

For real values of the Nakagami fading parameter m , the integral in eqn. 9 can be easily computed numerically. However, for integer values of m , a closed form result may be obtained. In this case, it can be shown that the hypergeometric function in eqn. 5 can be written as (see Appendix of [9]).

$${}_2F_1\left(1, m + \frac{1}{2}; m + 1; x\right) = \frac{1}{\binom{2m}{m}\left(\frac{x}{4}\right)^m} \left[\frac{1}{\sqrt{1-x}} - \sum_{k=0}^{m-1} \binom{2k}{k} \left(\frac{x}{4}\right)^k \right] \quad (11)$$

Further, using integration by parts in eqn. 6, and the fact that [8]

$$\cos^{2n}(x) = \frac{1}{2^{2k}} \left\{ \binom{2n}{n} + \sum_{k=0}^{n-1} \binom{2n}{k} \cos[2(n-k)x] \right\}$$

Q_3 may be evaluated in closed form [4]. After some simplification the result is

$$P_e(M) = \frac{M-1}{M} - \beta_U \left\{ \left(\frac{1}{2} + \frac{\tan^{-1} \alpha_U}{\pi} \right) \times \sum_{k=0}^{m-1} \left(\frac{m/4}{\alpha\gamma_0 \sin^2(\pi/M) + m} \right)^k \times \left[\binom{2k}{k} + \frac{1}{\pi} \sum_{l=0}^{k-1} \binom{2k}{l} \frac{\sin[2(k-l)\tan^{-1} \alpha_U]}{(k-l)} \right] \right\} \quad (12)$$

where $\alpha_U = \sqrt{\{\alpha\gamma_0/[\alpha\gamma_0 \sin^2(\pi/M) + m]\} \cos(\pi/M)}$ and $\beta_U = \sqrt{\{\alpha\gamma_0/[\alpha\gamma_0 \sin^2(\pi/M) + m]\} \sin(\pi/M)}$. Fig. 1 shows the average symbol error rate for several values of m and M .

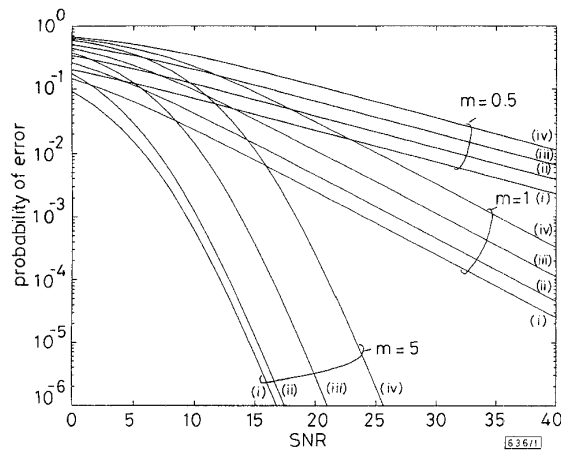


Fig. 1 Average symbol error rate for MPSK in Nakagami fading channel

- (i) $M = 2$
- (ii) $M = 4$
- (iii) $M = 8$
- (iv) $M = 16$

Effect of diversity: The results can be easily extended to include the effect of diversity on system performance. When N -order MRC diversity is employed at the receiver to mitigate the effect of multipath fading, the PDF of the output SNR is given by

$$f(\gamma) = \left(\frac{m}{\gamma_0}\right)^{Nm} \frac{\gamma^{Nm-1}}{\Gamma(Nm)} \exp\left(-\frac{m}{\gamma_0} \gamma\right) \quad (13)$$

