THE EFFECT OF TIME DELAY ON THE PERFORMANCE OF DIVERSITY COMBINING SYSTEMS FOR HIGH-LEVEL QAM DIGITAL MICROWAVE RADIO SYSTEMS

by

Jose R. Coll M.

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ABSTRACT

The effect of time delay difference ocurring between the received radio signals on the performance of second-order space diversity combining systems (SOSDCS) using quadrature amplitude modulation (QAM) schemes is studied. Radio frequency (RF), intermediate frequency (IF) and baseband (BB) combining are considered.

Models to represent the combining systems affected by the aforementioned phenomenon are developed and analyzed to determine the impairment mechanism and the effect of corrective devices that can be incorporated in the system.

Computer simulation of the above models was employed to determine the performance of 64 and 256 QAM SOSDCS, with or without corrective operations incorporated.

It has been found that when no corrective devices are incorporated in the system, small values of time delay difference between the RF received signals render inoperable RF and IF 64 and 256 QAM SOSDCS, while the degradation produced when combining is made at baseband, is significantly lower.

A similar performance as in BB combining can be achieved in RF and IF combining by employing high precision phase compensation.

ACKNOWLEDGEMENTS

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Special thanks are due to his wife and children for their understanding and encouragement during this work.

CONTENTS

CHAPTER III

LIST OF ABBREVIATIONS

AGC	Amplitude gain control
ATDL	Adaptive time delay line
APSK	Amplitude and phase shift keying
AWGN	Additive white Gaussian noise
BB	Baseband
BER	Bit error rate
BW	Bandwidth
C	Carrier power
D/C	Down converter
E_b	Energy per bit
E_s	Energy per symbol
erfc	Complementary error function
FFT	Fast Fourier transform
f_c	Carrier frequency
f_b	Bit rate
f_{IF}	IF frequency
f_s	Symbol rate
GHz	Gigahertz
I	In-phase
IF	Intermediate frequency
ISI	Inter-Symbol-Interference
Mb/s	Megabit per second
N_o	Noise power density
N	Noise power
NRZ	Non return to zero

ns	Nano-second
P(e)	Probability of error
P_b(e)	Bit error probability
P_s(e)	Symbol error probability
PSK	Phase shift keying
Q	Quadrature
QAM	Quadrature amplitude modulation
QPSK	Quadrature phase shift keying
RF	Radio frequency
SFAPS	Single frequency adaptive phase shifter
SOSDCS	Second-order space diversity combining system
T_b	Bit duration
T_s	Symbol duration
WAPS	Wideband adaptive phase shifter
α	Raised cosine filter roll-off factor
β	Phase error
ϕ	Band-center-frequency phase
τ	Time delay difference

LIST OF FIGURES

CHAPTER I

INTRODUCTION

The number of bits per Hertz that a microwave radio system is able to transmit has become a topic of considerable interest in recent years. In order to increase the number of bits that can be assigned to one symbol, multilevel modulation schemes such as quadrature phase shift keying (QPSK), 8 phase shift keying (8PSK) and 16-ary quadrature amplitude modulation (16 QAM), have been developed [1-4].

Today higher-level modulation schemes, such as 32 and 64 QAM, are already in use and 256 QAM is under development [5-10].

Radio systems using this kind of modulation are quite sensitive to manufacturing imperfections and channel perturbations [11-12]. Therefore, any countermeasure that can be implemented to improve the system performance is highly desirable.

Diversity combining is a technique that has been used to significantly improve the performance and availability of microwave radio systems. Concepts and principles of diversity combining techniques, mainly applied to analog systems, have been investigated since 1950 [13,14]. It is well known that in second-order diversity combining systems (SODCS), a carrier-to-noise ratio (C/N) up to 3 dB higher than in uncombined systems can be theoretically obtained [12].

When diversity combining techniques are applied to digital systems, problems related to time variable delay difference between the RF received signals, caused by various fading phenomena, can become extremely important when high-level QAM modulation schemes are employed [12].

In this thesis, the degradation effect of a time delay difference between the RF received signals is studied for RF, IF and BB QAM SOSDCS. The performance improvement caused by using phase difference compensation in RF and IF combining systems is also studied. Two phase compensation methods to improve IF combining system performance are presented and demonstrated to be theoretically equivalent.

Space diversity combining was selected for this study given that it has been found to provide better performance than frequency diversity when higher RF frequency systems are considered [16]. RF and IF space diversity switching are not recommendable for digital radio systems because of the excessive number of interruptions that would be produced during switching, which in turn may be caused by fading. Space diversity switching can be made at baseband but this is not an economical solution since two complete receiver-demodulators are required [16].

Although our study is restricted to second-order diversity combining systems and specific results are presented only for those systems using 64 and 256 QAM modulation; concepts are applicable to n-order diversity combining and M-ary QAM modulation schemes in general.

1.1 THESIS OUTLINE

The general principle of diversity combining and equations to compute error probability of ideal 64 and 256 QAM second-order space diversity combining systems (SOSDCS), are presented in Chapter II.

In Chapter III a theoretical analysis of performance degradation caused by a time delay difference ocurring between the RF received signals in RF, IF and BB QAM SOSDCS is presented followed by the determination, by means of computer simulation, of this performance degradation in 64 and 256 QAM SOSDCS. Performance improvement achieved by means of phase compensation is also studied in Chapter III.

In Chapter IV the computer programs employed to simulate 90 Mb/s - 4 GHz - 64 and 256 QAM SOSDCS, under different conditions of time delay impairment, with or without the effect of corrective measures, are described.

Chapter V contains recommended further research.

Chapter VI contains conclusions.

CHAPTER II

IDEAL COMBINING OF 64 AND 256 QAM SECOND-ORDER SPACE DIVERSITY SIGNALS

2.1 GENERAL PRINCIPLE OF SOSDCS

The advantage of combining two noisy signals, with the same desired information content, but arriving at the receive end of a radio system through two different path which are separated in space is well known and lies in the fact that the noise signal affecting the information signal can normally be considered as uncorrelated; therefore, the combiner adds the information signals coherently (on a voltage basis), the noise signals incoherently (on a power basis) and an equivalent carrier-to-noise (C/N) improvement as high as 3 dB can be theoretically obtained [12,13].

2.2 EQUATIONS TO COMPUTE BIT ERROR PROBABILITY $P_b(e)$ IN RECTANGULAR 64 and 256 QAM SOSDCS

The theoretical $P_b(e)$ performance of 64 and 256 QAM systems using rectangular signals sets can be derived following the standard procedure employed for M-ary QAM systems in general [17,18,19,20,21]. The corresponding equations are found to be:

For 64 QAM systems:

$$P_b(e) = (7/24)\ \mathrm{erfc}\ \sqrt{(1/7)(E_b/N_o)} \qquad (2.1)$$

For 256 QAM systems:

$$P_b(e) = (15/64)\ \mathrm{erfc}\ \sqrt{(4/85)(E_b/N_o)} \qquad (2.2)$$

Where $P_b(e)$ is bit error probability, erfc(x) is the complementary error function, E_b is the average signal energy per bit and N_o is AWGN power density at the receiver filter input.

To appreciate the improvement effect that can be provided by a second-order diversity combiner, let us consider the model illustrated in figure 2.1, in which, under ideal conditions, we will assume that:

$s_1(t) = s_2(t) = s(t)$: Information signal

$n_1(t)$, $n_2(t)$: Gaussian, uncorrelated and equal-power noise signals

The combined information signal $s_c(t)$ is given by:

$$s_c(t) = s_1(t) + s_2(t) = 2\ s(t) \tag{2.3}$$

The total noise $n_c(t)$ can be expressed:

$$n_c(t) = n_1(t) + n_2(t) \tag{2.4}$$

To determine the error probability, let us note that $n_c(t)$ is also gaussian [22]. Then:

$$\sigma_c^2 = \sigma_1^2 + \sigma_2^2 = 2\sigma^2 \tag{2.5}$$

Where:

$\sigma_1^2 = \sigma_2^2 = \sigma^2$ Average power in $n_1(t)$ or $n_2(t)$

and:

σ_c^2 = Average power in $n_c(t)$

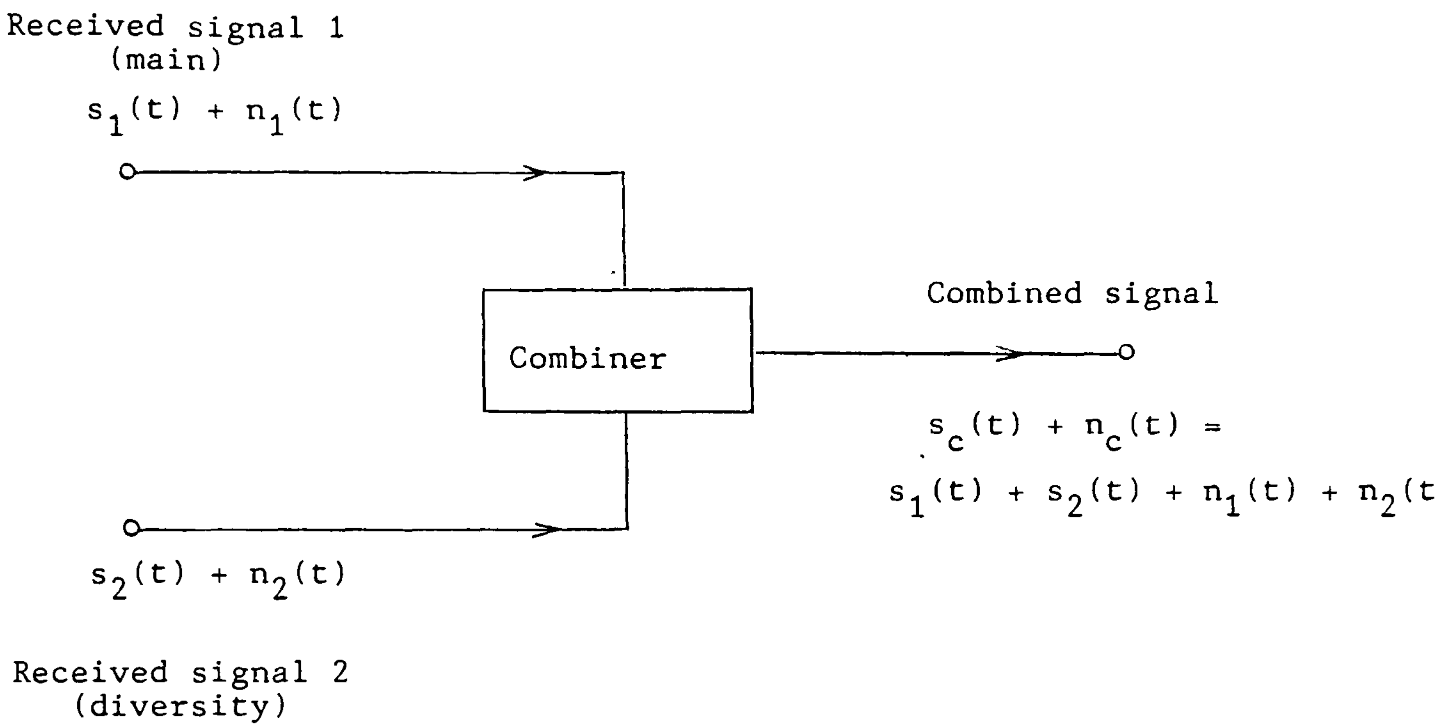


Fig. 2.1 Second order combiner model

From equations 2.3 and 2.5, the combined system can be interpreted, in relation to the uncombined system, as having a half distance between levels, say 2A intead of A, and double noise power $2\sigma^2$ in the decision instant. The probability 2p that the noise signal exceeds in magnitude one half the distance between levels is therefore:

$$2p = \operatorname{erfc}\left(2A/\sqrt{2\sigma^2}\sqrt{2}\right) = \operatorname{erfc}\left(A/\sigma\right) \qquad (2.6)$$

Using p in the derivation of $P_b(e)$ for 64 and 256 QAM, the following equations for the combined systems are found:

<u>64 QAM</u>

$$P_b(e) = (7/24)\operatorname{erfc}\sqrt{(2/7)(E_b/N_o)} \qquad (2.7)$$

<u>256 QAM</u>

$$P_b(e) = (15/64)\operatorname{erfc}\sqrt{(8/85)(E_b/N_o)} \qquad (2.8)$$

By comparing these equations with those obtained for the uncombined systems, an equivalent E_b/N_o improvement of 3 dB is noted as expected.

$P_b(e)$ as a function of E_b/N_o for 64 and 256 QAM uncombined and combined systems are plotted in figure 2.2, where the 3 dB equivalent improvement in E_b/N_o can be verified. (The corresponding curves for 4 QAM and 16 QAM are also shown in figure 2.2).

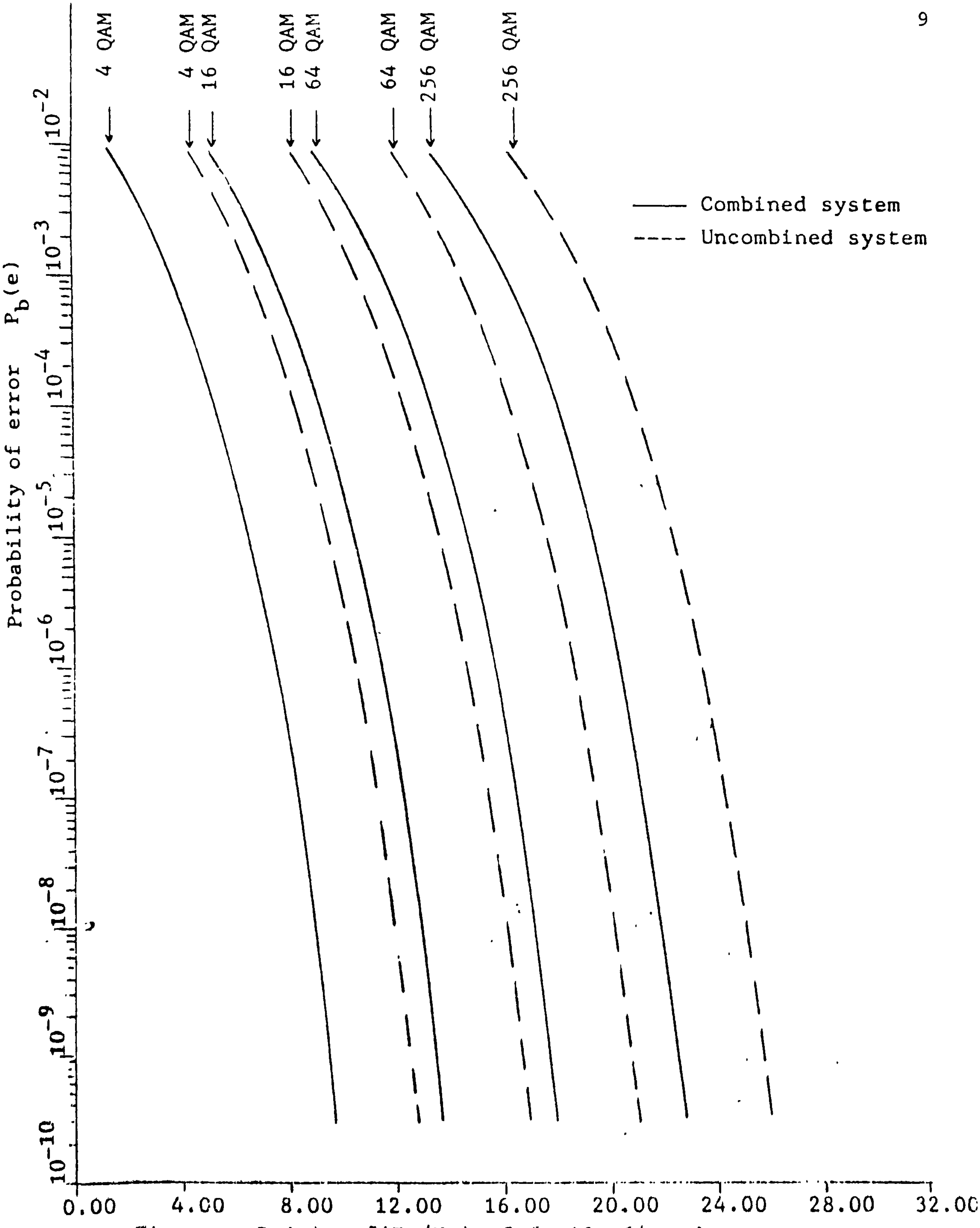


Fig. 2.2 $P_b(e) = f(E_b/N_o)$ of 4, 16, 64 and 256 QAM ideal second-order combined and uncombined systems.

$P_b(e)$ can be expressed in terms of carrier-to-noise ratio in a given noise bandwidth $(C/N)_{bw}$, employing the conversion formula [23]:

$$(C/N)_{bw} = (E_b/N_o)/(f_b/BW) \qquad (2.9)$$

Where f_b is the bit rate and BW is the bandwidth in which noise is measured.

CHAPTER III

RF, IF and BB COMBINING OF TWO RELATIVELY TIME DELAYED QAM DIVERSITY RECEIVED SIGNALS

Time variable propagation delay differences between the two RF received rays can occur in practical second-order space diversity combining systems (SOSDCS) [24,25,26] and considerably degrade their performance even in the abcense of multipath fading, unless some correction operation, such as dynamic phase compensation, is included in the system [26].

Feher and Chan [27] have shown that a small amount of time delay difference between the two inputs of a four-phase coherent phase shift keying (PSK) combining system, causes considerable degradation. More sensitive systems, like those using 64 or 256 QAM, degrade even more under the same environmental conditions.

In this Chapter, we investigate the mechanism of degradation in RF, IF and BB 64 and 256 QAM SOSDCS when a time delay difference occurs between the two RF received signals. A computer simulation-aided theoretical approach will be followed.

3.1 THEORETICAL ANALYSIS OF RF, IF AND BB QAM SOSDCS WITH A TIME DELAY DIFFERENCE BETWEEN THE RF RECEIVED SIGNALS

The main and diversity rays in a SOSDCS are illustrated in figure 3.1. Absolute time delays present in those channels at a particular instant are designated by τ_m and τ_d, respectively τ_m and τ_d are, in general, time variable because of daily angle-of-arrival variations [26] and various fading phenomena [15].

In figure 3.2, the way in which time delays τ_m and τ_d affect main and diversity signals is illustrated by means of the phase transfer function associated with each channel. To analyze only the effect of time delay difference between the RF received signals, the magnitude of the transfer function in both channels is assumed to be equal to unity.

Equal time delays in the received signals do not cause any degradation in the combining system performance, hence only the time delay difference $\tau = \tau_d - \tau_m$ will be considered in our analysis. In this way, instead of the phase transfer function $\phi_m(f) = -\tau_m f$ and $\phi_d(f) = -\tau_d f$, illustrated in figure 3.2, affecting the main and diversity channels, respectively, we will consider the difference phase transfer function $\phi(f) = -\tau f$, also illustrated in figure 3.2, affecting only the diversity channel.

In figures 3.3, 3.4 and 3.5, RF, IF and BB SOSDCS models for this study are illustrated. It is supposed that the transmitting and receiving filters are square root of raised cosine, with and without x/sinx equalization, respectively, chosen to assure inter-symbol-interference (ISI) free transmission and optimal P(e) performance, i.e.: to constitute an optimal Nyquist channel when time delay difference between the RF received signals is not present in the

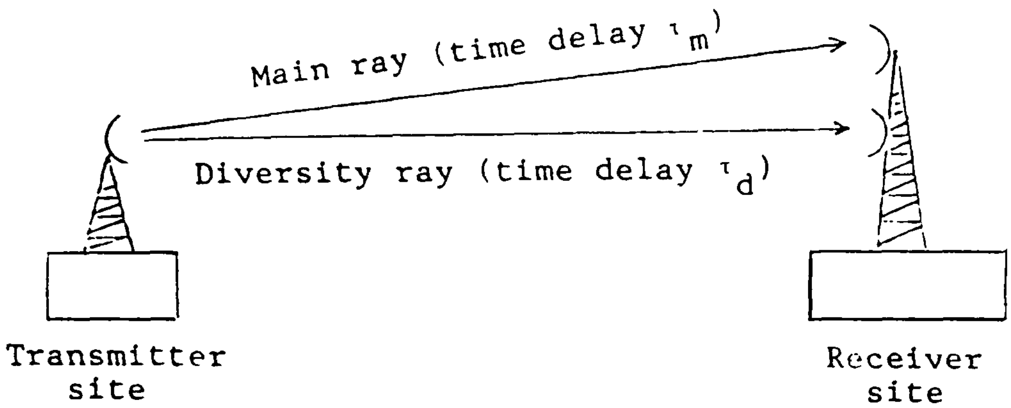


Fig. 3.1 Main and diversity rays in second-order diversity combining systems (space diversity is illustrated)

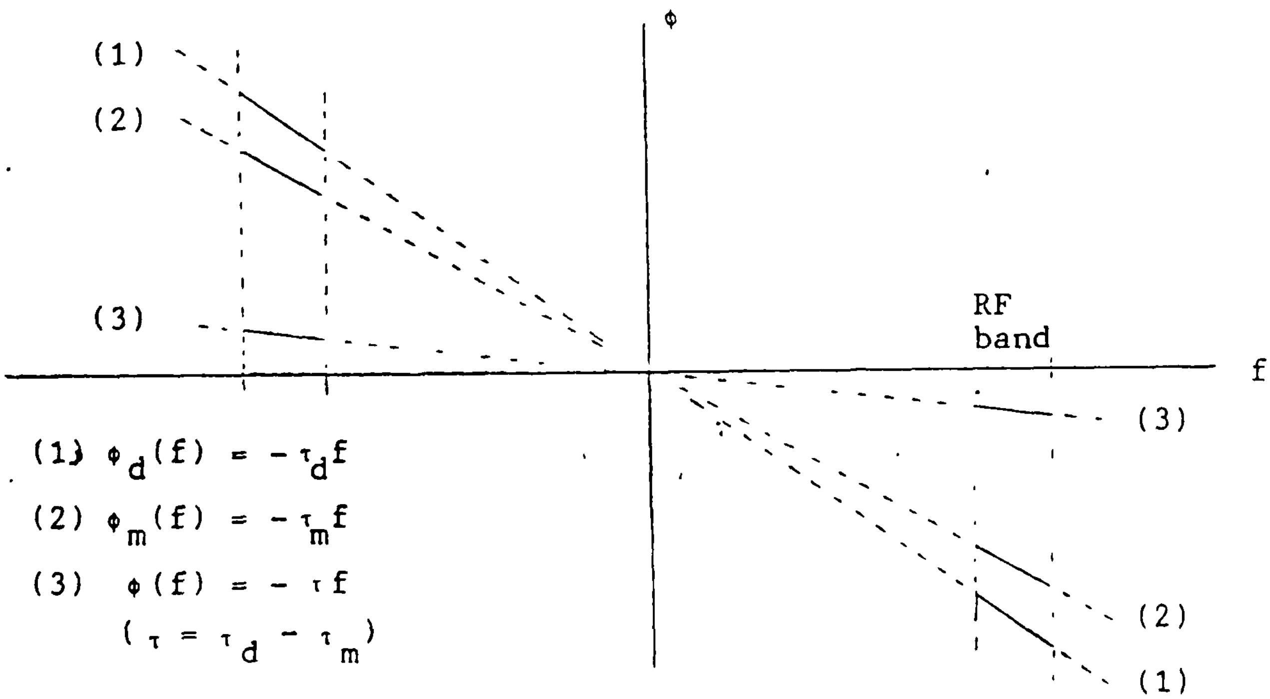


Fig. 3.2 Absolute phase transfer functions $\phi_d(f)$ and $\phi_m(f)$, in the diversity and main channels respectively, and relative phase tranfer function $\phi(f) = \phi_d(f) - \phi_m(f)$, of diversity channel with respect to the main channel.

