MF correlator is realised with the help of shift registers using binaries. The input signal-plus-noise is hard limited before it is fed to the shift register. It has been found experimentally that with four samples per bit the loss in the SNR due to the digital processing was negligible whereas only two samples per bit required a 2.14 db greater input SNR.

Synchronous and asynchronous operation of the MF correlators have been studied. The asynchronous operation, though inferior by 6 db to the synchronous one, is inevitable where synchronisation may not be available.

The effectiveness of the combined codes has been tested in the laboratory by making a MF correlator using the RC-active-delay-lines mentioned above for a combined code of length 15 x 7 where the inner and the outer codes are the maximal PN sequences of length 15 and 7 respectively. The practical results have been found to be close to the theoretical results within experimental limits.

An experimental system using combined code has also been made to enable the transmission of teleprinter signals of bandwidth 50 Hz over a 3 KHz telephone channel and thus, requiring a bandwidth expansion of 60. This expansion of the time-bandwidth product by 60 has been achieved by using a 15 x 4 length combined code where the inner code is a 15 length maximal PN sequence and the outer code is a 4 length Barker sequence. The performance of the MF correlator, which consists of a two stage RC-active-delay-line as mentioned above and a nonlinear pulse stretcher, has been tested in presence of Gaussian noise. The final print out on the receive-teleprinter showed that the character probability of error was 10^{-3} with an input SNR of -2.1 db whereas theoretically the required SNR should have been only -9.7 db. The experimental results have been taken with a nonlinear pulse stretcher and without a subpulse-matched-filter. A linear pulse stretcher would improve the SNR by 2 db and the subpulse-matched-filter by another 3 db. With 10 simultaneous users, the input SNR would be -9.54 db and thus, the system described here can theoretically cater for 10 simultaneous users with a character error probability of 10^{-3}.

Studies here have been confined to the processing of the signal at the video level only because of the simplicity of implementation and experimentation. The MF correlator at the RF level is difficult to implement but the problem is easily solved by bringing down the RF signal to the video level through the use of the I-&-Q-channel detection of the RF signal to avoid any losses and doing the subsequent processing at the video level. The use of I & Q channels would double the size of the MFC correlator. If, however, only a single-channel-processing is done, there may be an average signal loss of 3 db. Theoretically, the I & Q channel detection of the RF signal and the subsequent video processing is equivalent in performance to the RF processing and detection.

Finally, a comparison of the various processors viz analogue or digital, synchronous or asynchronous, using a single or combined code, has been made with reference to a multiaccess system using spread-spectrum-technique for 20 simultaneous users. The results are presented in Table-II.

Table-II : Comparison of MFC Processors for 20 simultaneous users and $P_e = 3 \times 10^{-5}$

Sl. No.	Code	Type of Processor	Operation	Required Code Length L	No. of delay units required
1	Single code	Analogue	Synchronous	152	152
2	Single code	Digital	Synchronous	241	964
3	Single code	Analogue	Asynchronous	608	608
4	Single code	Digital	Asynchronous	964	3856
5	Combined code	Analogue with linear stretcher	Synchronous	152 (8 x 19)	35
6	Combined code	Analogue with nonlinear stretcher	Synchronous	241 (13x19)	32
7	Combined code	Digital	Synchronous	403 (13x31)	176

It is interesting to observe that for a bit-error-probability of 3×10^{-5} and the number of simultaneous users $K = 20$, an all analogue MF correlator using combined code requires a total number of 35 delay units having individual delay lines of length 8 and 19 only whereas a digital MF correlator would require 176 delay units. On the other hand if single long codes were used the analogue processor would require 152 delay units in one delay line and the corresponding digital processor would require 964 units.

In conclusion it can be said that the efficacy of combined codes in spread-spectrum applications has been amply demonstrated. These codes can replace single long codes in many multiaccess and multiplex situations most advantageously — particularly in the great simplification of the hardware at the receiver. In some applications of multiaccessing when the transmitters are completely independent of each other, a compromise has to be struck between the advantages of the receiver simplification through the use of the combined codes and the disadvantages of larger cross-correlation-peaks between the code words. Matched-filter-correlators for multiplex applications of the combined codes have been built satisfactorily using RC-active-delay-lines developed for the purpose. These delay lines can be used in wide variety of applications including equalisers, digital communications etc. It has also been shown that spread-spectrum-technique results in suppression of impulse noise in addition to providing immunity to interference and jamming.