For real values of the Nakagami fading parameter m , the result becomes

$$P_e(M) = \frac{2\Gamma(2Nm)(m/4)^{Nm}}{\Gamma(Nm)\Gamma(Nm+1)} \left\{ \frac{\sqrt{\alpha\gamma_0} \sin \theta_U}{(\alpha\gamma_0 \sin^2 \theta_U + m)^{Nm+\frac{1}{2}}} \times {}_2F_1\left(1, Nm + \frac{1}{2}; Nm + 1; \frac{m}{\alpha\gamma_0 \sin^2 \theta_U + m}\right) - \frac{Nm}{\pi} \left(\frac{1}{\alpha\gamma_0 + m}\right)^{Nm} \int_0^1 \frac{t^{Nm-1} v(t)}{\sqrt{1-t}} \tan^{-1}(v(t) \tan \theta_U) dt \right\} \quad (14)$$

while for integer values of m , we have

$$P_e(M) = \frac{M-1}{M} - \beta_U \left\{ \left(\frac{1}{2} + \frac{\tan^{-1} \alpha_U}{\pi} \right) \times \sum_{k=0}^{Nm-1} \left(\frac{m/4}{\alpha\gamma_0 \sin^2(\pi/M) + m} \right)^k \times \left[\binom{2k}{k} + \frac{1}{\pi} \sum_{l=0}^{k-1} \binom{2k}{l} \frac{\sin[2(k-l)\tan^{-1} \alpha_U]}{(k-l)} \right] \right\} \quad (15)$$

Conclusion: We have derived new expressions for the symbol error rate performance of coherent M -ary phase shift keyed signals in slow nonselective Nakagami fading and additive white Gaussian noise channel. For integer values of the Nakagami fading parameter m , a closed-form expression is derived for the symbol error rate. However, for real values of m , a simple formula which can be easily computed numerically is presented. The effect of N -order MRC diversity reception on the MPSK system is also considered. The results presented are sufficiently general to offer a convenient method to evaluate the performance of digital land mobile coherent MPSK systems.

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Differential matched filter architecture for spread spectrum communication systems

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Indexing terms: Spread spectrum communication, Matched filters

The authors present a new digital matched filter architecture: digital differential matched filter (DDMF), that employs novel schemes to reduce the number of multiplications and accumulations (M and A). Theoretical analysis shows that the DDMF saves half of the M and A hardware in comparison with the conventional filter, and maintains identical processing gain. This makes the proposed DDMF more suitable for direct sequence spread spectrum (DSSS) communication systems and low power VLSI implementations.

Introduction: The direct sequence spread spectrum communication system has many attractive properties compared with other communication techniques. The most well known properties are its anti-jamming capability, multipath rejection, low probability of intercept, etc. [1]. In those systems the despreading of random code is an important issue [1]. Among many despreading algorithms, the use of a matched filter is supposed to be a fast way to acquire the random code [2]. However, the major disadvantage of conventional digital matched filters is that as the number of stages increases the amount of multiplication and accumulation (M and A) will be greatly increased. This constrains the chip rate and the hardware implementations. The number of stages of commercial digital matched filters is mostly in the range 8–64, and the chip rate is limited to <30 M chip/s [3].

We propose a digital differential matched filter (DDMF), which employs novel schemes to reduce the amount of M and A. For those matched filters with long stages, this new architecture saves half the number of M and A in comparison with the conventional filter, while maintaining an identical processing gain (PG). By cutting down the number of M and A, not only the hardware implementation but also the power consumption is reduced. This makes digital matched filters more suitable for low power personal communication system (PCS) implementation.

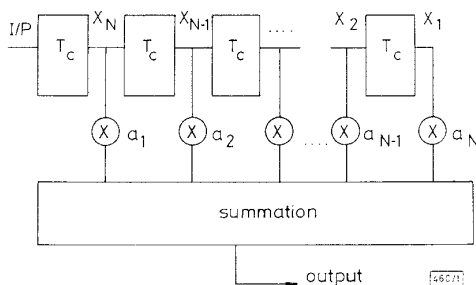


Fig. 1 Conventional matched filter structure

Architecture: A block diagram of the conventional digital matched filter (CDMF) with N stages is shown in Fig. 1 [1], where a block with T_c is a tap with one chip time delay. The output of the matched filter at time $n-1$ and n can thus be expressed as follows:

$$f_{C,n-1} = a_N x_0 + a_{N-1} x_1 + \dots + a_2 x_{N-2} + a_1 x_{N-1} \quad (1)$$