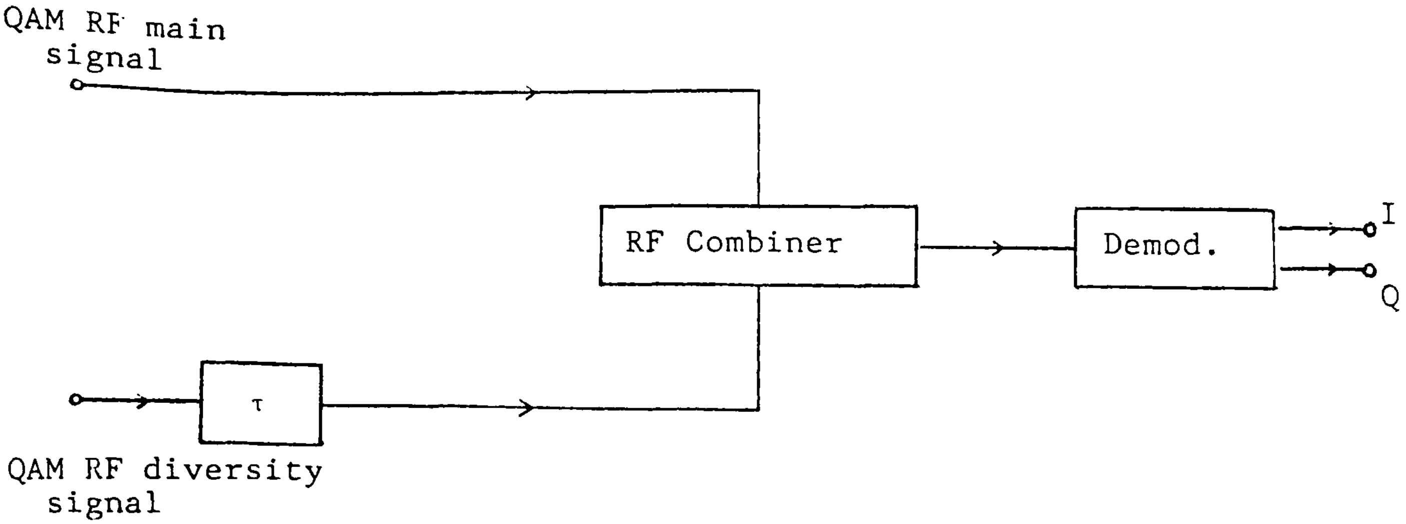


Fig. 3.3 Basic scheme of RF QAM SOSDCS

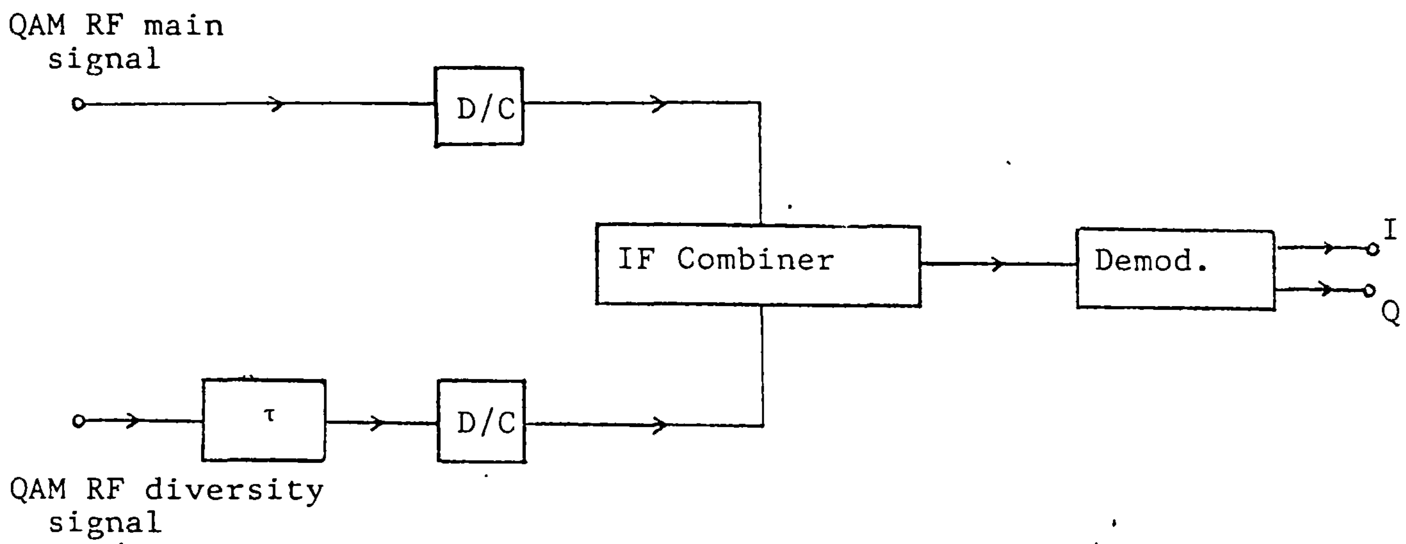


Fig. 3.4 Basic scheme of IF QAM SOSDCS

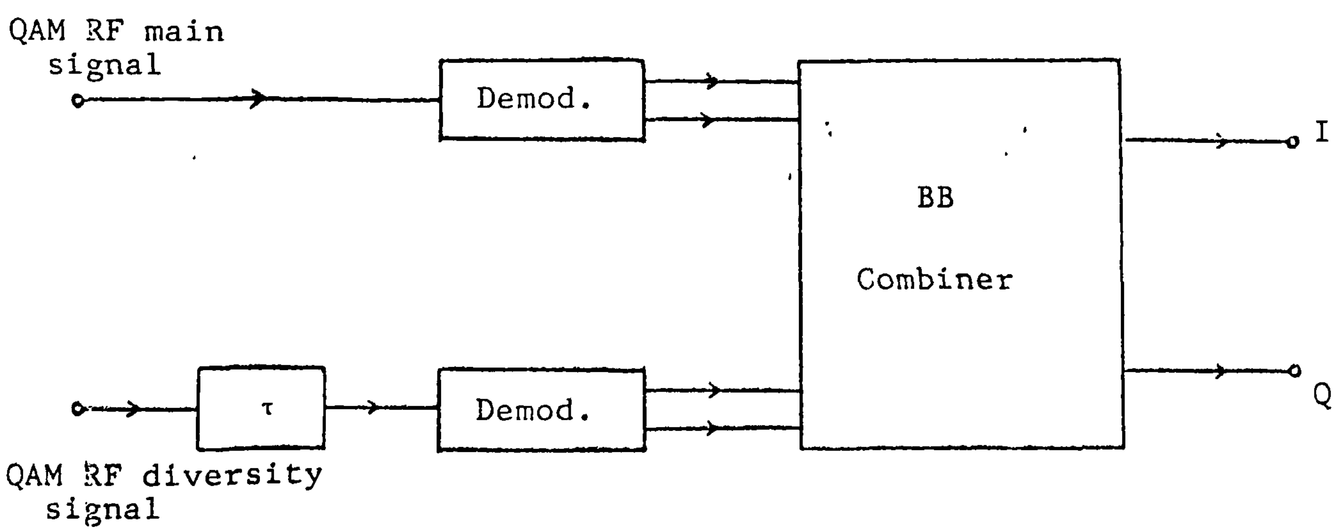


Fig.3.5 Basic scheme of BB QAM SOSDCS

system. Details about this filtering strategy can be found in reference [28].

In our RF and IF SOSDCS models, in which demodulation is made after combination, demodulation carriers recovered from one of the combining signals is supposed. For BB SOSDCS, where combination is made after demodulation, it is assumed that demodulation carriers recovered from the corresponding signal are used in each of the combining channels.

It is convenient to analyze systems of figures 3.3, 3.4 and 3.5, through their equivalent baseband models. To this purpose the complex envelope concept [29,30] is used.

The same transfer function $H_{C1}(f)$, given by equation 3.1, was found for both RF and IF SOSDCS illustrated in figures 3.3 and 3.4, to represent their equivalent baseband models between the transmitting filter output and the receiving filter input.

$$H_{C1}(f) = \begin{cases} 1 + e^{-j2\pi(f+f_c)\tau} & ; \quad -(BW/2) \leq f \leq (BW/2) \\ 0 & ; \quad \text{elsewhere} \end{cases} \qquad (3.1)$$

Where BW is the RF channel bandwidth.

For BB SOSDCS modeled in figure 3.5 the corresponding baseband model transfer function $H_{C2}(f)$ is:

$$H_{C2}(f) = \begin{cases} 1 + e^{-j2\pi f\tau} & ; \quad -(BW/2) \le f \le (BW/2) \\ 0 & ; \quad \text{elsewhere} \end{cases} \qquad (3.2)$$

In the following we analyze the performance of RF, IF and BB SOSDCS. For RF and IF combining a common analysis can be made because of their common equivalent baseband model.

3.1.1 RF and IF SOSDCS

The demodulated complex baseband signal, before regeneration, for the two systems under consideration can be written as:

$$i_o(t)+jq_o(t) = [i_i(t)+jq_i(t)]*h_T(t)*h_{C1}(t)*h_R(t) \qquad (3.3)$$

Where $i_o(t)$ and $-q_o(t)$ are demodulated signals from in-phase and quadrature channels, $i_i(t)$ and $-q_i(t)$ are $\sqrt{M}$ -level NRZ data signals at the system inputs, $h_T(t)$ and $h_R(t)$ are transmitting and receiving filter inpulse responses, $h_{C1}(t)$ is given in the frequency domain by equation 3.1 and "*" indicates convolution operation between two functions.

Given the optimum-Nyquist-channel characteristic of the convolution $h_T(t)*h_R(t) = h_N(t)$, equation 3.3 can be written:

$$i_o(t)+jq_o(t) = [i(t)+jq(t)]*h_{C1}(t) \qquad (3.4)$$

Where:

$$i(t) = i_i(t) * h_N(t) \qquad (3.5)$$

$$q(t) = q_i(t) * h_N(t) \qquad (3.6)$$

are signals with the same or proportional values as $i_i(t)$ and $q_i(t)$, respectively, at the sampling instants.

Finally, from the convolution operation indicated in equation 3.4, it can be obtained that:

$$i_o(t) = [i(t)] + [\cos \omega_c \tau\, i(t-\tau) + \sin \omega_c \tau\, q(t-\tau)] \qquad (3.7)$$

$$q_o(t) = [q(t)] + [\cos \omega_c \tau\, q(t-\tau) - \sin \omega_c \tau\, i(t-\tau)] \qquad (3.8)$$

In equations 3.7 and 3.8 the first term in the brackets can be interpreted as baseband signals from one of the combining channels and the other term in the brackets as baseband signal from the other combining channel.

At regeneration, signals $i_o(t)$ and $q_o(t)$ are sampled and regenerated.

When no time delay difference exists between the two RF received signals ($\tau=0$), there is no difference between the two contributing terms in equations 3.7 and 3.8, hence:

$$i_o(t) = 2i(t) \tag{3.9}$$

$$q_o(t) = 2q(t) \tag{3.10}$$

And, therefore:

$$i_o(t_k) = 2i_i(t_k) \tag{3.11}$$

$$q_o(t_k) = 2q_i(t_k) \tag{3.12}$$

Where t_k ($k=1,2,3,.....$) are sampling instants.

As expected for this condition ($\tau=0$), an equivalent 3 dB improvement in C/N is obtained with respect to the uncombined channel, if uncorrelated equal-power AWGN noise signals are assumed at the combiner inputs.

When $\tau \neq 0$, the signal from the diversity channel, in both, $i_o(t)$ and $q_o(t)$, is affected with respect to the one from the main channel in the following way:

1) An attenuation $\cos \omega_c \tau$ is introduced making the system performance to depend on the RF carrier frequency ω_c and the time delay τ.

2) A time delay τ is introduced.

3) An interference of the type known as "quadrature distortion" [30] is introduced.

In conclusion, when a time delay difference occurs between the RF received signals in RF or IF QAM SOSDCS, the same performance degradation is introduced and it depends, in general, on:

1) The amount of time delay difference τ

2) The RF band-center-frequency f_c

3) The filter roll-off factor α

4) The system bit rate f_b

Given this multiple dependence a further general theoretical study of performance is rather complicated. Instead, a computer simulation study is made in section 3.2.

3.1.2 BB SOSDCS

To determine the demodulated complex baseband signal before regeneration, equation 3.3 can be applied if $h_{C1}(t)$ is substituted by $h_{C2}(t)$ which is given, in the frequency domain, by equation 3.2.

Following a similar procedure as in section 3.3.3, it can be found that:

$$i_o(t) = i(t) + i(t-\tau) \qquad (3.13)$$

$$q_o(t) = q(t) + q(t-\tau) \qquad (3.14)$$

With $i_o(t)$, $q_o(t)$, $i(t)$ and $q(t)$ defined as in section 3.1.1.

It can be noted that performance dependence on RF carrier f_c as well as the quadrature interference, which were present in RF and IF combining, have been eliminated in this case and the only remaining degrading factor is the time delay τ between the two baseband combining signals.

In section 3.2 we extend this study of BB SOSDCS performance degradation, by means Of computer simulation. For the present, we can expect that disappearence of the fast periodical changing terms cos $\omega_c \tau$ and sin $\omega_c \tau$ in equations 3.13 and 3.14 causes a significantly better performance in BB SOSDCS when compared to RF and IF SOSDCS without correction. This is indeed due to the corrective effect of using demodulation carriers recovered from the respective received signals in each combining channel, which is valid in this case.

3.2 COMPUTER SIMULATION STUDY OF $P_b(e)$ PERFORMANCE IN RF, IF AND BB 64 AND 256 QAM SOSDCS WITH A TIME DELAY DIFFERENCE τ BETWEEN THE TWO RF RECEIVED SIGNALS

This study will be made employing the programs listed in appendix A and described in Chapter IV.

The simulation is based on the complex BB equivalent models presented in section 3.1. It is made for systems with an RF band-center-frequency f_c = 4 GHz and bit rate f_b = 90 Mb/s. These values are representative of practical systems. Results can be used for other values of f_c and f_b by means of normalization.

3.2.1 Computer simulation results for RF and IF SOSDCS

In figures 3.6 and 3.7, $P_b(e)$ as a function of E_b/N_o and C/N in the double sided Nyquist bandwidth [31], are presented for 64 and 256 QAM systems respectively. Values of τ close to zero were considered and it was found that for such small values of τ there is not significant performance variation when different values of α are considered.

To explain this result, let us consider equations 3.7 and 3.8, values of τ relatively small when compared to the system symbol duration T_s, can be neglected in terms $i(t-\tau)$ and $q(t-\tau)$ because they cause little change in $i(t)$ and $q(t)$. Then, equa-

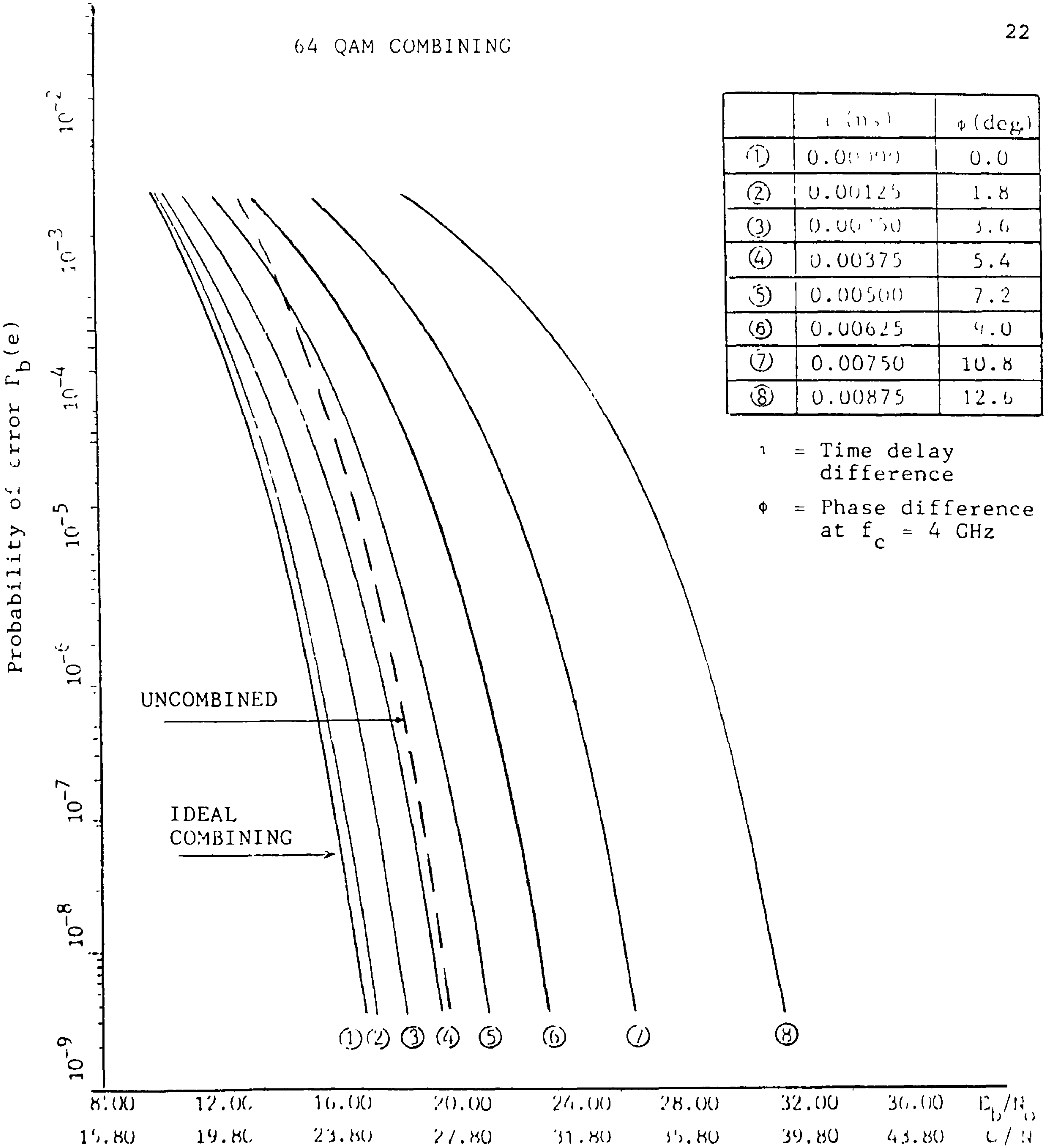
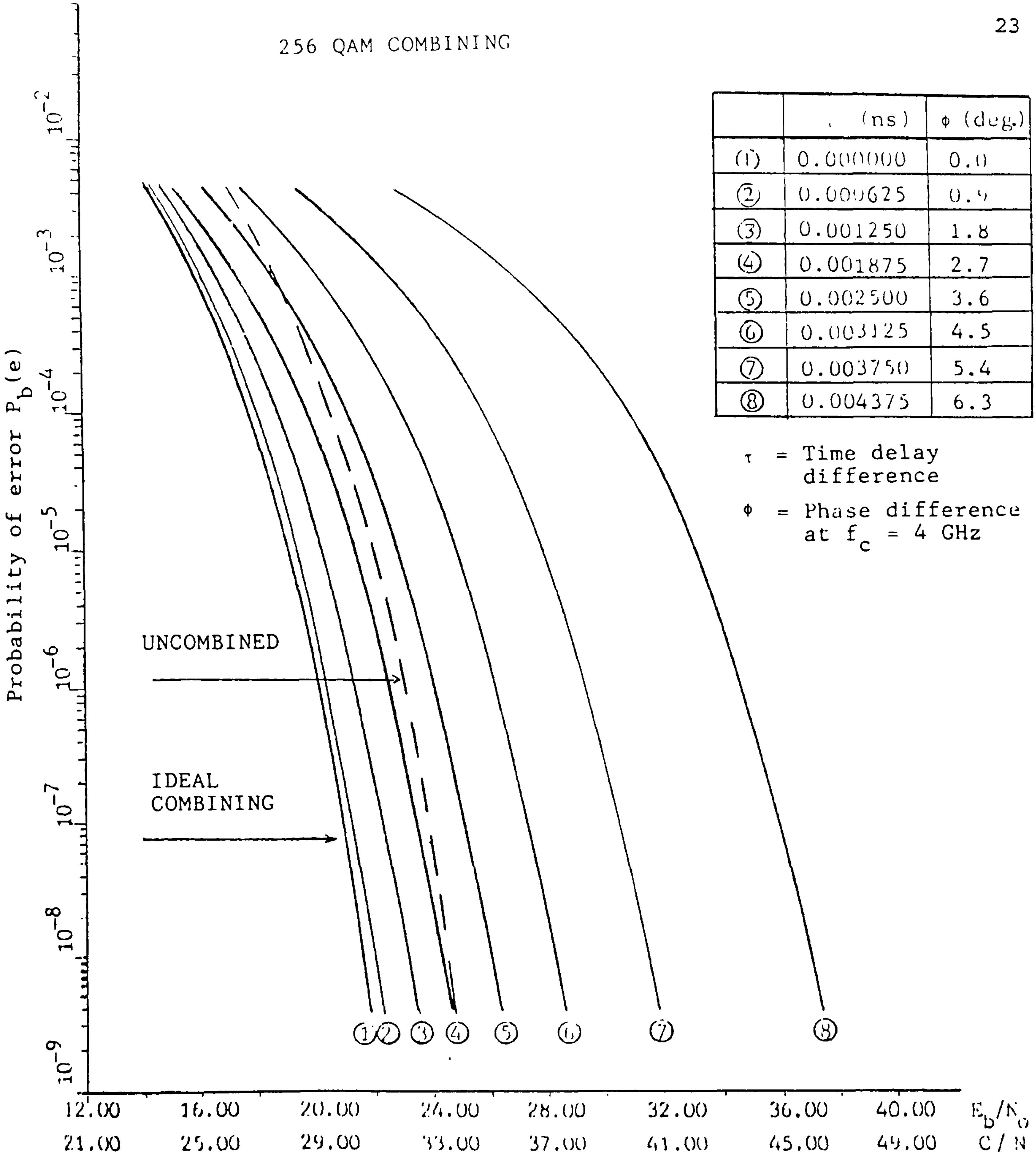


Fig. 3.6 Bit error probability performance for 64 QAM RF and IF SOSDCS with time delay difference close to 0 ns. (Without phase compensation) Nyquist filters roll-off factors: α = 0.0, 0.4, 1.0



	τ (ns)	φ (deg.)
①	0.000000	0.0
②	0.000625	0.9
③	0.001250	1.8
④	0.001875	2.7
⑤	0.002500	3.6
⑥	0.003125	4.5
⑦	0.003750	5.4
⑧	0.004375	6.3

Fig. 3.7 Bit error probability performance for 256 QAM RF and IF SOSDCS with time delay difference close to 0 ns. (Without phase compensation)
Nyquist filters roll-off factors: α = 0.0, 0.4, 1.0

tions 3.7 and 3.8, for such small values of τ, can be written:

$$i_o(t) = i(t) + \cos \omega_c \tau\, i(t) + \sin \omega_c \tau\, q(t) \qquad (3.15)$$

$$q_o(t) = q(t) + \cos \omega_c \tau\, q(t) - \sin \omega_c \tau\, i(t) \qquad (3.16)$$

And this corresponds to the suppression of the effect of other symbols on the detection of a particular one.