$$f_{C,n} = a_N x_1 + a_{N-1} x_2 + \dots + a_2 x_{N-1} + a_1 x_N$$

where N is the length of the PN sequence, $a_{i,i} = 1 \sim N$ are the PN sequence coefficients, and $x_{i,i} = 0 \sim N$ is the received digital signal. By subtracting two consecutive outputs of this matched filter, a new sequence $f_{D,i} = f_{C,i} - f_{C,i+1}$, $i = 1, 2, \dots$ can be generated, where

$$\begin{aligned} f_{D,n} &= b_{N+1} x_0 + b_N x_1 + b_{N-1} x_2 + \dots + b_2 x_{N-1} + b_1 x_N \\ &= -a_N x_0 + (a_N - a_{N-1}) x_1 + \dots + (a_2 - a_1) x_{N-1} + a_1 x_n \end{aligned} \quad (2)$$

Apparently, by accumulating the $f_{D,i}$, $i = 1, 2, \dots, n$, $f_{C,n}$ can be obtained. Therefore, a DDMF structure, as shown in Fig. 2, can be constructed. Since a_i , $i = 1, 2, \dots, N$ are either 1 or -1 , the b_i , $i = 1, 2, \dots, N$ in DDMF are 2, -2 , or 0 except b_1 and b_{N+1} . For a coefficient of zero, there is no need for multiplication. Thus the number of multiplication is reduced. Table 1 is a comparison of the number of multiplications in CDMF and in DDMF. In this comparison, a well known maximum length random code (m-sequence) [1] is used.

Table 1: Comparison of M&A number between conventional and proposed matched filter

	CDMF	DDMF	Ratio of M&A number between CDMF and DDMF
Number of M&A	$2^r - 1$	$2^{r-1} + 1$	$2^{r-1} + 1 / 2^r - 1 \approx 0.5$

r : order of generator polynomial of m-sequence [1]

It is obvious from the comparison that the number of M and A is reduced by one half. For a system with M sample/chip, the

DDMF furthermore reduces the number of M and A to $1/2M$ times. Thus the power and the hardware are reduced accordingly in a real implementation, while the original property of CDMF, such as processing gain, is still retained.

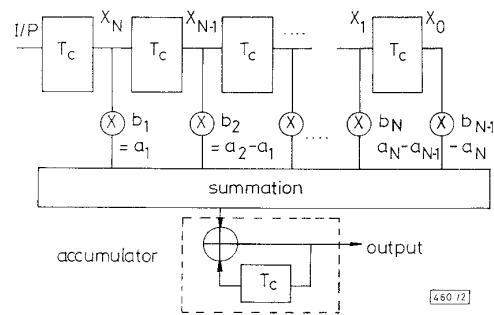


Fig. 2 Proposed differential matched filter structure with half the coefficients equal to zero

Conclusion: A differential digital matched filter for DSSS is proposed. This filter reduces the number of M and A by half for one sample/chip system, and to $1/2M$ for M sample/chip system, while maintaining all the properties of the conventional matched filter. These advantages will be more apparent for matched filters with long stages, thus leading to a more suitable implementation of low power VLSI in long operation time handset application.

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Improving the performance of cell-loss recovery in ATM networks

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Indexing terms: Forward error correction, Asynchronous transfer mode

A new method for improving the performance of cell-loss recovery using FEC (forward error correction) in ATM networks is proposed. This method provides more correcting coverage than existing methods.

Introduction: The major source of errors in high-speed networks such as B-ISDN is buffer overflow during congested conditions which results in cell loss. Conventional cell loss recovery methods using FEC in ATM networks recover up to 16 consecutive cell losses because lost cells are recognised from a 4 bit sequence number (SN). A cell loss recovery method which can recover up to 18 consecutive cell losses in ATM networks is presented in [1]. This means that the method cannot extend the row length of the coding matrix to more than 18. We review the method briefly and then employ a new algorithm for recovering more cell losses. The performance estimation is based on the two-state Markov model