To illustrate this in more detail, in figures 3.8 and 3.9 we present the eye diagrams corresponding to τ = 0 ns and τ = 0.0075 ns, respectively, for the simulated 64 QAM SOSDCS with α = 1.0. In figures 3.10 and 3.11, eye diagrams corresponding to the same values of τ but using α = 0, are shown. For both values of α, the effect of τ is to compress the eye diagram (effect of factor $\cos \omega_c \tau$ in equations 3.7 and 3.8) and produce eight possible values (same values independently of α) instead of one at any sampling instant (effect of the quadrature distortion interference). But no appreciable ISI has been introduced and the system performance is practically independent of α.

To continue with our study, we can infer from figures 3.6 and 3.7, that very small values of τ cause a complete degradation in the system. Nevertheless, the appearence of periodic terms $\cos \omega_c \tau$ and $\sin \omega_c \tau$ in equations 3.7 and 3.8 suggests the possibility of some kind of periodicity respect to τ in the system performance. In effect, if values of τ such that $\omega_c \tau = k 2\pi$ (k=0, 1,2,...), are considered, equations 3.6 and 3.7 reduce to:

τ = 0.00000 ns

α = 1.0

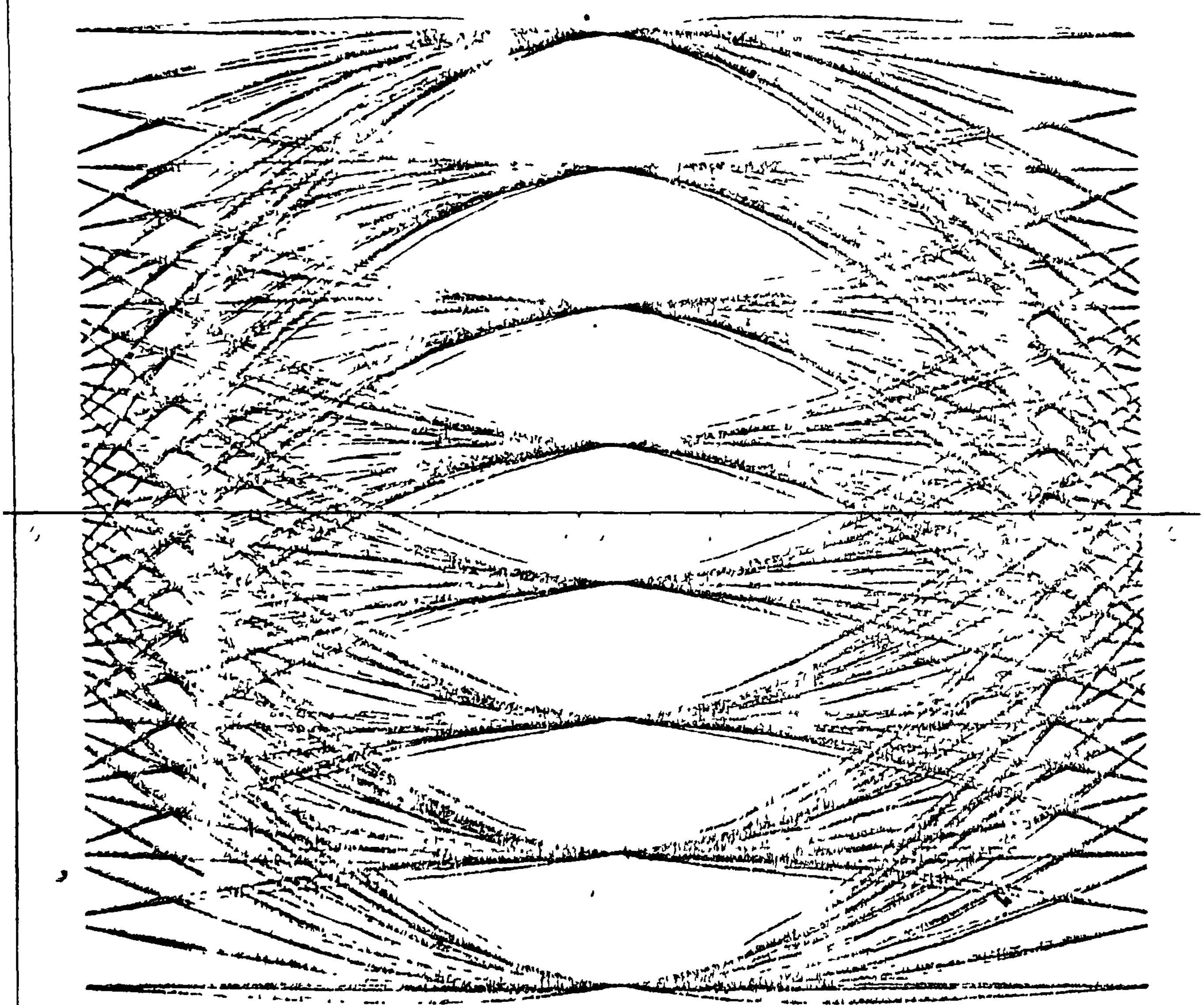


Fig. 3.8 Eye diagram for 64 QAM RF and IF SOSDCS

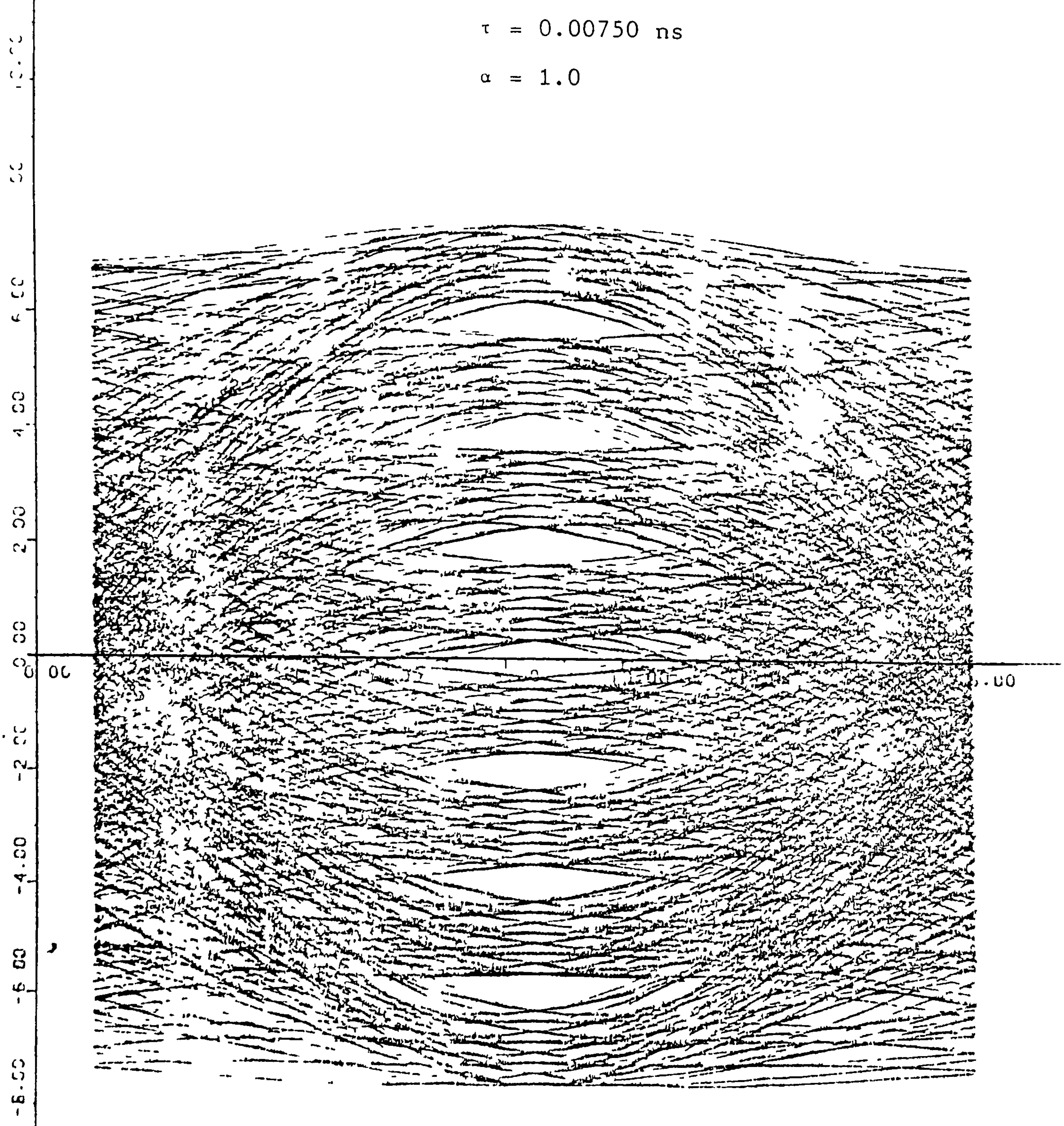


Fig. 3.9 Eye diagram for 64 QAM RF and IF SOSDCS

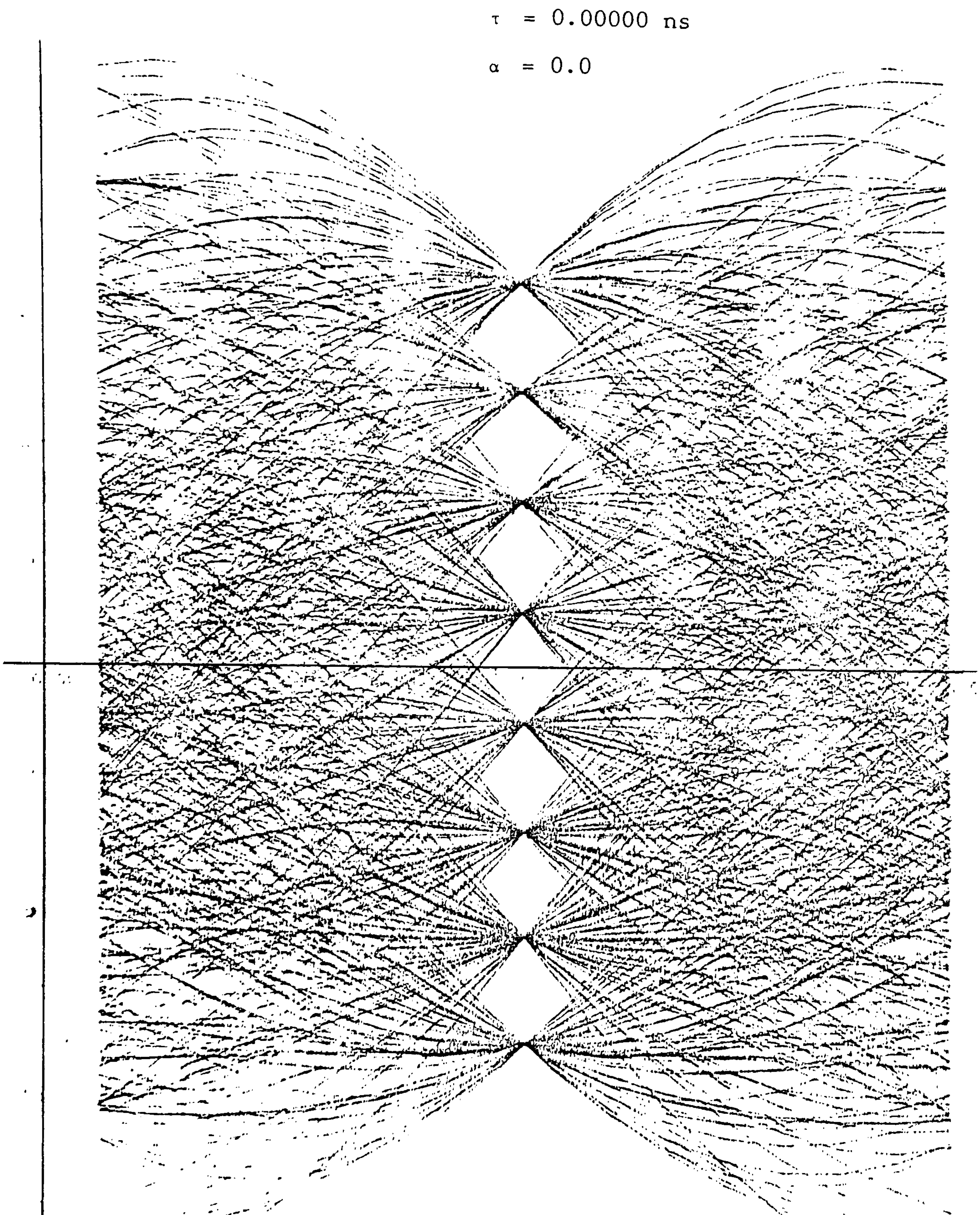


Fig. 3.10 Eye diagram for
64 QAM RF and IF
SOSDCS

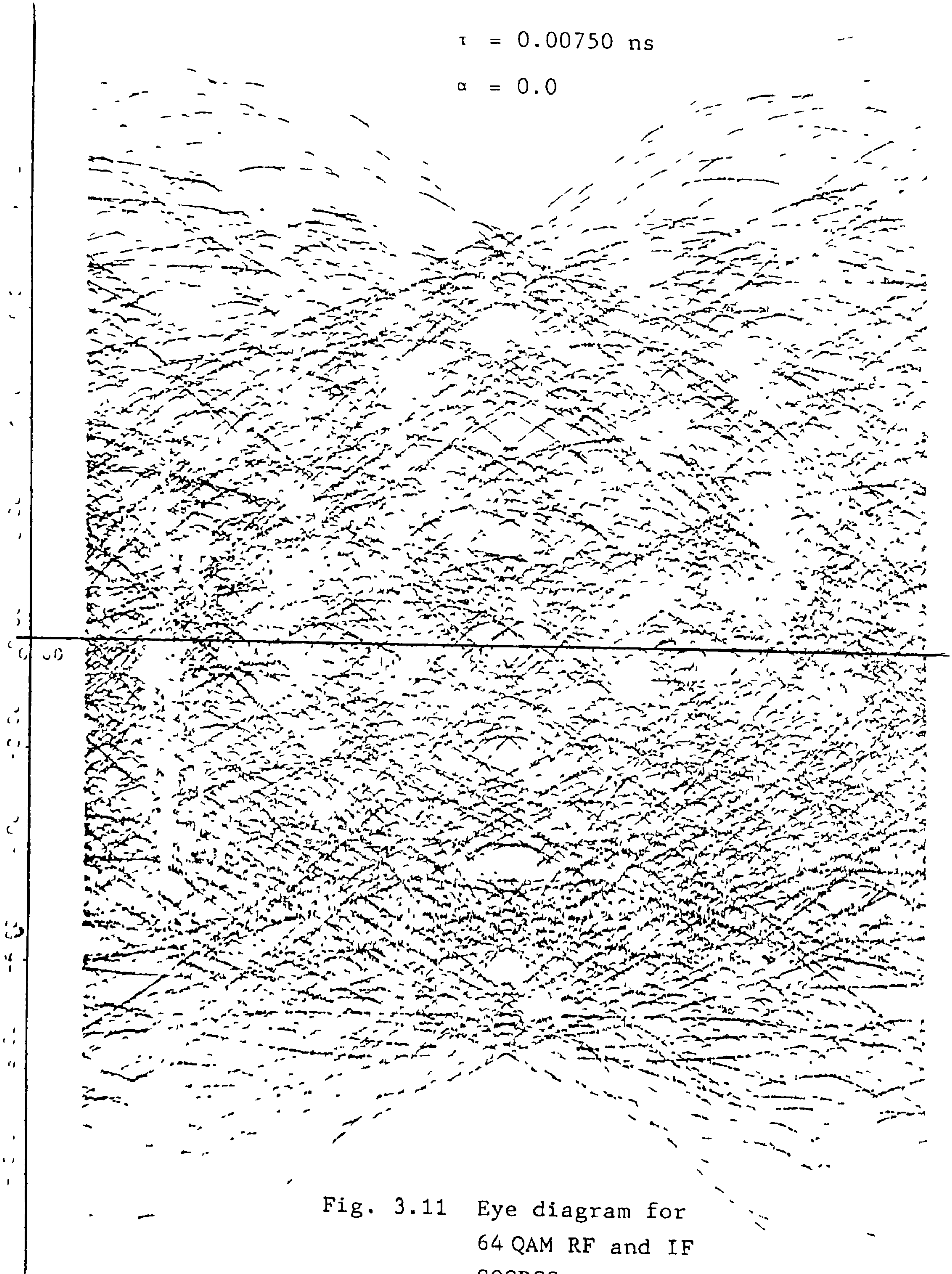


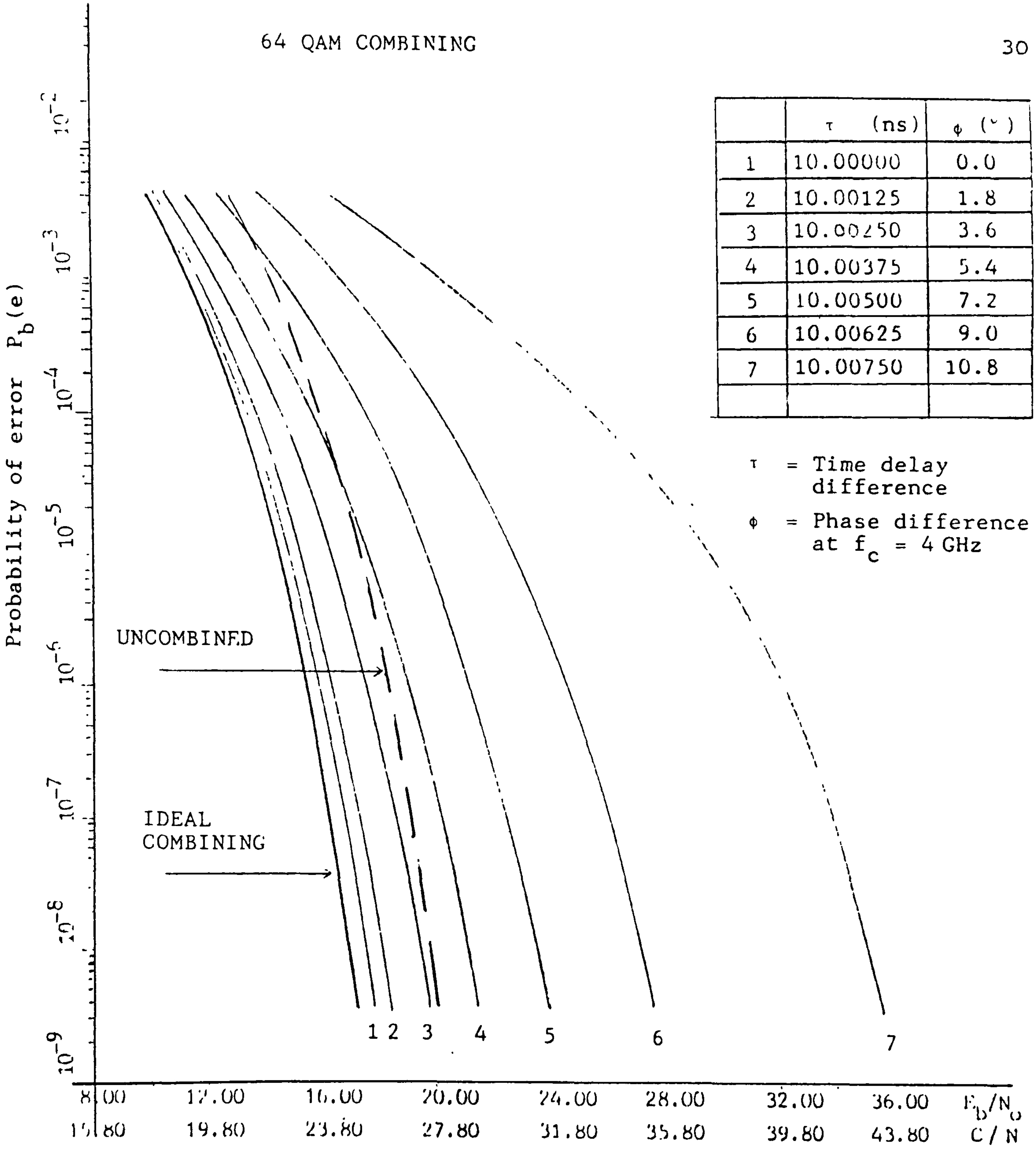
Fig. 3.11 Eye diagram for 64 QAM RF and IF SOSDCS

$$i_o(t) = i(t) + i(t-\tau) \qquad (3.17)$$

$$q_o(t) = q(t) + q(t-\tau) \qquad (3.18)$$

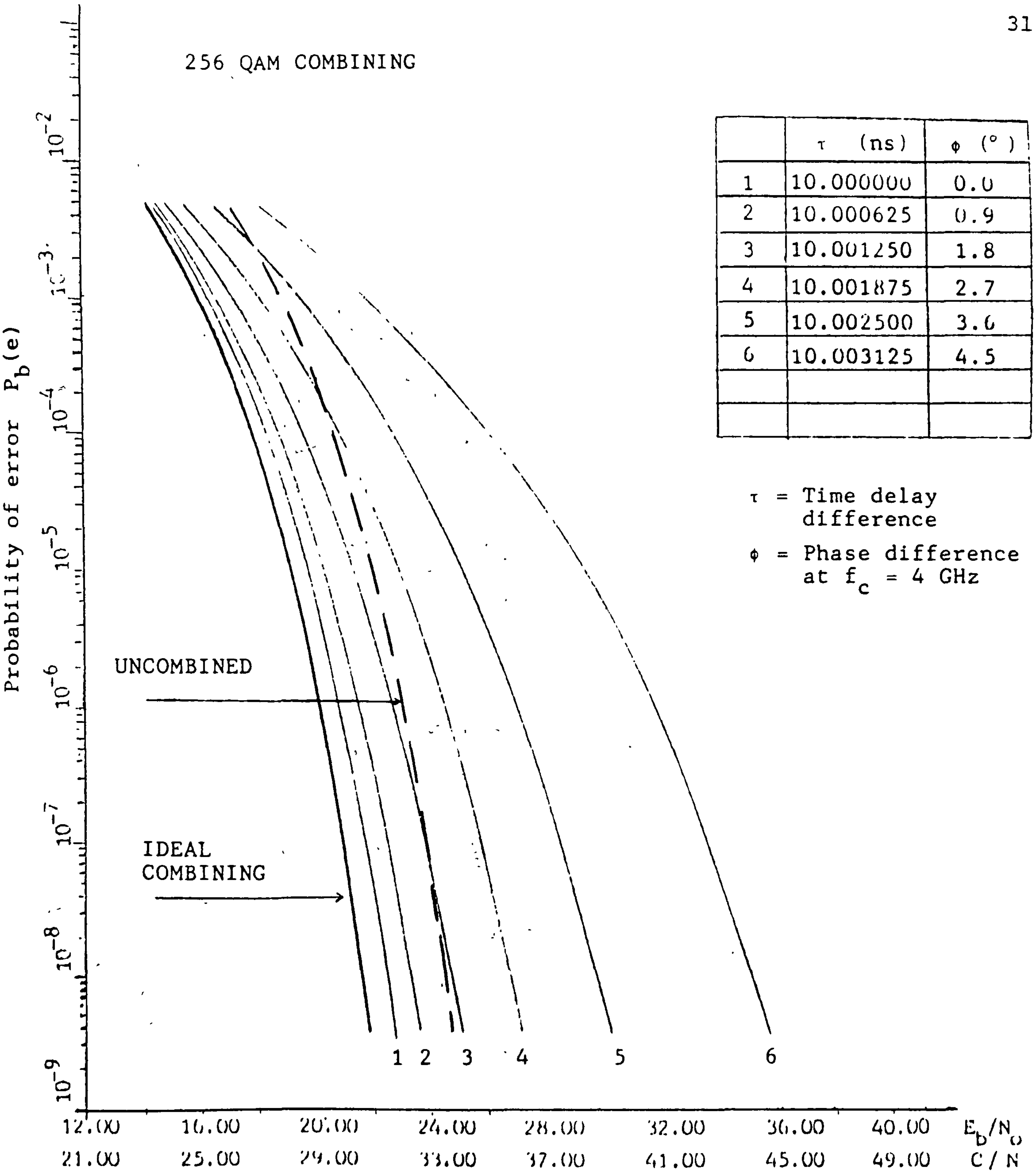
Where we can appreciate that only the delay effect of τ on $i(t)$ and $q(t)$ remains and affects the otherwise perfect periodicity in performance. This expectation was confirmed by our computer simulation when a relatively large range of values of τ was considered. A complete degradation was observed except for values of τ very close to k/f_c $(k=0,1,2,...)$, for which the system performance is gradually impaired with respect to the one observed close to $\tau = 0$ ns. Typical $P_b(e) = f(E_b/N_o)$ curves for this case are illustrated in figures 3.12 and 3.13 for $\alpha = 0.4$ and τ close to 10 ns.

In figures 3.14 and 3.15, we present the eye diagrams corresponding to $\tau = 10.0050$ ns for the 64 QAM system with $\alpha = 1.0$ and 0.0 respectively. In both diagrams the appearence of ISI, in addition to the quadrature interference effect, can be noted.



	τ (ns)	φ (°)
1	10.00000	0.0
2	10.00125	1.8
3	10.00250	3.6
4	10.00375	5.4
5	10.00500	7.2
6	10.00625	9.0
7	10.00750	10.8

Fig. 3.12 Bit error probability performance for 64 QAM RF and IF SOSDCS with time delay difference close to 10 ns. (Without phase compensation) Nyquist filters roll-off factor: $\alpha = 0.4$



	τ (ns)	φ (°)
1	10.000000	0.0
2	10.000625	0.9
3	10.001250	1.8
4	10.001875	2.7
5	10.002500	3.6
6	10.003125	4.5

Fig. 3.13 Bit error probability performance for 256 QAM RF and IF SOSDCS with time delay difference close to 10 ns. (Without phase compensation) Nyquist filters roll-off factor: $\alpha = 0.4$

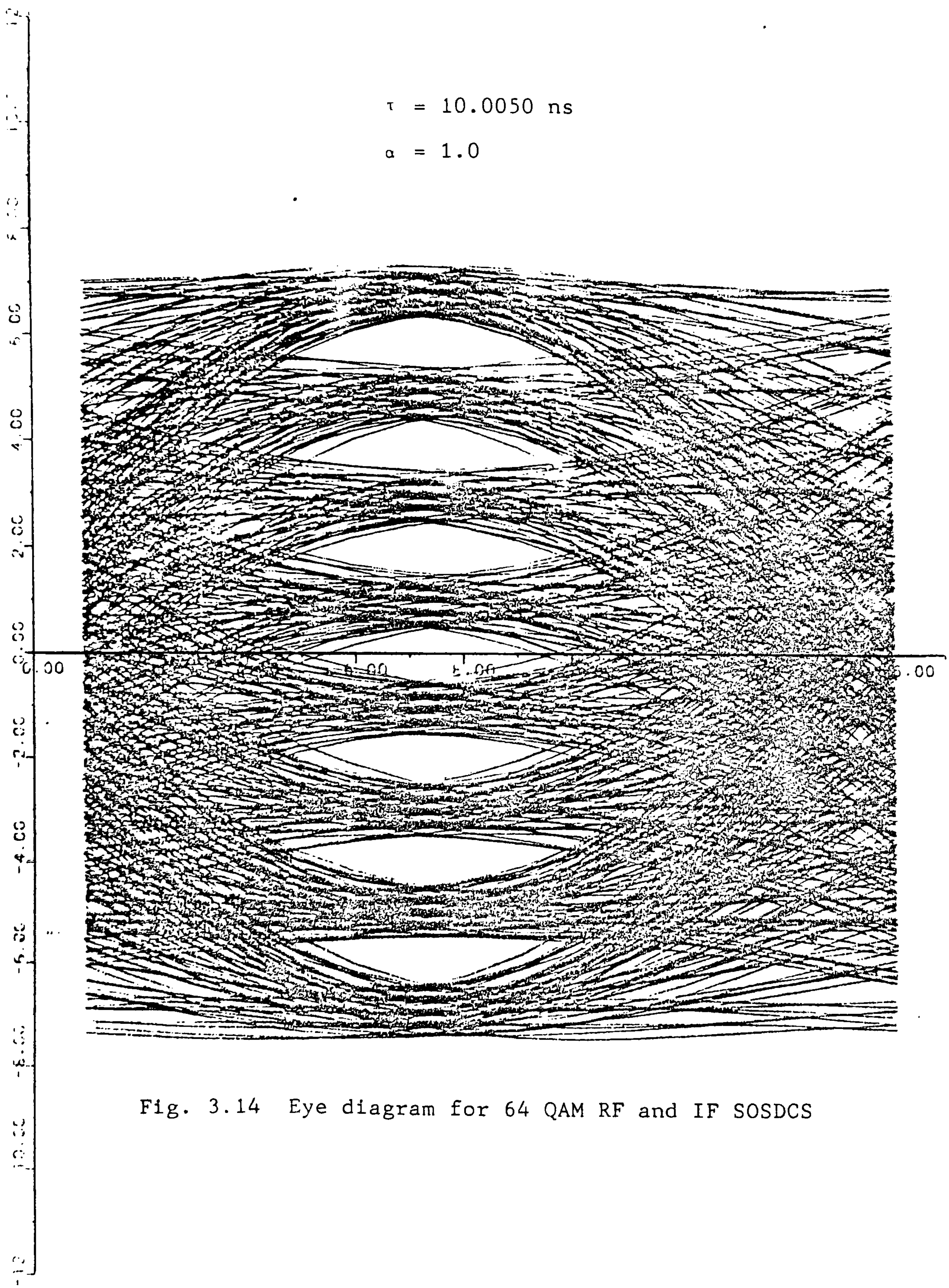


Fig. 3.14 Eye diagram for 64 QAM RF and IF SOSDCS

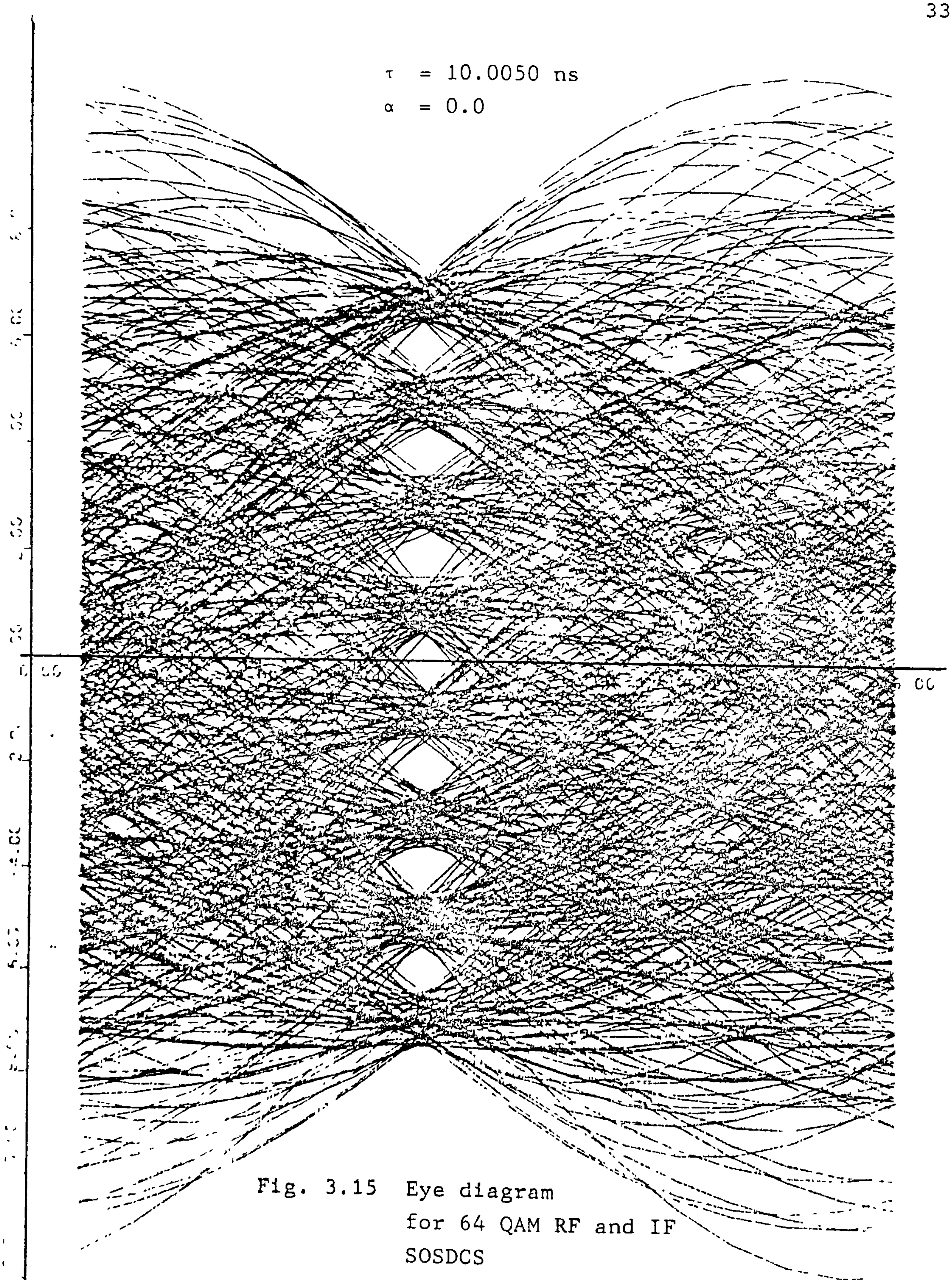


Fig. 3.15 Eye diagram for 64 QAM RF and IF SOSDCS

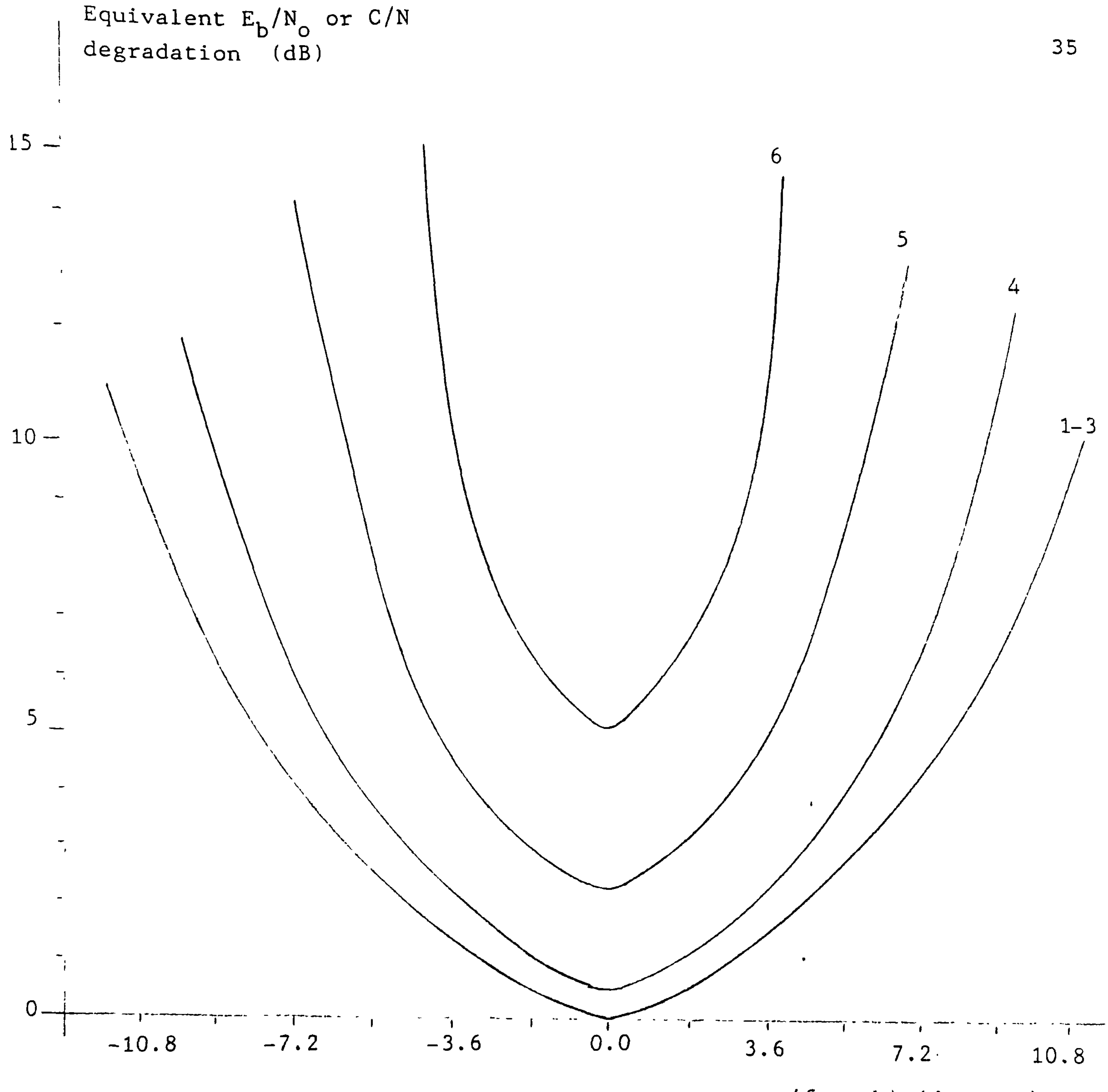
Besides, comparison of these diagrams reveals that little system performance dependence on α can be expected.

In figures 3.16 and 3.17 equivalent E_b/N_o or C/N degradation curves, at $P_b(e) = 10^{-6}$, with respect to the ideal case of combining with $\tau = 0$, are illustrated for the simulated RF and IF 64 and 256 QAM SOSDCS respectively. A value of $\alpha = 0.4$ was considered in both cases. These curves are presented as a function of the band-center-frequency phase difference $(f_c\tau - k)$ expressed in degree, where k is the integer number of carrier cycles in τ.

To illustrate how to use curves of figures 3.16 and 3.17, let us find the degradation caused by $\tau = 6.0025$ ns in our 64 and 256 QAM SOSDCS with $\alpha = 0.4$, at $P_b(e) = 10^{-6}$. In $\tau = 6.0025$ ns the entire number of periods $1/f_c = 1/4\text{GHz} = 0.25$ ns, is $k = 24$ and $(f_c\tau - k) = 0.01$ cycle $= 3.6$ degree. Using the curves of figures 3.16 and 3.17, for $k = 24$ and $(f_c\tau - k) = 3.6$ degree, it is found that the equivalent C/N degradation in the system performance is 1.7 dB for 64 QAM SOSDCS and 6 dB for 256 QAM SOSDCS.

Curves of figures 3.16 and 3.17 reveal that RF and IF 64 and 256 QAM SOSDCS are very sensitive to time delay difference between the RF received signals. Values of τ as small as 0.00375 ns cause equivalent C/N degradation, at $P_b(e) = 10^{-6}$, in the order of 3 dB and 11.5 dB in 64 and 256 QAM SOSDCS, respectively if systems with $f_c = 4$ GHz and $\alpha = 0.4$ are considered.

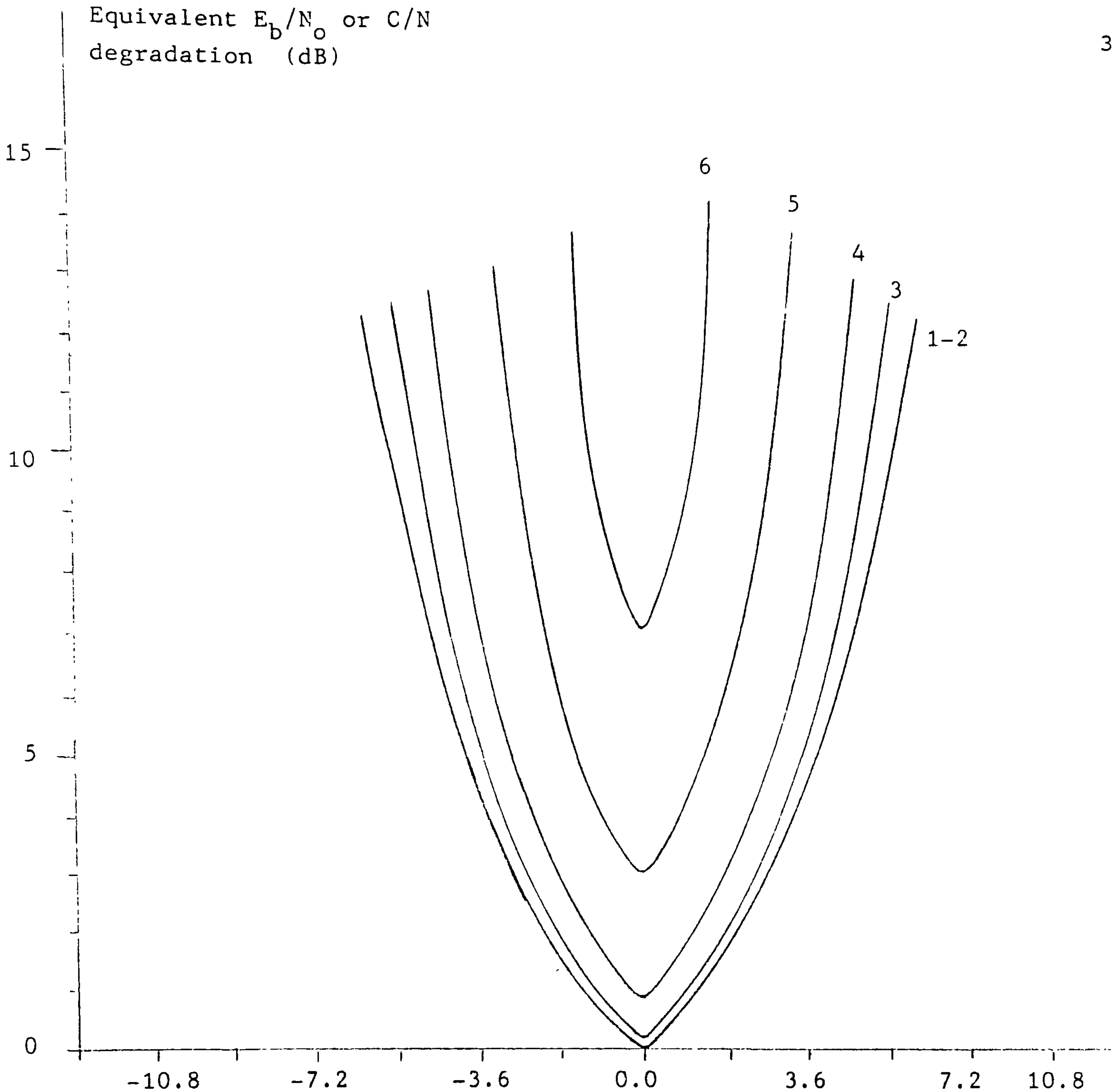
Values of τ in the order of 5 ns can be expected in some practical cases [24,25,27], which implies that, unless some kind



	k	τ (ns)
1	0	0.0
2	20	5.0
3	26	6.5

	k	τ (ns)
4	40	10.0
5	58	14.5
6	70	17.5

Fig. 3.16 Equivalent degradation of E_b/N_o or C/N, at $P_b(e) = 10^{-6}$, due to time delay difference between the RF received signals in RF and IF 64 QAM SOSDCS with f_c = 4 GHz, α = 0.4 and f_b = 90 Mb/s.



	k	τ (ns)
1	0	0.0
2	22	5.5
3	24	6.0

	k	τ (ns)
4	40	10.0
5	58	14.5
6	70	17.5

Fig. 3.17 Equivalent degradation of E_b/N_o or C/N, at $P_b(e) = 10^{-6}$, due to time delay difference between the RF received signals in RF and IF 256 QAM SOSDCS with $f_c = 4$ GHz, $\alpha = 0.4$ and $f_b = 90$ Mb/s.

of correction is made, none of these systems can be used in those cases.

Although similar curves exist in the environs of any $\tau = l/f_c$ (l=0,1,2,...), those shown in figures 3.16 and 3.17 are enough to illustrate the basic trends of the system behavior at $P_b(e) = 10^{-6}$, which are:

1) Substantial degradation exists for all values of τ except those close to l/f_c (l=0,1,2,...).

2) For relatively small values of τ (with respect to the symbol duration T_s), the system performance, around $\tau = l/f_c$, is approximately the same.

3) For relatively large values of τ (with respect to the symbol duration T_s), the system performance, around $\tau = l/f_c$, is increasingly worse.

3.2.2 Computer simulation results for baseband combining

In figures 3.18 and 3.19, $P_b(e)$ as a function of E_b/N_o and C/N in the double sided Nyquist bandwidth, are illustrated for the simulated 64 and 256 QAM SOSDCS with α = 0.4 and different values of τ.

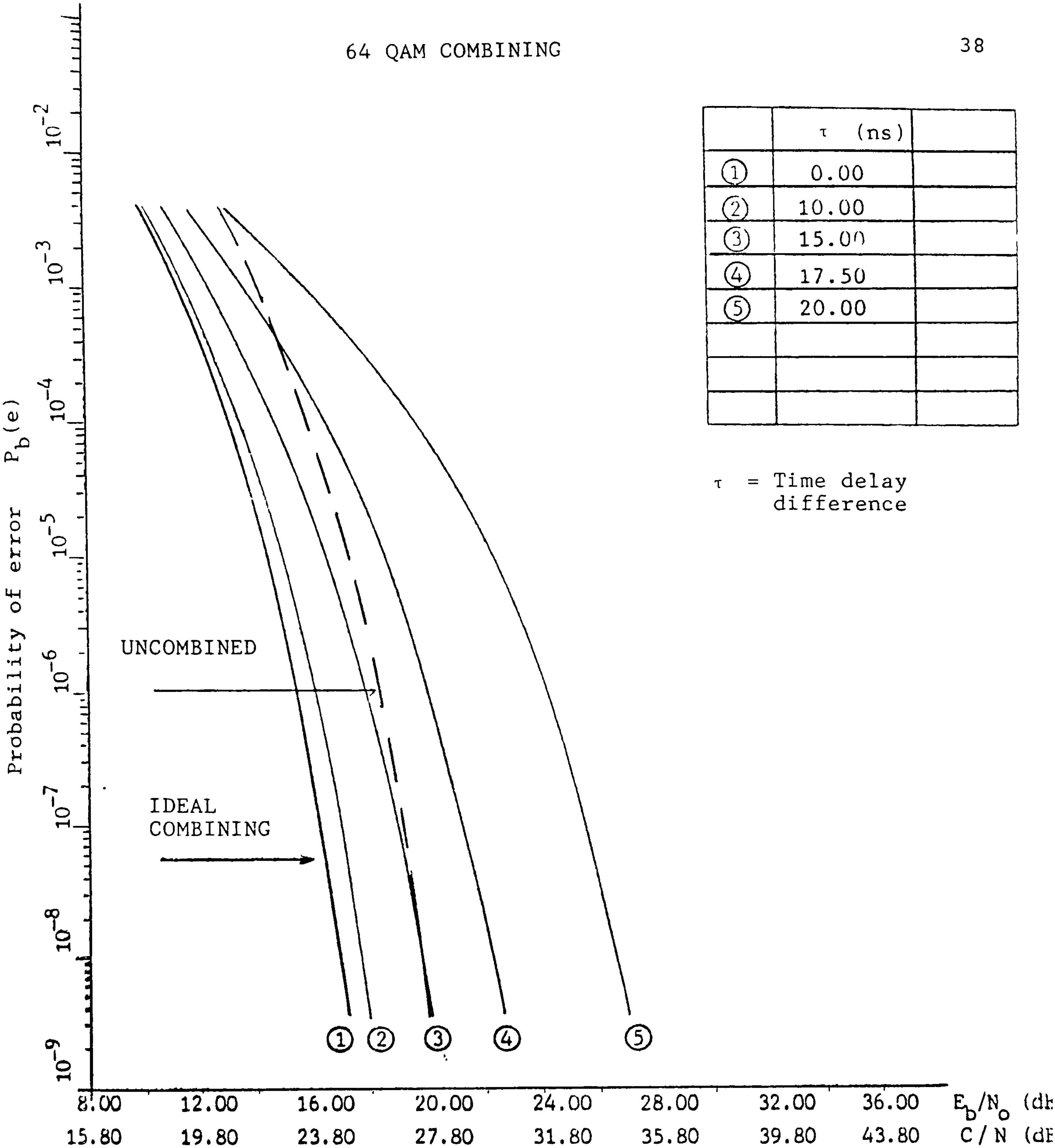


Fig. 3.18 Bit error probability performance for 64 QAM BB SOSDCS with time delay difference τ between the RF received signals. Nyquist filters roll-off factor: α = 0.4

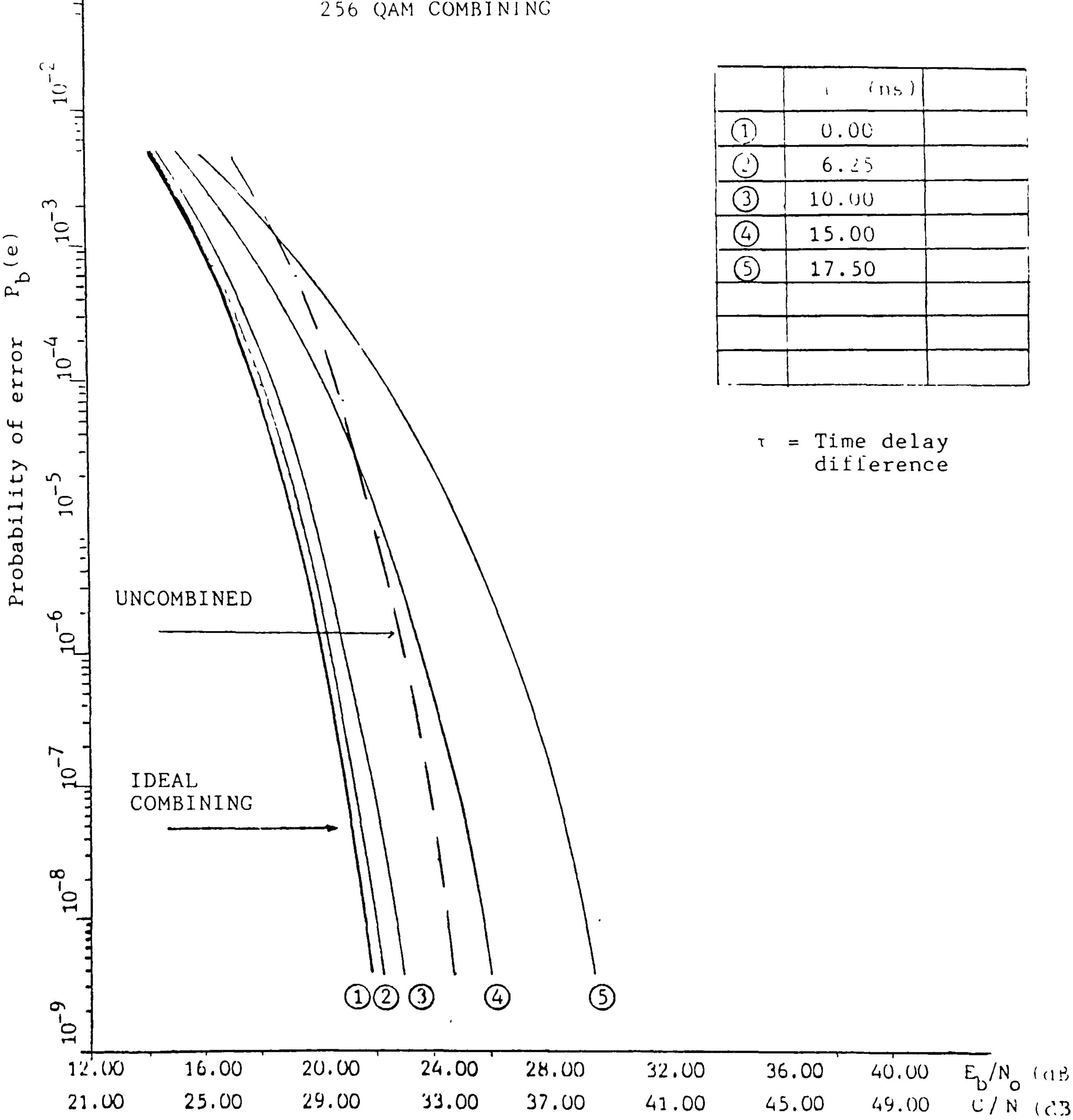


Fig. 3.19 Bit error probability performance for 256 QAM BB SOSDCS with time delay difference τ between the RF received signals. Nyquist filters roll-off factor: $\alpha = 0.4$

In figure 3.20, equivalent E_b/N_o or C/N degradation curves at $P_b(e) = 10^{-6}$ are presented.

In figures 3.21 and 3.22 we illustrate the eye diagrams corresponding to τ = 10 ns, for the 64 QAM simulated system with α = 1.0 and 0.0 respectively. The ISI nature of the impairment can be noted in both of these diagrams and a comparison of them permits to conclude that little dependence on α can be expected in the system performance.

In all the figures of this section we indicate, in addition to τ, the corresponding value of $(\tau/T_s)100\%$, in order to make them useful to analyze systems with different bit rates. In this way, it is easy to find, from curves in figure 3.20, that if, for example, expected values of τ in the system are in the order of 5 ns, it is possible to use 64 QAM BB SOSDCS with f_b as large as 170 Mb/s, without any kind of correction included, if an equivalent C/N degradation in the order of 0.5 dB at $P_b(e) = 10^{-6}$ is tolerable. The corresponding maximum value of f_b for 256 QAM BB SOSDCS is 150 Mb/s.

3.3 RF AND IF COMBINING SYSTEM PERFORMANCE IMPROVEMENT BY MEANS OF PHASE COMPENSATION

The degradation caused by a time varying delay difference between the received radio signals in RF, IF and BB SOSDCS, can be totally

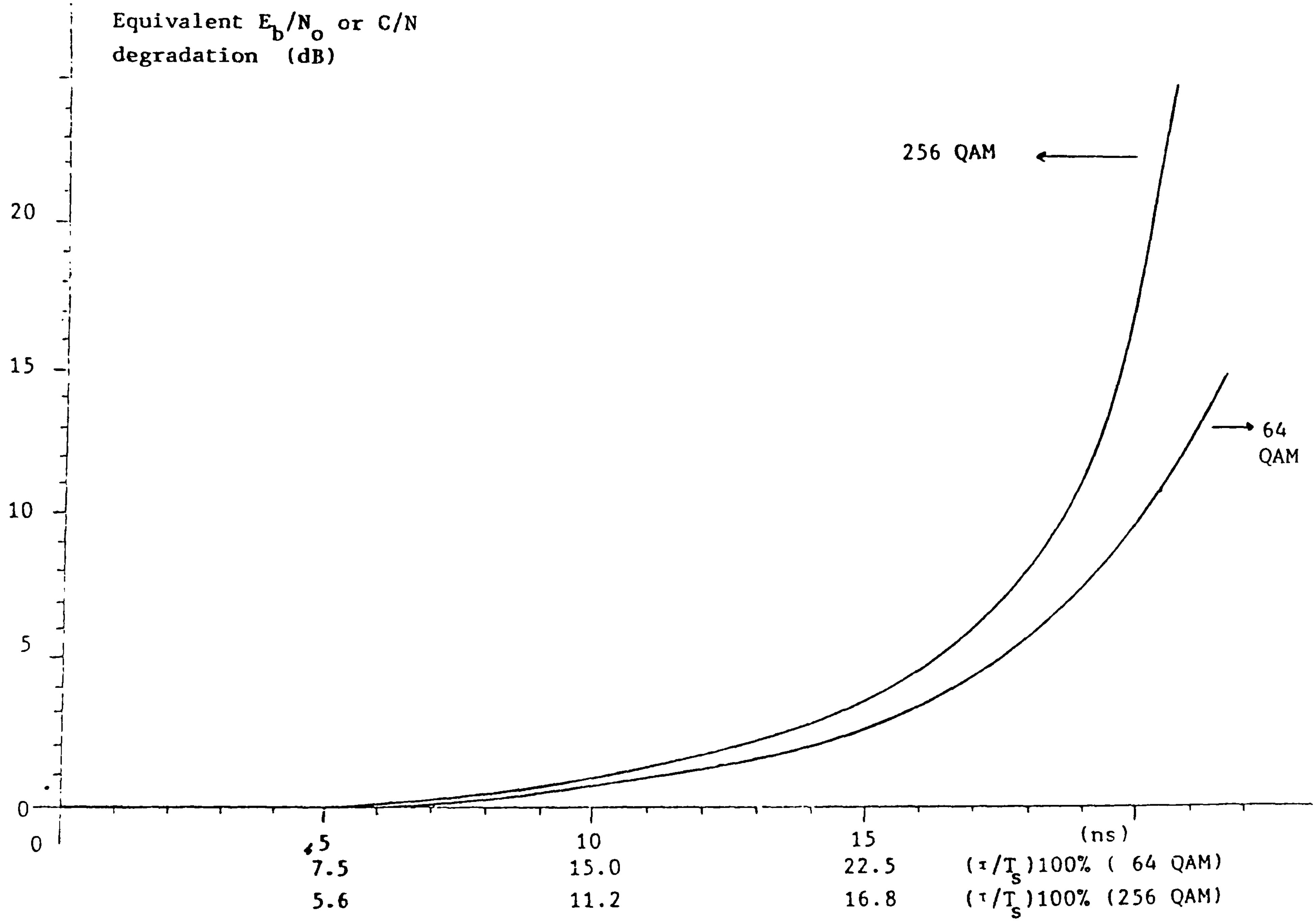


Fig. 3.20 Equivalent E_b/N_o or C/N degradation due to time delay difference τ between the RF received signals in BB 64 and 256 QAM SOSDCS

τ = 10.0000 ns

α = 1.0

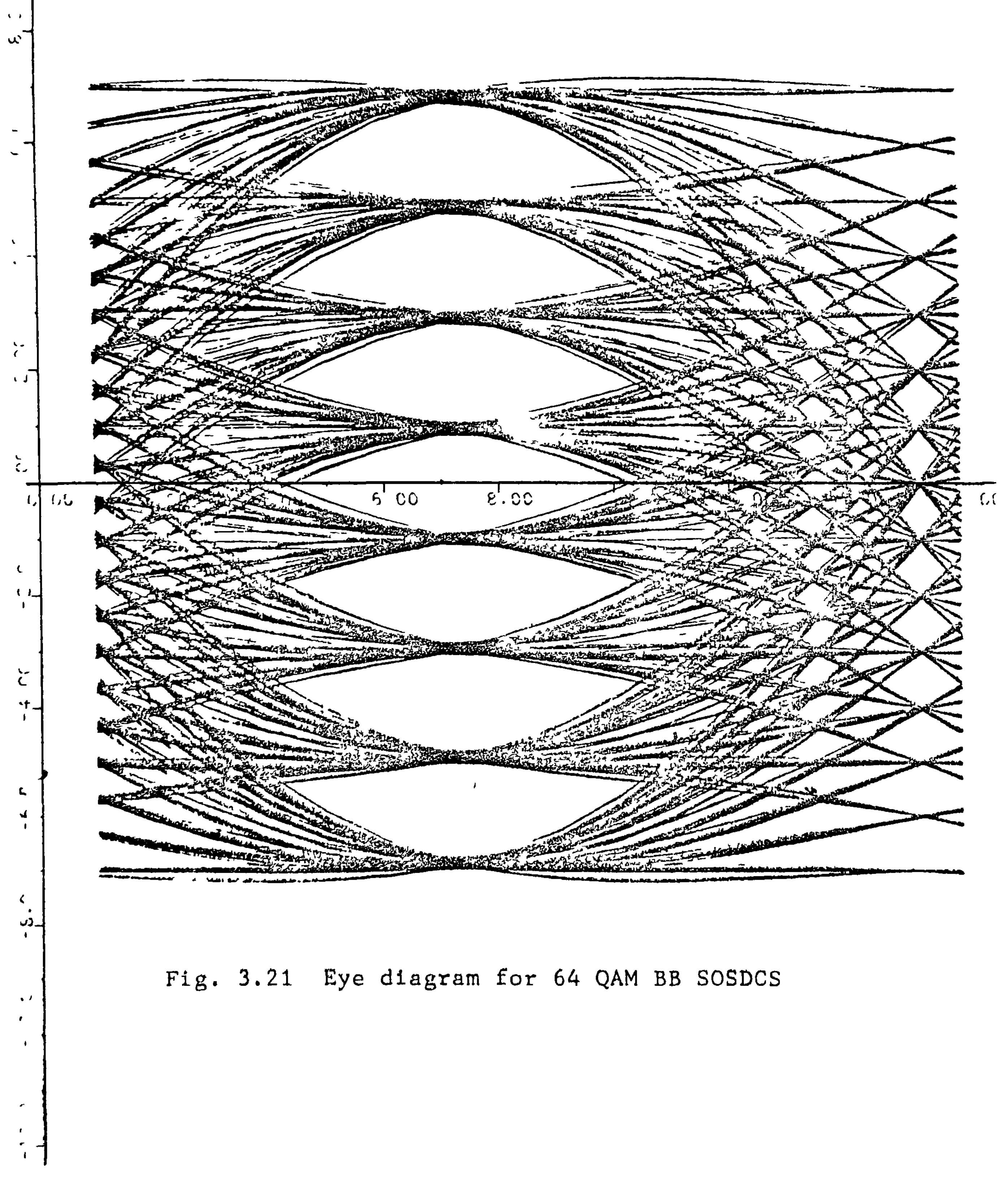


Fig. 3.21 Eye diagram for 64 QAM BB SOSDCS

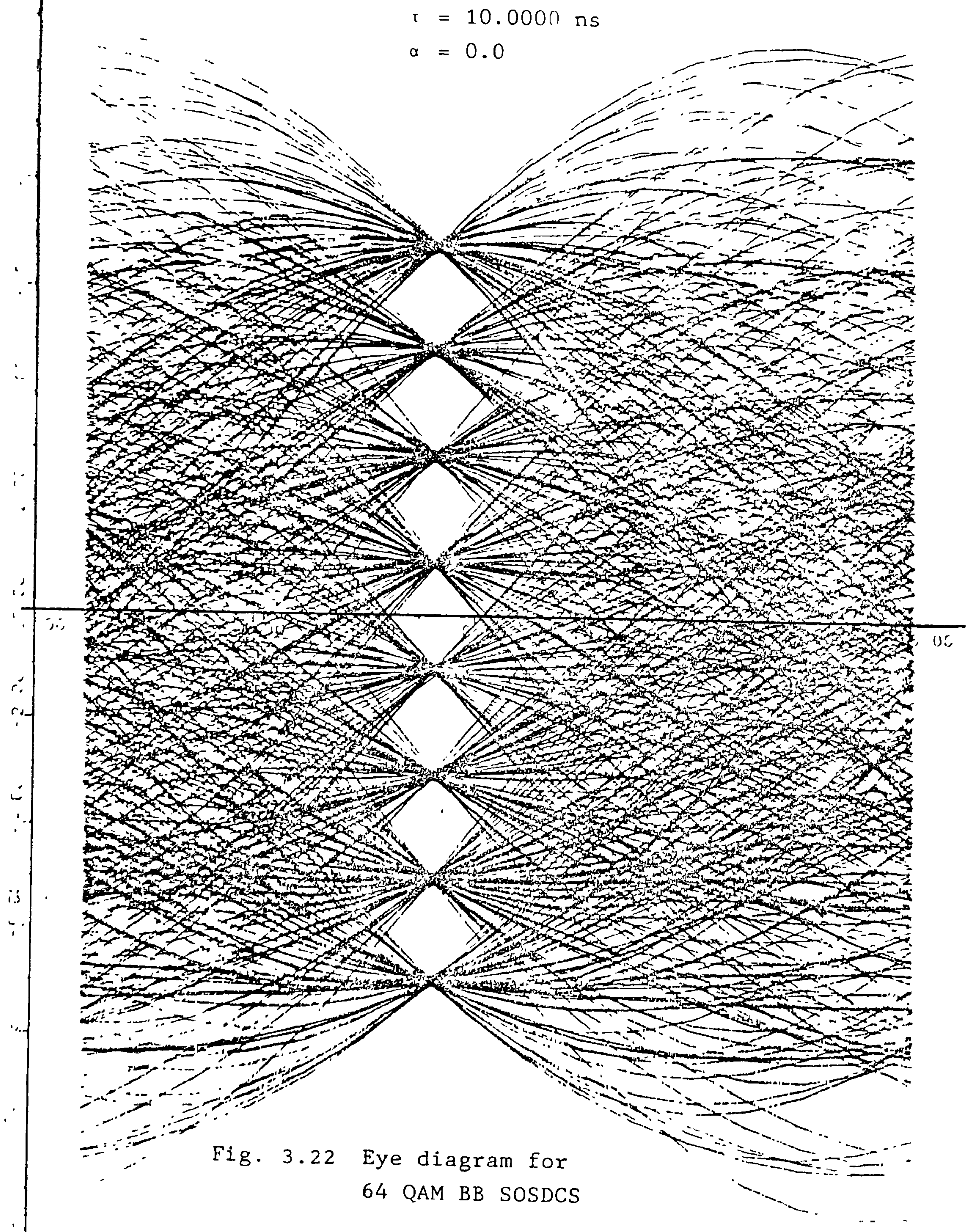


Fig. 3.22 Eye diagram for
64 QAM BB SOSDCS

canceled by employing adaptive time delay lines (ATDL). An ATDL continually compensates for the time delay difference between the received signals but it is very difficult to design [27].

On the other hand , it was observed in section 3.1.2 that by employing coherent demodulation in each of the combining channels, which can be made for BB but not for RF and IF combining systems, an important performance improvement can be obtained.

Employment of coherent demodulation in each of the combining channels implies a phase difference $2\pi f_c\tau$ between the demodulation carriers of frequency f_c, in second order combining systems with delay difference τ between the combining channels.

It is obvious that for RF and IF combining systems, where demodulation is made after combining the received signals, it is impossible to take advantage of individual coherent demodulation. Nevertheless, for the type of demodulation employed, the effect of a phase shift in the demodulation carrier is identical to the effect of introducing a compensating constant phase shift across the signal bandwith [30].

The same benefit obtained by using coherent demodulation in each channel of a BB SOSDCS can then be achieved in RF and IF SOSDCS by introducing a compensating constant phase shift $2\pi f_c\tau$ across the bandwidth of one of the combining signals.

In our computer program, phase compensation was simulated by means of the appropriate subroutine, as explained in Chapter IV. Results for the simulated RF and IF SOSDCS with phase compensation were exactly the same as previously found for BB SOSDCS.

The fact that analog phase shifters are not infinitely precise or that discrete phase shifters provide discrete phase compensation, is studied in Appendix B where we determine the corresponding system performance degradation.

In Appendix C an alternative way of obtaining phase compensation for IF combining systems is analyzed and found to provide the same system performance as the method presented in this section.

CHAPTER IV

COMPUTER SIMULATION

Computer simulation is a powerful alternative to theoretical analysis when, as in our case, it becomes extremely complicated. Therefore, computer simulation programs has been developed for this thesis, to be able to extend our theoretical discussion of Chapter III.

In general our programs are based on those given in the Digital Communication Group (DCG) Report # 112 [32]. Basic modifications similar to those made in reference [21], to adapt the programs for the study of 64 QAM system performance, were made to design the basic program for the study of 256 QAM system performance.

4.1 SIMULATION MODEL

Our computer simulation model for QAM SOSDCS is based on the equivalent baseband concept and is illustrated in Fig. 4.1. To simulate both of the combining signals, a non-return-to-zero (NRZ) data signal was created using a pseudo-random generated sequence of length 2^{11}. This data was then converted into the complex baseband format with the in-phase component corresponding to the real part of the complex data and the quadrature component corresponding to the imaginary part.

The transmit and receive filters are modelled in the equivalent low-pass complex format. They are phase-equalized root of raised

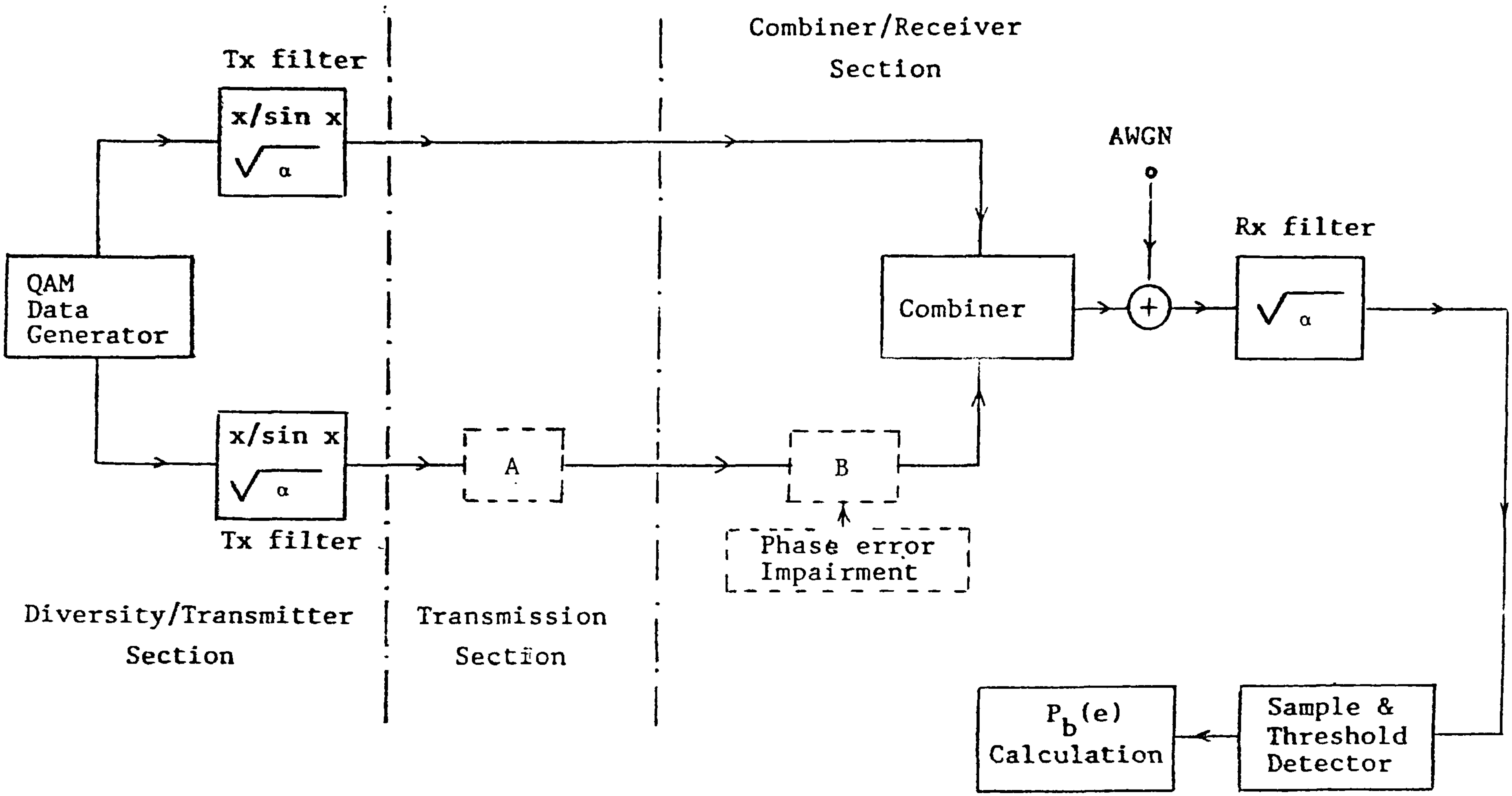


Fig. 4.1 Computer simulation model

A = Time delay difference
B = Wideband phase shifter

cosine filters, with and without x/sinx amplitude equalization and roll-off factor α.

The filtering process is effected by using fast Fourier transform (FFT) techniques. The complex baseband data is transformed into the frequency domain where it is multiplied by the complex low-pass transfer function of the filter under consideration and then transformed back to the time domain.

A time delay difference between the RF received signals and wide-band phase correction are simulated, in the frequency domain, by introducing the appropriate phase variation in one of the combining signals, in accordance with models developed in Chapter III. Perfect or imperfect phase correction can be selected by setting the parameter BETA to zero or the desired value of phase error.

In the case of BB SOSDCS the time delay difference between the BB received signals can be simulated also by rotating one of the data signals by the corresponding number of sample intervals.

Operation of the combiner is simulated by adding information signals in a voltage basis and noise signals in a power basis.

In the receiver, the detector is represented by a sampler and threshold detector. The sampler is simulated by a synchronization subroutine which samples the data in points as close as possible to maximum eye opening.

Finally the program calculates the bit error probability $P_b(e)$ for different values of E_b/N_o, assuming the noise present at the decision point is a Gaussian signal. The method of calculating $P_b(e)$ as a function of E_b/N_o is explained in section 4.2.

Adequate subroutines to plot $P_b(e)$ versus E_b/N_o performance curves and eye patterns are also included in the program.

4.2 PROBABILITY OF ERROR CALCULATION

In the calculation of probability of error performance of digital systems, one essential feature is the inclusion of the combined effect of inter-symbol-interference (ISI) and noise. ISI tends to be data dependent, for this reason and in a strict sense, the data sequence used in the simulation must contain all possible data transition and the number of transitions must be balanced so that all data transitions are given equal weight in the determination of P(e) performance [21].

On the other hand, a large number of samples per symbol must be used in high-level QAM systems to assure adequate precision in the selection of the sampling instants at detection.

Necessity of a large data sequence and number of samples per symbol, plus the fact that as the total number of samples handled by the program is increased its operability is reduced, made us to initially test the programs with different values of sequence length and number of samples per symbol, in order to determine practical values to be used in our simulation.

Reasonable results and operability were found using a sequence length of 2^{11} and 64 samples per symbol, i.e., handling a total number of 2^{18} samples per signal in the programs.

Once the set of points corresponding to the sampling instants has been selected by the synchronization subroutine, the detector separates the complex received data into its two component parts (in-phase and quadrature) and calculates, for each symbol, the probability that the corresponding space diagram point is out of the correct decision region for a specified value of Gaussian noise power at the decision point.

Once the symbol error probability for the selected range of noise power is obtained for each symbol, the detection sub-routine reduces them to bit error probability, considering a Gray code scheme and computes the average value over the entire symbol sequence, relating them to the corresponding E_b/N_o value in an ideal uncombined system.

Specifically, for a given value of Gaussian noise power N, the average bit error probability $P_b(e)$ for the M-ary QAM system (M = 64 or 256 in our case), is given by:

$$P_b(e) = (1/n \log_2 \sqrt{M}) \sum_{i=1}^{n} P_i(e) \qquad (4.1)$$

Where n is the total number of symbols including both, in-phase and quadrature components and $P_i(e)$ is the error probability of the i^{th} symbol. If the i^{th} transmitted symbol is an end point (those in-phase or quadrature symbols for which the transmitted value was a positive or negative extreme), $P_i(e)$ is given by:

$$P_i(e) = (1/2) \operatorname{erfc}[(\bar{S}_i - TH1_i)/\sqrt{2N}] \qquad (4.2)$$

And, if not an end point, by:

$$P_i(e) = (1/2)\ \mathrm{erfc}[(\bar{S}_i - TH1_i)/\sqrt{2N}] + (1/2)\ \mathrm{erfc}[(TH2_i - \bar{S}_i)/\sqrt{2N}] \qquad (4.3)$$

Where $\bar{S}_i$ is the magnitude of the i^{th} symbol sample, erfc(x) is the complementary error function as in equations 2.1 and 2.2 $TH1_i$ is the magnitude of the lower threshold level corresponding to the i^{th} transmitted symbol, $TH2_i$ is the magnitude of the upper threshold level corresponding to the i^{th} transmitted symbol and N is noise power at the decision point.

CHAPTER V

RECOMMENDED FURTHER RESEARCH

Our study was restricted to systems in which only time delay difference between the RF received signals occurs. In practical systems, amplitude variation can affect the combining signals making necessary the usage of amplitude gain control (AGC) devices.

A logical second step in this research is to study the effect of amplitude variation of the combining signals when AGC circuits are employed in the system.

A possible further step is the consideration of combining signals affected by multipath fading and/or interference in addition to time delay difference and/or amplitude variation.

The programs used to simulate basic 64 and 256 QAM systems are extensions of those developed in reference [12] for lower level systems. Problems observed in these programs, specifically the very long computing time caused by the synchronization subroutine and inoperability resulting when the total number of samples handled by the programs is increased, make desirable to investigate and design particular programs for high-level systems.

CHAPTER VI

CONCLUSION

RF and IF second-order space diversity combining systems (SOSDCS) with high-order QAM modulation schemes are very sensitive to time delay difference between the RF received signals. A few nanosecons of time delay difference render inoperable these systems.

Corresponding BB SOSDCS are much less sensitive to time delay difference but twice as many demodulators and combiners are required in comparison with RF and IF SOSDCS.

High precision phase compensation for the band-center-frequency phase difference, across the signal bandwidth, makes RF and IF combining systems to exhibit the same P_b (e) performance as BB combining systems, independently of the carrier frequency employed. The maximum permisible bit rate in the system is a function of the time delay difference and the required bit error probability.

High precision phase shifters required to perform the phase correction operation and the fact that phase correction in RF combining systems must be done for all the frequency components in the RF signal bandwidth while in IF combining systems RF single frequency phase correction can be implemented, make IF combining systems more attractive when practical realization is considered.

Performance and sensitivity curves determined in Chapter III and Appendix B, respectively, facilitate the design of second-order combining systems with 64 and 256 QAM modulation schemes, when the time delay difference is statistically known.

APPENDIX A

PROGRAMS TO SIMULATE RF, IF AND BB 64 AND 256 QAM SOSDCS

In this Appendix we present listing of the computer programs which were used to evaluate the performance of RF, IF and BB 64 and 256 QAM SOSDCS. The subroutines which are essentially common to the programs are listed only in the first program.

```
C***********************************************************************
C       PROGRAM TO SIMULATE THE P(E) PERFORMANCE OF 64 QAM
C       SECOND-ORDER DIVERSITY COMBINING SYSTEMS
C       CF    = RF CARRIER FREQUENCY
C       TD    = TIME DELAY BETWEEN RF RECEIVED SIGNALS
C       BETA  = PHASE ERROR IN PHASE CORRECTION
        AA    = EQUIVALENT ATTENUATION OF THE COMBINED SIGNAL
C       THA   = THRESHOLD ADJUSTMENT PARAMETER
C***********************************************************************
        COMMON /NUMB1/ LDIM,LSAMPL
        COMMON /NUMB2/ BAUD,ALPHA,FBW
        COMMON /NUMB3/ NSYMB,NSNR,NERROR
        COMMON /NUMB4/ PER
        COMMON /NUMB5/ BETA
        COMPLEX DATA1(131072),TFTX(131072),DATA2(131072),TFRX(131072)
        COMPLEX DATA3(131072),TFPC(131072),TFTD(131072),DATA4(131072)
        DIMENSION NI(2048),NQ(2048),EBNO(36),PEI(36)
        DIMENSION XARRAY(38),YARRAY(38)
C
C       INITIALIZE PROGRAM.
C
        BETA=0.0
        NERROR=0
        LSAMPL=64
        LDIM=131072
        BAUD=15.
        ALPHA=0.4
        FBW=7.5
        NSYMB=2048
        NSNR=36
        NRUNS=9
        CF=4000.0
        PER=10.0
        T1=(0.00125)/(10.0**3.0)
        T2=((0.25)/(10.0**3.0))*(PER)
        WRITE(5,10)
   10   FORMAT(5X,'SIMULATION OF 64-QAM COMBINING'/)
        WRITE(5,11) LSAMPL,LDIM,BAUD,ALPHA
   11   FORMAT(5X,'SAMPLES PER SYMBOL=',I4,/5X,'LDIM=',I7,/
      &5X,'SYMBOL RATE=',F7.2,/5X,'ALPHA=',F7.2,/)
C
C       END OF INITIALIZATION.
C
C       START OF COMPUTATIONS.
C
        CALL RCOSTX(TFTX)
        CALL RCOSRX(TFRX)
        CALL HHGG(TFRX,PNOISE)
C
        PNOISE=PNOISE*0.5
        WRITE(5,156)PNOISE
  156   FORMAT(10X,'PNOISE NOW =',F7.3,/)
C
C       CHANNEL 1
C
        CALL LO64(DATA1,NI,NQ)
        CALL FILTER(DATA1,TFTX)
        CALL ENERGY(DATA1,EB)
        WRITE(5,155)EB
  155   FORMAT(10X,'EB=',E14.8)
C
        DO 158 KM=1,NRUNS
C
C
C       CHANNEL 2
C
        DO 21 I=1,LDIM
        DATA2(I)=DATA1(I)
   21   CONTINUE
        V=FLOAT(KM)
        TD=(T2)+((T1)*(V-1.0))
        AA=1.0-(0.02)*V
        THA=1.00000000
        IT=((KM)*3)+(3)
        CALL TDEL(TD,CF,TFTD)
        CALL FILTER(DATA2,TFTD)
        CALL PHCO(TD,CF,TFPC)
        CALL FILTER(DATA2,TFPC)
C       CALL TIME(DATA2,IT)
```

```
C
C      ADD THE TWO DATA SIGNALS
C
       DO 401 I=1,LLIM
  401  DATA3(I)=(DATA1(I)+DATA2(I))/2.000000
C      CALL ATA(DATA3,AA)
C
C      DECODING  (THE COMBINED CHANNEL)
C
       CALL FILTER(DATA3,TFRX)
       CALL SYNCRO(DATA3,PNOISE,NI,NQ,MI,MQ,EP,THA)
C      CALL EYEQ(DATA3)
       CALL DECO(DATA3,PNOISE,MI,MQ,NI,NQ,EBNO,PEI,FB,THA)
C
C      CALL THE DRAWING ROUTINE
C
       DO 157 I=1,NSNR
       XARRAY(I)=EBNO(I)
  157  YARRAY(I)=PEI(I)
       CALL DRAW(XARRAY,YARRAY,KM,NRUNS)
  158  CONTINUE
       STOP
       END
C
C*********************************************************************
C     RAISED COSINE FILTER WITH ARBITRARY ALPHA AND X/SINX
C*********************************************************************
       SUBROUTINE RCOSTX(TF1)
       COMPLEX TF1(131072)
       COMMON /NUMB1/ LDIM,LSAMPL
       COMMON /NUMB2/ BAUD,ALPHA,FBW
       SBANDW=BAUD*FLOAT(LSAMPL)
       NO=LDIM/2
       NO1=NO+1
       IF (ALPHA.EQ.0.0) ALPHA=0.0001
       FN=LDIM*(FB./SBANDW)
       F1=(1.-ALPHA)*FN
       F2=(1.+ALPHA)*FN
       IFN=IFIX(FN)
       IF1=IFIX(F1)+1
       IF2=IFIX(F2)+1
C
C     THE AMPLITUDE CHARACTERISTICS
C
       A1=3.141592/(2.*FLOAT(IFN))
       DO 8 I=2,IF1
       TF1(1)=CMPLX(1.0,0.0)
       J=I-1
C
C      X/SIN(X) EQUALIZATION
C
       A2=(FLOAT(J)*A1)/(SIN(FLOAT(J)*A1))
       TF1(I)=CMPLX(A2,0.0)
    8  CONTINUE
       JK=IF1+1
       DO 9 J=JK,IF2
       I=J-1
C
C      ROOT OF RAISED COSINE
C
       A3=(FLOAT(I)*A1)/(SIN(FLOAT(I)*A1))
       A=(3.141516/(2.0*ALPHA))*((FLOAT(I)/FLOAT(IFN))-1.)
       TF1(J)=CMPLX(SQRT(0.5*(1.0-SIN(A))),0.0)
       TF1(J)=TF1(J)*CMPLX(A3,0.0)
    9  CONTINUE
       JH=IF2+1
       DO 10 I=JH,NO1
       TF1(I)=CMPLX(0.0,0.0)
   10  CONTINUE
       NO2=NO1+1
       DO 5 I=NO2,LDIM
       TF1(I)=CONJG(TF1(LDIM+2-I))
    5  CONTINUE
       RETURN
       END
```

```
C
C*****************************************************************
C     RAISED COSINE FILTER WITH ARBITRARY ALPHA
C*****************************************************************
      SUBROUTINE RCOSRX(TF2)
      COMPLEX TF2(131072)
      COMMON /NUMB1/ LDIM,LSAMPL
      COMMON /NUMB2/ BAUD,ALPHA,FBW
      SBANDW=BAUD*FLOAT(LSAMPL)
      NO=LDIM/2
      NO1=NO+1
      IF (ALPHA.EQ.0.0) ALPHA=0.0001
      FN=LDIM*(FBW/SBANDW)
      F1=(1.-ALPHA)*FN
      F2=(1.+ALPHA)*FN
      IFN=IFIX(FN)
      IF1=IFIX(F1)+1
      IF2=IFIX(F2)+1
C
C     AMPLITUDE CHARACTERISTIC
C
      A1=3.141592/(2.*FLOAT(IFN))
      DO 8 I=2,IF1
      TF2(1)=CMPLX(1.0,0.0)
      J=I-1
      TF2(I)=CMPLX(1.,0.0)
    8 CONTINUE
      JK=IF1+1
      DO 9 J=JK,IF2
      I=J-1
C
C     ROOT OF RAISED COSINE
C
      A=(3.141592/(2.0*ALPHA))*((FLOAT(I)/FLOAT(IFN))-1.)
      TF2(J)=CMPLX(SQRT(0.5*(1.0-SIN(A))),0.0)
    9 CONTINUE
      JH=IF2+1
      DO 10 I=JH,NO1
      TF2(I)=CMPLX(0.0,0.0)
   10 CONTINUE
      NO2=NO1+1
      DO 5 I=NO2,LDIM
      TF2(I)=CONJG(TF2(LDIM+2-I))
    5 CONTINUE
      RETURN
      END
C*****************************************************************
C     THE FOLLOWING SUBROUTINE PERFORMS THE FILTERING
C     PROCESS ON THE DATA SEQUENCE.
C*****************************************************************
C
      SUBROUTINE FILTER(DATA,TF)
      COMPLEX DATA(131072),TF(131072)
      DIMENSION IWK(18)
      COMMON /NUMB1/ LDIM,LSAMPL
      CALL FFT2C(DATA,17,IWK)
      DO 1 I=1,LDIM
    1 DATA(I)=CONJG(DATA(I)*TF(I))
      CALL FFT2C(DATA,17,IWK)
      DO 2 I=1,LDIM
    2 DATA(I)=CONJG(DATA(I))/FLOAT(LDIM)
      RETURN
      END
C
```

```
C*************************************************************
C     THIS SUBROUTINE COMPUTES THE EFFECTIVE NOISE (PNOISE)
C     AT THE OUTPUT OF THE RECEIVE FILTER
C*************************************************************
C
      SUBROUTINE HHGG(TF,PNOISE)
      COMPLEX TF(131072)
      COMMON /NUMB1/ LDIM,LSAMPL
      COMMON /NUMB2/BAUD,ALPHA,FBW
      SUM=0.0
      SBANDW=BAUD*LSAMPL
      DO 1 L=1,LDIM
      HH=(CABS(TF(L)))**2
    1 SUM=SUM+HH
      PNOISE=SUM*SBANDW/FLOAT(LDIM)/2.
      WRITE (5,2) PNOISE
    2 FORMAT(5X,'PNOISE=',F7.3,/)
      RETURN
      END
C
C
C*************************************************************
C     BIT ENERGY COMPUTATION
C *************************************************************
      SUBROUTINE ENERGY(DATA,EB)
      COMPLEX DATA(1)
      WATTS=0.
      COMMON /NUMB1/ LDIM,LSAMPL
      COMMON /NUMB2/ BAUD,ALPHA,FBW
      DO 1 I=1,LDIM
      WATTS=WATTS+((CABS(DATA(I)))**2.)
    1 CONTINUE
      WATTS=WATTS/(FLOAT(LDIM))
      EB=WATTS/(6.*BAUD)
      RETURN
      END
C*************************************************************
C     THIS SUBROUTINE GENERATE A BASEBAND 64-QAM SIGNAL
C*************************************************************
C
      SUBROUTINE LO64(DATA,NI,NQ)
      COMPLEX DATA(1)
      DIMENSION NI1(2048),NQ1(2048),NX(38),NY(38),NI(1),NQ(1)
      COMMON /NUMB1/ LDIM,LSAMPL
      DATA IT1,IT2,IT3,IT4/2,11,0,0/
      DATA MREG,MODTYP/11,3/
C
      NO=MREG
      NM=NO-1
      N1=NO+1
      N2=NO+2
      N3=NO+3
      N4=NO+4
C
      NMX=2**NO
      NMX1=NMX-1
      NO2=IFIX(FLOAT(NO)/2.)
C
C     INITIALIZE THE SHIFT REGISTER
C
      DO 34 I=1,NO
      IF(I.LT.NO2) NX(I)=1
      IF(I.GE.NO2) NX(I)=-1
      NY(I)=-NX(I)
   34 CONTINUE
C
C     PN SEQUENCE GENERATION
C
      NX(N3)=1
      NY(N3)=1
      NX(N4)=1
      NY(N4)=1
      DO 36 K=1,NMX1
      NX(N1)=NX(IT1)
      NX(N2)=NX(IT2)
      IF(IT3.GT.0) NX(N3)=NX(IT3)
      IF(IT4.GT.0) NX(N4)=NX(IT4)
```

```
      NY(N1)=NY(IT1)
      NY(N2)=NY(IT2)
      IF(IT3.GT.0) NY(N3)=NY(IT3)
      IF(IT4.GT.0) NY(N4)=NY(IT4)
      DO 37 I1=1,NM
      NX(N1-I1)=NX(NO-I1)
      NY(N1-I1)=NY(NO-I1)
   37 CONTINUE
      NX(1)=NX(N1)*NX(N2)*NX(N3)*NX(N4)
      NY(1)=NY(N1)*NY(N2)*NY(N3)*NY(N4)
      NI1(K)=NX(1)
      NQ1(K)=NY(1)
   36 CONTINUE
      NI1(NMX)=-1
      NQ1(NMX)=-1
C
C     GENERATE THE 64 QAM DATA
C
      K1=1
      K2=3
      K3=5
      M1=2**(MODTYP-1)
      M2=MODTYP-1
      M3=0
      IF(MODTYP.EQ.3) M3=1
      DO 100 I2=1,NMX
      IF(K3.GT.NMX) K3=1
      IF(K2.GT.NMX) K2=1
      NI(I2)=NI1(K1)*M1+NI1(K2)*M2+NI1(K3)*M3
      NQ(I2)=NQ1(K1)*M1+NQ1(K2)*M2+NQ1(K3)*M3
C
C     LOAD INTO SAMPLE ARRAY, LSAMPL SAMPLES PER SYMBOL
C
      J1=(I2-1)*LSAMPL+1
      J2=I2*LSAMPL
      DO 101 J=J1,J2
      DATA(J)=CMPLX(FLOAT(NI(I2)),FLOAT(NQ(I2)))
  101 CONTINUE
      K1=K1+1
      K2=K2+1
      K3=K3+1
  100 CONTINUE
      RETURN
      END
C
C*****************************************************************
C     THE FOLLOWING SUBROUTINE SYNCHRONIZES THE RECEIVED DATA.
C*****************************************************************
C
      SUBROUTINE SYNCRO(DATA,PNOISE,NI,NQ,MI,MQ,EB,THA)
      INTEGER Q7FLAG
      COMPLEX DATA(1),AMP
      DIMENSION NI(1),NQ(1)
      COMMON /NUMB1/LDIM,LSAMPL
      COMMON /NUMB2/ BAUD,ALPHA,FBW
      COMMON /NUMB3/ NSYMB,NSNR,NERROR
      NERROR=0
C
C     SYNCHRONIZE THE RECEIVED DATA
C
  111 NOLD=0
      NOF=0
      K=1
  300 CONTINUE
      NEW=0
      DO 200 J=1,NSYMB
      J1=K+(J-1)*LSAMPL
      IF(J1.GT.LDIM) J1=J1-LDIM
      AXBAR=REAL(DATA(J1))
      AYBAR=AIMAG(DATA(J1))
      SS=AXBAR*NI(J)
      IF (SS .GT. 0.) NEW=NEW+1
      SS=AYBAR*NQ(J)
      IF (SS .GT. 0.) NEW=NEW+1
  200 CONTINUE
```

```
      IF(NOLD.GE.NEW) GO TO 399
      NOLD=NEW
      NOF=K
  399 K=K+1
      LFU=2*NSYMB
C
C     SHIFT RECEIVED DATA RIGHT UNTIL ALL SYMBOLS ARE LINED UP
C

      IF (NOLD.LT.LFU.AND.K.LE.LDIM) GO TO 300
      NOF=NOF-1
      WRITE (5,260) NOF
  260 FORMAT(10X,'NOF=',I4,/)
      IF (NOF.EQ.0) GO TO 230
      LD=LDIM-1
      DO 250 I=1,NOF
      AMP=DATA(1)
      DO 240 J=1,LD
      DATA(J)=DATA(J+1)
  240 CONTINUE
      DATA(LDIM)=AMP
  250 CONTINUE
C
C      OPTIMIZE THE SAMPLING INSTANT
C
  230 MI=1
      MQ=1
      EOI=FLOAT(NSYMB) + 1.
      EOQ=FLOAT(NSYMB) + 1.
      VARIAN=PNOISE*EB/73.43
      WRITE(5,55)VARIAN
   55 FORMAT(10X'VARIAN=',F7.3,/)
      SIGMA=SQRT(VARIAN)
      WRITE(5,56)SIGMA
   56 FORMAT(10X,'SIGMA=',F7.3,/)
      DO 90 J=1,LSAMPL
      EI=0.
      EQ=0.
      DO 70 K=1,NSYMB
C     VERIFY THAT NO SAMPLED SYMBOL IS IN ERROR
      J1=(K-1)*LSAMPL+J
      IF(J1.EQ.LDIM) GO TO 20
      AXBAR=(REAL(DATA(J1))+REAL(DATA(J1+1)))/2.
      AYBAR=(AIMAG(DATA(J1))+AIMAG(DATA(J1+1)))/2.
      IF(J1.LT.LDIM) GO TO 21
   20 AXBAR=(REAL(DATA(LDIM))+REAL(DATA(1)))/2.
      AYBAR=(AIMAG(DATA(LDIM))+AIMAG(DATA(1)))/2.
   21 INDEXI=0
      SS=FLOAT(NI(K))*AXBAR
      IF (SS .LT. 0.) INDEXI=1
      AMPX=ABS(AXBAR)
      AMPI=ABS(FLOAT(NI(K)))
      THR1I=(AMPI - 1.0)*(THA)
      THR2I=(AMPI + 1.0)*(THA)
      I7FLAG=0
      IF (AMPI.EQ. 7.0) GO TO 30
      IF (( AMPX.GE.THR2I).OR.(AMPX.LE.THR1I)) INDEXI=1
      GO TO 40
   30 IF(AMPX.LE.THR1I) INDEXI=1
      I7FLAG=1
   40 CONTINUE
      INDEXQ=0
      SS=FLOAT(NQ(K))*AYBAR
      IF(SS .LT. 0.) INDEXQ=1

      AMPY=ABS(AYBAR)
      AMPQ=ABS(FLOAT(NQ(K)))
      THR1Q=(AMPQ - 1.0)*(THA)
      THR2Q=(AMPQ + 1.0)*(THA)
      Q7FLAG=0
      IF (AMPQ.EQ. 7.0) GO TO 50
      IF (( AMPY.GE.THR2Q).OR.(AMPY.LE.THR1Q)) INDEXQ=1
      GO TO 60
   50 IF(AMPY.LE.THR1Q) INDEXQ=1
      Q7FLAG=1
   60 CONTINUE
      IF(INDEXI.EQ.1) EI=EI + 1.
      IF(INDEXQ.EQ.1) EQ= EQ + 1.
```

```
C
C
C      COMPUTE THE PROBABILITY OF ERROR FOR THIS SYMBOL AT EB/NO=19DB
C
       D1=ABS(AMPX-THR1I)
       ARG=D1/(SIGMA*SQRT(2.))
C      CHECK IF ARG IS LARGE INWHICH CASE PE IS INSIGNIFICANT
C
       IF (ARG.GT.12.) ARG=12.
       EI=EI + ERFC(ARG)/2.
       IF (I7FLAG .EQ. 1) GO TO 65
       D2=ABS(THR2I-AMPX)
       ARG=D2/(SIGMA*SQRT(2.))
       IF (ARG.GT.12.) ARG=12.
       EI=EI + ERFC(ARG)/2.
    65 D1=ABS(AMPY-THR1Q)
       ARG=D1/(SIGMA*SQRT(2.))
       IF (ARG.GT.12.) ARG=12.
       EQ=EQ + ERFC(ARG)/2.
       IF (Q7FLAG .EQ. 1) GO TO 70
       D2=ABS(THR2Q-AMPY)
       ARG=D2/(SIGMA*SQRT(2.))
       IF (ARG.GT.12.) ARG=12.
       EQ=EQ + ERFC(ARG)/2.
    70 CONTINUE
       IF(EOI.LE.EI) GO TO 75
       EOI=EI
       MI=J
       WRITE(5,67)MI
    67 FORMAT(10X,'MI=',I4,/)
    75 CONTINUE
       IF (EOQ.LE.EQ) GO TO 80
       EOQ=EQ
       MQ=J
       WRITE(5,68)MQ
    68 FORMAT(10X,'MQ=',I4,/)
    80 CONTINUE
       MOFF=IABS(MI-MQ)
       WRITE(5,63)MOFF
    63 FORMAT(10X,I2,/)
       IF(MOFF.NE.0) WRITE(5,90) MOFF
    90 FORMAT(5X,'SAMPLING POINTS FOR I CHANNEL DIFFER BY',
      *I2,' SIXTEENTHS OF THE SYMBOL INTERVAL',/)
       PM=FLOAT(MI)
       WRITE(5,261)PM
   261 FORMAT(10X,'PM=',F7.3,/)
       PF=FLOAT(NOF)
       WRITE(5,262)PF
   262 FORMAT(10X,'PF=',F7.3,/)
       PL=FLOAT(LSAMPL)
       WRITE(5,263)PL
   263 FORMAT(10X,'PL=',F7.3,/)
       OFF=(FLOAT(MI) + FLOAT(NOF))/FLOAT(LSAMPL)
       WRITE(5,64)OFF
    64 FORMAT(10X,F7.3,/)
       WRITE(5,95) OFF
    95 FORMAT(5X,'RECEIVED DATA IS DELAYED BY ',F7.3,' SYMBOLS',/)
       RETURN
       END
C
C*****************************************************************
C      THE FOLLOWING SUBROUTINE DECODES THE RECEIVED DATA
C*****************************************************************
       SUBROUTINE DECO (DATA,PNOISE,MI,MQ,NI,NQ,EBNO,PE,FB,THA)
       COMPLEX DATA(1)
       DIMENSION EBNO(1),PE(1),NI(1),NQ(1)
       COMMON /NUMB1/ LDIM,LSAMPL
       COMMON /NUMB2/ BAUD,ALPHA,FBW
       COMMON /NUMB3/ NSYMB,NSNR,NERROR
       NERROR=0
       DO 1 I=1,NSNR
     1 PE(I)=0.0
C      CHECK AND COUNT ERRORS IN THE SAMPLING INSTANTS,
C      PE COMPUTATION
       DO 2 K=1,NSYMB
       J1=(K-1)*LSAMPL+MI
       J2=(K-1)*LSAMPL+MQ
       AXBAR=(REAL(DATA(J1))+REAL(DATA(J1+1)))/2.
       AYBAR=(AIMAG(DATA(J2))+AIMAG(DATA(J2+1)))/2.
```

```
      INDEXI=0
      OVI=FLOAT(NI(K))
      THR1I=(OVI-1.0)*(THA)
      THR2I=(OVI+1.0)*(THA)
      IF(OVI.EQ.7.0) GO TO 3
      IF(OVI.EQ.-7.0) GO TO 4
      IF((AXBAR.GE.THR2I).OR.(AXBAR.LE.THR1I)) INDEXI=1
      GO TO 5
    3 IF(AXBAR.LE.THR1I) INDEXI=1
      GO TO 5
    4 IF(AXBAR.GE.THR2I) INDEXI=1
    5 CONTINUE
C
      INDEXQ=0
      OVQ=FLOAT(NQ(K))
      THR1Q=(OVQ-1.0)*(THA)
      THR2Q=(OVQ+1.0)*(THA)
      IF(OVQ.EQ.7.0) GO TO 6
      IF(OVQ.EQ.-7.0) GO TO 7
      IF((AYBAR.GE.THR2Q).OR.(AYBAR.LE.THR1Q)) INDEXQ=1
      GO TO 8
    6 IF(AYBAR.LE.THR1Q) INDEXQ=1
      GO TO 8
    7 IF(AYBAR.GE.THR2Q) INDEXQ=1
    8 CONTINUE
      IF((INDEXI.EQ.1).OR.(INDEXQ.EQ.1)) NERROR=NERROR+1
C
C     COMPUTE THE PROBABILITY OF ERROR FOR THIS SYMBOL
C
      DO 9 M=1,NSNR
      PEI1=0.0
      PEI2=0.0
      PEQ1=0.0
      PEQ2=0.0
      VARIAN=PNOISE*ES*(10.**(-0.1*FLOAT(M)))
      SIGMA=SQRT(VARIAN)
      IF(OVI.EQ.-7.0) GO TO 10
      DI=AXBAR-THR1I
      D1=ABS(DI)
      ARG=D1/(SIGMA*SQRT(2.))
      IF(ARG.GT.12.0) ARG=12.0
      PEI1=ERFC(ARG)/2.0
      IF(DI.LE.0.0) PEI1=1.-PEI1
      IF(OVI.EQ.7.0) GO TO 11
   10 DII=THR2I-AXBAR
      D2=ABS(DII)
      ARG=D2/(SIGMA*SQRT(2.))
      IF(ARG.GT.12.0) ARG=12.0
      PEI2=ERFC(ARG)/2.0
      IF(DII.LE.0.0) PEI2=1.-PEI2
   11 PEI=(PEI1+PEI2)/3.0
C
      IF(OVQ.EQ.-7.0) GO TO 12
      DI=AYBAR-THR1Q
      D1=ABS(DI)
      ARG=D1/(SIGMA*SQRT(2.))
      IF(ARG.GT.12.0) ARG=12.0
      PEQ1=ERFC(ARG)/2.0
      IF(DI.LE.0.0) PEQ1=1.0-PEQ1
      IF(OVQ.EQ.7.0) GO TO 13
   12 DII=THR2Q-AYBAR
      D2=ABS(DII)
      ARG=D2/(SIGMA*SQRT(2.))
      IF(ARG.GT.12.0) ARG=12.0
      PEQ2=ERFC(ARG)/2.0
      IF(DII.LE.0.0) PEQ2=1.0-PEQ2
   13 PEQ=(PEQ1+PEQ2)/3.0
C
      IF(PEI.LT.1.E-15) PEI=0.0
      IF(PEQ.LT.1.E-15) PEQ=0.0
    9 PE(M)=PE(M)+(PEI+PEQ)/2.0
    2 CONTINUE
C
      DO 14 I=1,NSNR
```

```
      PE(I)=PE(I)/FLOAT(NSYMB)
   14 EBNO(I)=FLOAT(I)
      WRITE(5,15)
   15 FORMAT(5X,'EB/NO',10X,'PROB. OF ERROR',/)
      WRITE(5,16)(EBNO(I),PE(I),I=1,NSNR)
   16 FORMAT(1X,F5.1,10X,E13.6)
      WRITE(5,17)NERROR
   17 FORMAT(/5X,'ERRORS=',I5,/)
      RETURN
      END
C
C*****************************************************************
C     THE FOLLOWING SUBROUTINE INTRODUCES A TIME DELAY
C     DIFFERENCE TO BETWEEN THE TWO RF RECEIVED SIGNALS
C*****************************************************************
C
      SUBROUTINE TDEL(TD,CF,TF3)
      COMMON /NUMB1/ LDIM,LSAMPL
      COMMON /NUMB2/ BAUD,ALPHA,FBW
      COMPLEX TF3(131072)
      SBANDW=FLOAT(LSAMPL)*BAUD
      SLP=-2.0*3.141592*TD
      N01=LDIM/2
      N02=N01+1
      DO 1 I=1,N01
      PHI1=(SLP)*((CF)+((FLOAT(I-1)*SBANDW)/(FLOAT(LDIM)+1.0)))
      TF3(I)=CMPLX(COS(PHI1),SIN(PHI1))
    1 CONTINUE
      DO 2 I=N02,LDIM
      PHI2=(SLP)*((CF)-((FLOAT(LDIM-I+1)*SBANDW)/(FLOAT(LDIM)+1.0)))
      TF3(I)=CMPLX(COS(PHI2),SIN(PHI2))
    2 CONTINUE
      RETURN
      END
C
C*****************************************************************
C     THE FOLLOWING SUBROUTINE COMPENSATES THE PHASE
C     DIFFERENCE BETWEEN THE COMBINING SIGNALS, IN A
C     VALUE EQUAL TO THE BAND-CENTER-FREQUENCY PHASE
C     DIFFERENCE, ACROSS THE SIGNAL BANDWITH.
C*****************************************************************
C
      SUBROUTINE PHCO(TD,CF,TF4)
      COMMON /NUMB1/ LDIM,LSAMPL
      COMMON /NUMB2/ BAUD,ALPHA,FBW
      COMMON /NUMB5/ BETA
      COMPLEX TF4(131072)
      SBANDW=FLOAT(LSAMPL)*BAUD
      SLP=2.0*3.141592*TD
      PHI1=(2.0*3.141592*BETA)/360.0
      PHI=((SLP)*(CF))-PHI1
      N01=LDIM/2
      N02=N01+1
      DO 1 I=1,N01
      TF4(I)=CMPLX(COS(PHI),SIN(PHI))
    1 CONTINUE
      DO 2 I=N02,LDIM
      TF4(I)=CMPLX(COS(PHI),SIN(PHI))
    2 CONTINUE
      RETURN
      END
```

```
C
C*****************************************************************
C      THIS SUBROUTINE DRAWS PROBABILITY OF ERROR CURVES FOR
C      P(E) AS LOW AS 1.0E-10 AND A EB/NO RATIO AS HIGH AS 36 DB.
C*****************************************************************
       SUBROUTINE DRAW(XARRAY,YARRAY,JCURV,JLAST)
       DIMENSION XARRAY(1),YARRAY(1),X(20),Y(20)
       COMMON /NUMB2/ BAUD,ALPHA,FBW
       COMMON /NUMB3/ NSYMB,NSNR,NERROR
       COMMON /NUMB4/ PER
       COMMON /NUMB5/ BETA
       M=NSNR
       DO 1 K=1,NSNR
       IF(YARRAY(K).LE.1.0E-10) GOTO 3
       GOTO 1
     3 M=K-1
       GOTO 4
     1 CONTINUE
     4 CONTINUE
       I=M+1
       II=I+1
       XARRAY(I)=0.0
       XARRAY(II)=2.0
       YARRAY(I)=1.E-10
       YARRAY(II)=0.4
       IF(JCURV.GT.1)GO TO 2
C
C     ESTABLISH THE SURFACE AREA.
C
       CALL PLOTS(30.0,27.5)
C
C      ESTABLISH THE ORIGIN.
C
       CALL PLOT(3.0,2.0,-3)
C
C      DRAW THE LOGARITHMIC Y-AXIS.
C
       CALL LGAXS(0.0,0.0,20HPROBABILITY OF ERROR,20,25.5,
      +90.,1.0E-10,.4)
C
C      DRAW THE LINEAR X-AXIS.
C
       CALL AXIS(0.0,0.0,11HEB/NO IN DB,-11,
      +18.0,0.0,0.0,2.0)
C
C      WRITE THE TITLE OF THE GRAPH.
C
       CALL SYMBOL(8.0,24.0,0.49,16H64 QAM COMBINING,0.0,16)
       CALL SYMBOL(8.0,23.0,0.28,15HTD=((0.00125)N+,0.0,15)
       CALL NUMBER(999.,999.,0.28,PER,0.0,1)
       CALL SYMBOL(999.,999.,0.28,10H*(0.25))NS,0.0,10)
       CALL SYMBOL(8.0,22.5,0.28,6HALPHA=,0.0,6)
       CALL NUMBER(999.,999.,0.28,ALPHA,0.0,1)
       CALL SYMBOL(8.0,22.0,0.28,7HCF=4GHZ,0.0,7)
       CALL SYMBOL(8.0,21.5,0.28,5HBETA=,0.0,5)
       CALL NUMBER(999.,999.,0.28,BETA,0.0,1)
C      PLOT THE IDEAL CURVE.
       DO 5 J=1,18
       X(J)=FLOAT(J)
     5 Y(J)=0.291666*ERFC(SQRT((10.**(.1*X(J)))/3.5))
       X(19)=0.0
       X(20)=2.0
       Y(19)=1.E-10
       Y(20)=0.4
       CALL LGLIN(X,Y,18,1,2,JCURV,1)
     2 CONTINUE
C
C      PLOT THE DATA IN LOG-LINEAR MODE.
C
       CALL LGLIN(XARRAY,YARRAY,M,1,2,JCURV,1)
       IF(JCURV.EQ.JLAST) CALL PLOT(0.0,0.0,999)
       RETURN
       END
```

```
C*************************************************************************
C       THIS SUBROUTINE DRAWS THE EYE DIAGRAM FOR A
C       SYMBOL DURATION.
C*************************************************************************
        SUBROUTINE EYEQ(DATA)
        DIMENSION DATA2(64),DATA3(64)
        COMPLEX DATA(131072)
        COMMON /NUMB1/ LDIM,LSAMPL
        COMMON /NUMB2/ BAUD,ALPHA,FBW
        COMMON /NUMB3/ NSYMB,NSNR,NERROR
        COMMON /NUMB4/ PER
C       ESTABLISH THE SURFACE AREA.
        CALL PLOTS(35.0,27.5)
C       ESTABLISH THE ORIGIN.
        CALL PLOT(3.0,13.0,-3)
C       WRITE THE TITLE OF THE GRAPH.
        CALL SYMBOL(4.0,13.0,0.49,17H64QAM EYE DIAGRAM,0.0,17)
        CALL SYMBOL(4.0,12.5,0.28,15HTD=((0.00000)N+,0.0,15)
        CALL NUMBER(999.,999.,0.28,PER,0.0,1)
        CALL SYMBOL(999.,999.,0.28,10H*(0.25))NS,0.0,10)
        CALL SYMBOL(4.0,12.0,0.28,6HALPHA=,0.0,6)
        CALL NUMBER(999.,999.,0.28,ALPHA,0.0,1)
        CALL SYMBOL(4.0,11.5,0.28,7HCF=4GHZ,0.0,7)
C       DRAW THE TIME AXIS.
        CALL AXIS(0.0,0.0,1H ,-1,17.,0.0,0.0,1.0)
C       DRAW THE AMPLITUDE AXIS.
        CALL AXIS(0.0,-12.0,1H ,1,24.0,90.0,-12.0,1.0)
C       PLOT THE DATA.
        JJ=1*LSAMPL
        DATA2(JJ+1)=0.0
        DATA2(JJ+2)=1.0
        DATA3(JJ+1)=0.0
        DATA3(JJ+2)=1.0
C       M2=NSYMB/1
        M3=256
        M4=M3-255
        M5=M3+256
        DO 4 KK=M4,M5
        DO 2 I=1,JJ
        II=(KK-1)*1*LSAMPL+I
        DATA2(I)=REAL(DATA(II))
      2 DATA3(I)=FLOAT(I)
        CALL LINE(DATA3,DATA2,JJ,1,0,0)
      4 CONTINUE
        CALL PLOT(0.0,0.0,999)
        RETURN
        END
C
C*************************************************************************
C       THIS SUBROUTINE SHIFTS THE DATA SAMPLES 'IT' POSITIONS.
C*************************************************************************
        SUBROUTINE TIME(DATA,IT)
        COMMON /NUMB1/ LDIM,LSAMPL
        COMPLEX DATA(131072),DAT(32)
        DO 10 I=1,IT
        DAT(I)=DATA(I)
     10 CONTINUE
        J1=IT+1
        DO 20 J=J1,LDIM
        DATA(J-IT)=DATA(J)
     20 CONTINUE
        K1=LDIM-IT
        DO 30 K=1,IT
        K2=K1+K
        DATA(K2)=DATA(K)
     30 CONTINUE
        RETURN
        END
C
C*************************************************************************
C       THIS ROUTINE ATTENUATE THE DATA BY AA
C*************************************************************************
        SUBROUTINE ATA(DATA,AA)
        COMMON /NUMB1/ LDIM,LSAMPL
        COMPLEX DATA(131072)
        DO 10 I=1,LDIM
     10 DATA(I)-DATA(I)*AA
        RETURN
        END
```

```
C****************************************************************
C       PROGRAM TO SIMULATE THE P(E) PERFORMANCE OF 256 QAM
C       SECOND-ORDER DIVERSITY COMBINING SYSTEMS
C       CF    = RF CARRIER FREQUENCY
C       TD    = TIME DELAY BETWEEN RF RECEIVED SIGNALS
C       BETA  = PHASE ERROR IN PHASE CORRECTION
C       AA    = EQUIVALENT ATTENUATION OF THE COMBINED SIGNAL
C       THA   = THRESHOLD ADJUSTMENT PARAMETER
C****************************************************************
        COMMON /NUMB1/ LDIM,LSAMPL
        COMMON /NUMB2/ BAUD,ALPHA,FBW
        COMMON /NUMB3/ NSYMB,NSNR,NERROR
        COMMON /NUMB4/ PER
        COMMON /NUMB5/ BETA
        COMPLEX DATA1(131072),TFTX(131072),DATA2(131072),TFRX(131072)
        COMPLEX DATA3(131072),TFPC(131072),TFTD(131072),DATA4(131072)
        DIMENSION NI(2048),NQ(2048),EBNO(36),PEI(36)
        DIMENSION XARRAY(38),YARRAY(38)
C
C       INITIALIZE PROGRAM.
C
        BETA=0.00
        NERROR=0
        LSAMPL=64
        LDIM=131072
        BAUD=11.25
        ALPHA=0.4
        FBW=5.625
        NSYMB=2048
        NSNR=36
        NRUNS=9
        CF=4000.0
        PER=10.0
        T1=(0.00125)/((2.0)*(10.0**3.0))
        T2=((0.25)/(10.0**3.0))*PER
        WRITE(5,10)
    10  FORMAT(5X,'SIMULATION OF 256-QAM COMBINING'/)
        WRITE(5,11) LSAMPL,LDIM,BAUD,ALPHA
    11  FORMAT(5X,'SAMPLES PER SYMBOL=',I4,/5X,'LDIM=',I7,/
       &5X,'SYMBOL RATE=',F7.2,/5X,'ALPHA=',F7.2,/)
C
C       END OF INITIALIZATION.
C
C       START OF COMPUTATIONS.
C
        CALL RCOSTX(TFTX)
        CALL RCOSRX(TFRX)
        CALL HHGG(TFRX,PNOISE)
C
        PNOISE=PNOISE*0.5
        WRITE(5,156)PNOISE
   156  FORMAT(10X,'PNOISE NOW =',F7.3,/)
C
C       CHANNEL 1
C
        CALL LO256(DATA1,NI,NQ)
        CALL FILTER(DATA1,TFTX)
        CALL ENERGY(DATA1,EB)
        WRITE(5,155)EB
   155  FORMAT(10X,'EB=',E15.8)
C
        DO 158 KM=1,NRUNS
C
C       CHANNEL 2
C
        DO 21 I=1,LDIM
        DATA2(I)=DATA1(I)
    21  CONTINUE
        V=FLOAT(KM)
        TD=(T2)+(T1*(V-1.0))
        AA=1.0-(0.04)*V
        THA=1.0000000
        IT=((KM)*2)+(2)
        CALL TDEL(TD,CF,TFTD)
        CALL FILTER(DATA2,TFTD)
        CALL PHCO(TD,CF,TFPC)
        CALL FILTER(DATA2,TFPC)
C       CALL TIME(DATA2,IT)
```

```
C
C       ADD THE TWO DATA SIGNALS
C
        DO 401 I=1,LDIM
  401   DATA3(I)=(DATA1(I)+DATA2(I))/2.000000
C       CALL ATA(DATA3,AA)
C
C       DECODING  (COMBINED CHANNEL)
C
        CALL FILTER(DATA3,TFRX)
        CALL SYNCRC(DATA3,PNOISE,NI,NC,MI,MQ,EB,THA)
C       CALL EYED(DATA3)
        CALL DECO(DATA3,PNOISE,MI,MQ,NI,NQ,EBNC,FEI,EB,THA)
C
C       CALL THE DRAWING ROUTINE
C
        DO 157 I=1,NSNR
        XARRAY(I)=EBNO(I)
  157   YARRAY(I)=FEI(I)
        CALL DRAW(XARRAY,YARRAY,KM,NRUNS)
  158   CONTINUE
        STOP
        END
C
C**********************************************************************
C       THIS SUBROUTINE GENERATE A BASEBAND 256-QAM SIGNAL
C**********************************************************************
C
        SUBROUTINE LD256(DATA,NI,NQ)
        COMPLEX DATA(1)
        DIMENSION NI1(2048),NQ1(2048),NX(38),NY(38),NI(1),NQ(1)
        COMMON /NUMB1/ LDIM,LSAMPL
        DATA IT1,IT2,IT3,IT4/2,11,0,0/
        DATA MREG/11/
C
        NO=MREG
        NM=NO-1
        N1=NO+1
        N2=NO+2
        N3=NO+3
        N4=NO+4

C
        NMX=2**NO
        NMX1=NMX-1
        NO2=IFIX(FLOAT(NO)/2.)
C
C       INITIALIZE THE SHIFT REGISTER
C
        DO 34 I=1,NC
        IF(I.LT.NO2) NX(I)=1
        IF(I.GE.NO2) NX(I)=-1
        NY(I)=-NX(I)
   34   CONTINUE
C
C       PN SEQUENCE GENERATION
C
        NX(N3)=1
        NY(N3)=1
        NX(N4)=1
        NY(N4)=1
        DO 36 K=1,NMX1
        NX(N1)=NX(IT1)
        NX(N2)=NX(IT2)
        IF(IT3.GT.0) NX(N3)=NX(IT3)
        IF(IT4.GT.0) NX(N4)=NX(IT4)
        NY(N1)=NY(IT1)
        NY(N2)=NY(IT2)
        IF(IT3.GT.0) NY(N3)=NY(IT3)
        IF(IT4.GT.0) NY(N4)=NY(IT4)
        DO 37 I1=1,NM
        NX(N1-I1)=NX(NO-I1)
        NY(N1-I1)=NY(NO-I1)
   37   CONTINUE
        NX(1)=NX(N1)*NX(N2)*NX(N3)*NX(N4)
        NY(1)=NY(N1)*NY(N2)*NY(N3)*NY(N4)
        NI1(K)=NX(1)
        NQ1(K)=NY(1)
   36   CONTINUE
        NI1(NMX)=-1
        NQ1(NMX)=-1
```

```
C
C     GENERATE THE 256 QAM DATA
C
      K1=1
      K2=3
      K3=5
      K4=7
      M1=8
      M2=4
      M3=2
      M4=1
      DO 100 I2=1,NMX
      IF(K4.GT.NMX) K4=1
      IF(K3.GT.NMX) K3=1
      IF(K2.GT.NMX) K2=1
      NI(I2)=NI1(K1)*M1+NI1(K2)*M2+NI1(K3)*M3+NI1(K4)*M4
      NQ(I2)=NQ1(K1)*M1+NQ1(K2)*M2+NQ1(K3)*M3+NQ1(K4)*M4
C
C     LOAD INTO SAMPLE ARRAY, LSAMPL SAMPLES PER SYMBOL
C
      J1=(I2-1)*LSAMPL+1
      J2=I2*LSAMPL
      DO 101 J=J1,J2
      DATA(J)=CMPLX(FLOAT(NI(I2)),FLOAT(NQ(I2)))
  101 CONTINUE
      K1=K1+1
      K2=K2+1
      K3=K3+1
      K4=K4+1
  100 CONTINUE
      RETURN
      END
```

APPENDIX B

PERFORMANCE DEGRADATION CAUSED BY IMPERFECT PHASE CORRECTION IN QAM RF AND IF SOSDCS

In this appendix we analyze the system performance degradation caused in 64 and 256 QAM SOSDCS by compensating the band-center-frequency phase difference between the RF or IF combining signals not exactly by $2\pi f_c \tau$ but $(2\pi f_c \tau - \beta)$, where β is a relatively small phase error due to imprecisions in analog phase shifters or inherent to the nature of discrete phase shifters.

Phase compensation with phase error β makes the complex envelope corresponding to equation 3.1 to change to:

$$H_{C1}(f) = \begin{cases} 1 + e^{-j(2\pi f\tau + \beta)} & ; -(BW/2) \le f \le (BW/2) \\ 0 & ; \text{elsewhere} \end{cases} \qquad (B.1)$$

Therefore, $i_o(t)$ and $q_o(t)$ reduce to:

$$i_o(t) = i(t) + \cos\beta \; i(t-\tau) + \sin\beta \; q(t-\tau) \qquad (B.2)$$

$$q_o(t) = q(t) + \cos\beta \; q(t-\tau) - \sin\beta \; i(t-\tau) \qquad (B.3)$$

i.e., a β-dependent effect, of the same type observed when no phase compensation is used, appears (compare equations B.2 and B.3 with equations 3.7 and 3.8, respectively). It can be noted in equations B.2 and B.3 that degradation exists even if $\tau = 0$ and that increasing degradation must be expected if larger values of τ or β are considered.

Inclussion of a phase error β in phase compensation devices was considered in our computer program, as it is explained in Chapter IV. Simulation was made for different values of τ and β for 64 and 256 QAM RF and IF SOSDCS with f_c = 4 GHz, f_b = 90 Mb/s and $\alpha = 0.4$.

In figures B.1 and B.2, equivalent E_b/N_o or C/N degradation curves at $P_b(e) = 10^{-6}$ are presented for those systems. As expected, increasing degradation with respect to perfect phase compensation is observed when larger values of β are considered.

These results, in conjunction with the required performance in a practical system and the statistical knowledge of τ, can be used to specify the phase shifter needed in the system.

Let us suppose, for example, that the required performance in the system allows equivalent C/N degradation caused by τ at $P_b(e) = 10^{-6}$, to be greater than 0.5 dB only during 0.01% of time. The value of τ that is exceeded only during this percentage of time is needed and it can be determined from its statistics. Let us assume that in our particular case this value is 7 ns. In figures B.1 and B.2 it can be appreciated that RF or IF SOSDCS with phase compensation can be used for both, systems using 64 or 256 QAM. The phase imprecision that can be allowed

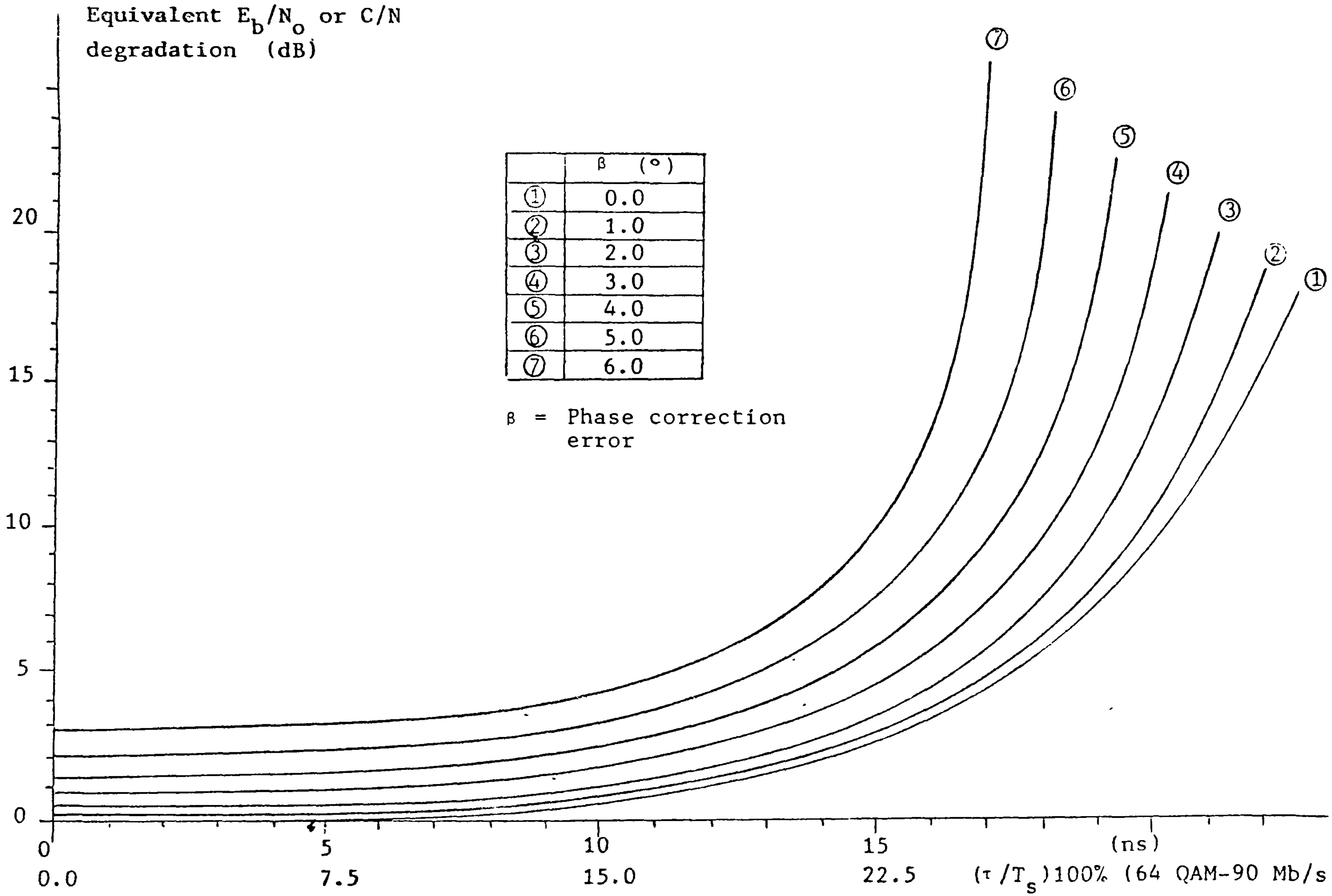


Fig. B.1 Equivalent degradation of E_b/N_o or C/N for RF and IF 64 QAM - SOSDCS with imperfect phase compensation

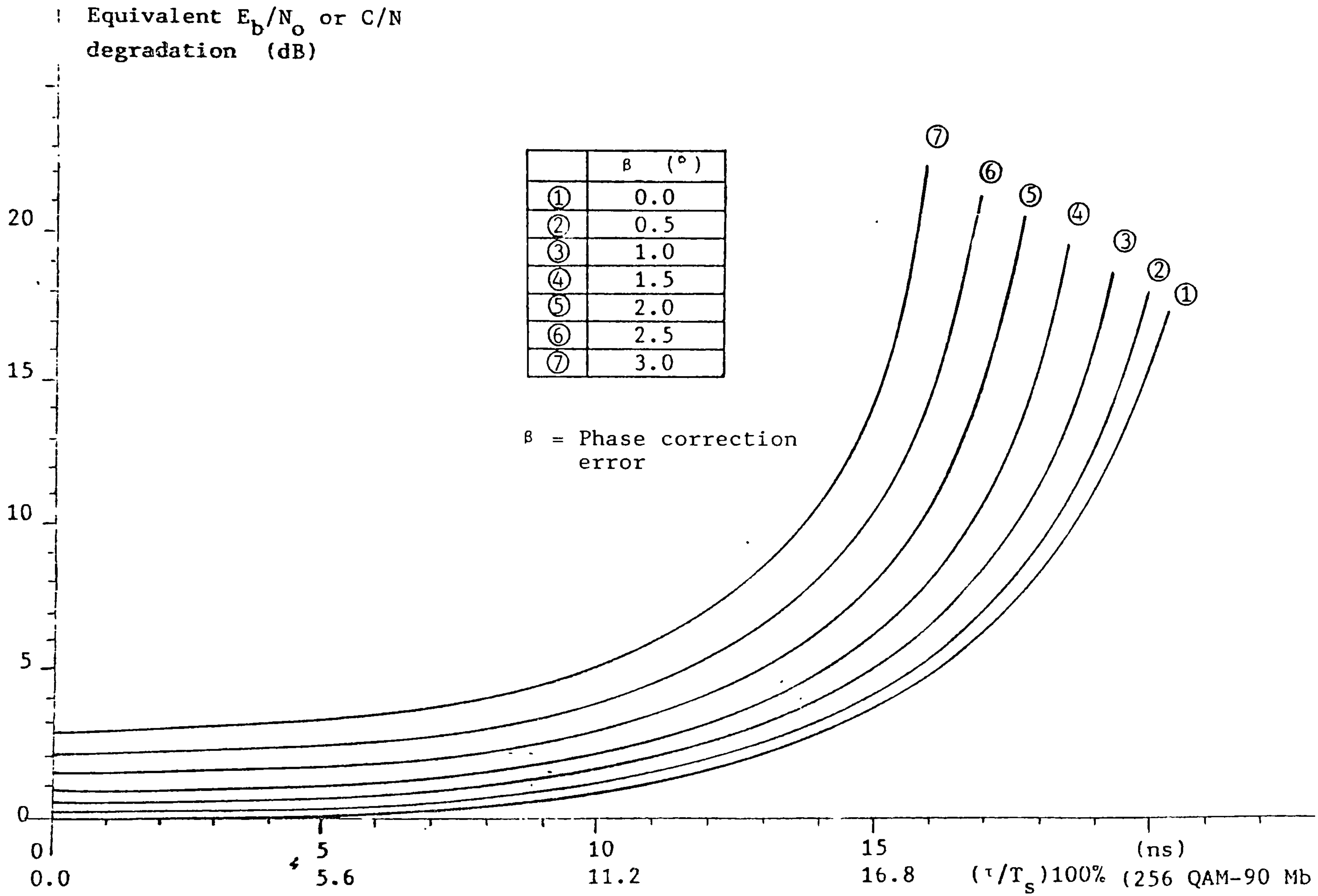


Fig. B.2 Equivalent degradation of E_b/N_o or C/N for RF anf IF 256 QAM SOSDCS with imperfect phase compensation is, at most, $\beta = 2.0$ degree if the modulation scheme is 64 QAM and $\beta = 0.5$ degree if 256 QAM is used.

The independence of f_c exhibited by equations B.2 and B.3, and representation of τ as a percentage of T_s (the system symbol duration), enable our results of figures B.1 and B.2 to be used for 64 and 256 QAM RF and IF SOSDCS in general.

APPENDIX C

PERFORMANCE EQUIVALENCE BETWEEN IF WIDEBAND PHASE SHIFTERS AND RF SINGLE FREQUENCY PHASE SHIFTERS IN COPHASING IF COMBINING SIGNALS

When combining of signals is made at RF, it is necessary to employ an RF wideband adaptive phase shifter (WAPS), as illustrated in figure C.1, to properly shift the frequency components in the RF signal and obtain the performance improvement indicated in Chapter III. In IF combining, the adequate phase shift across the signal bandwidth can be obtained also by using an IF WAPS, as illustrated in figure C.2, but in this case the fact that when a signal is shifted in the frequency domain, by the action of a sinusoidal tone, its phase characteristic is affected in accordance with the tone phase [30], suggests the possibility of implementing IF combining systems in which a wideband adaptive phase compensation is obtained by properly shifting the relative phase of the sinusoidal down-converter (D/C) tones employed in the system, as illustated in figure C.3, where a single frequency adaptive phase shifter (SFAPS) is used instead of a WAPS.

Next we demonstrate the performance equivalence between the schemes with SFAPS presented in figure C.3 and the one with WAPS illustrated in figure C.2. Although our discussion concerns to systems employing QAM, demonstration is made in a general way and therefore it is valid for combining systems using any kind of modulation.

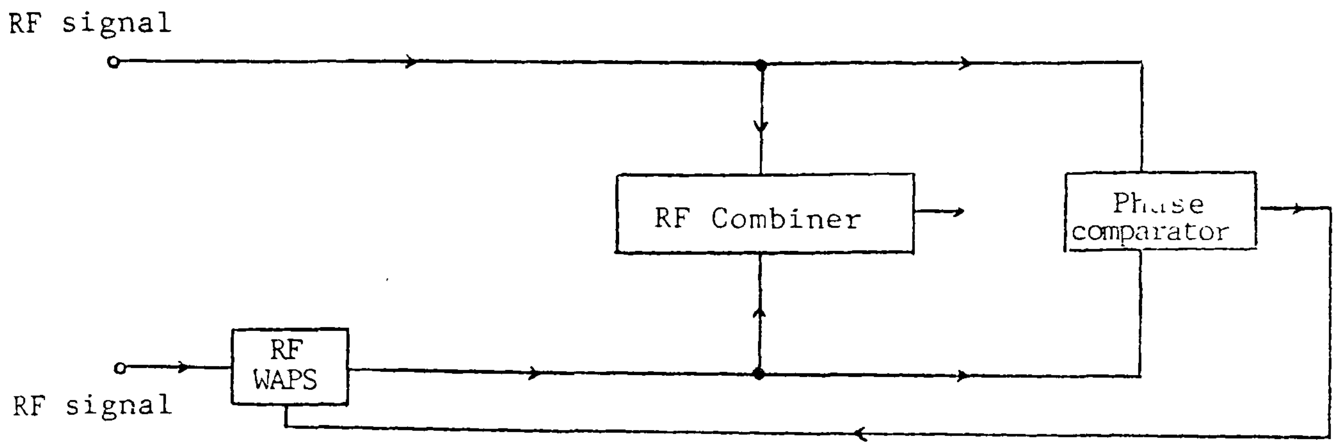


Fig. C.1 RF Combiner with WAPS

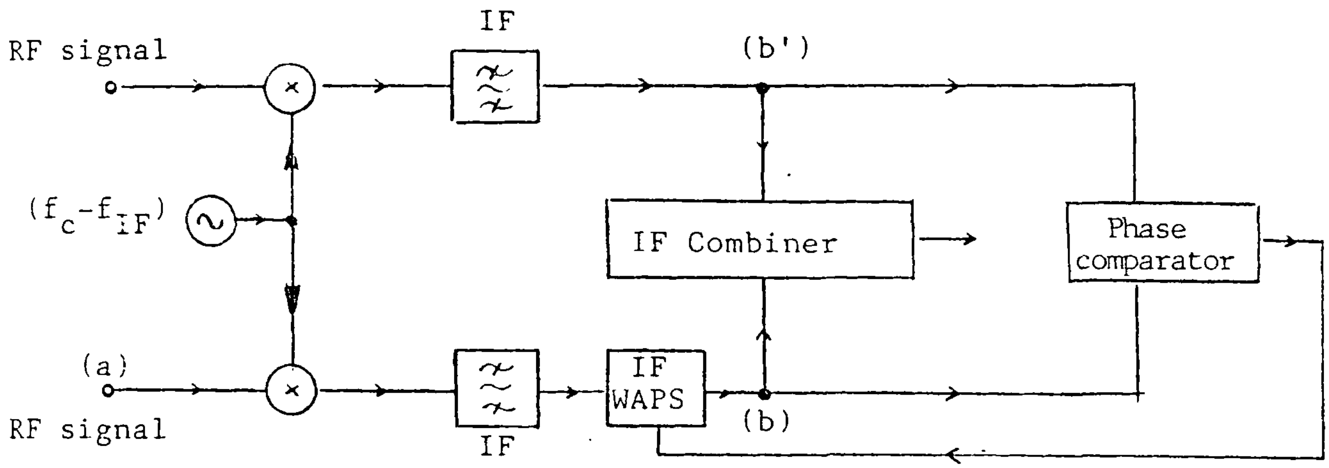


Fig. C.2 IF Combiner with WAPS

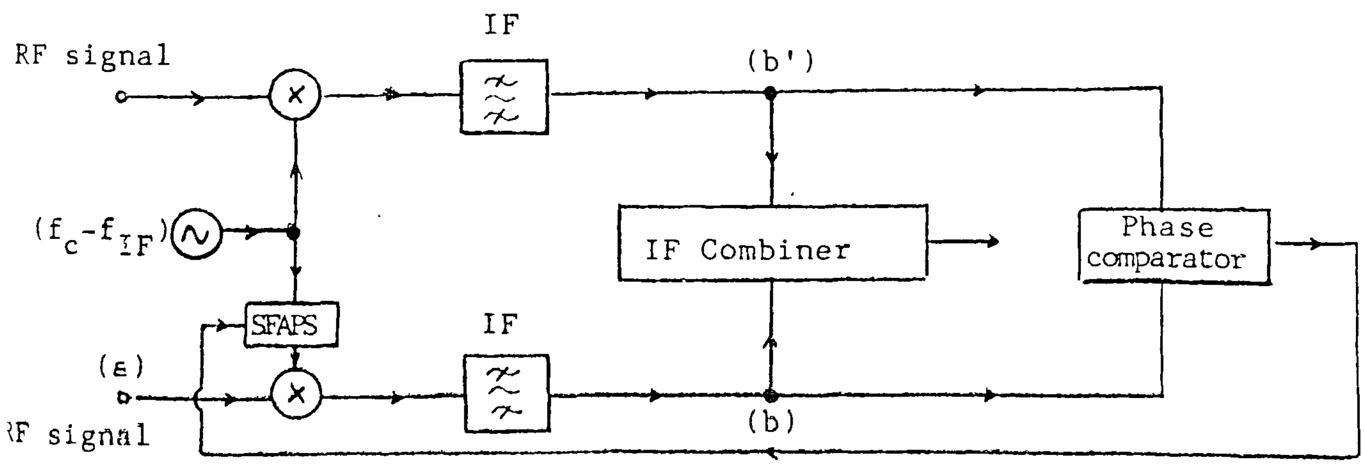


Fig. C.3 IF Combiner with SFAPS

DEMONSTRATION

Systems in figures C.3 and C.2 will have the same performance, if the received signals at points (b) and (b') are respectively the same in both cases.

Let us call x(t) the band-limited RF received signals at point (a) in both cases. In the system with WAPS, the signal $y_1(t)$ at point (b) is then:

$$y_1(t) = [x(t) \cos(\omega_c - \omega_{IF})t] * h_1(t) * h_2(t) \qquad (C.1)$$

In the system with SFAPS, the signal $y_2(t)$ at point (b) is:

$$y_2(t) = \{x(t) \cos[(\omega_c - \omega_{IF})t - \phi]\} * h_1(t) \qquad (C.2)$$

In equations C.1 and C.2, ω_c and ω_{IF} are the RF and IF angular frequencies respectively, $h_1(t)$ is the impulse response of an IF filter, $h_2(t)$ is the impulse response of an ideal WAPS device (for a certain phase adjustment value ϕ), and "*" indicates convolution operation between two functions.

Equation C.1 can be interpreted in the frequency domain as:

$$Y_1(f) = (1/2)\{X(f) * [\delta(f-\Delta) + \delta(f+\Delta)]\} H_1(f) H_2(f)$$

$$= (1/2)[X_-(f+\Delta) + X_-(f-\Delta) + X_+(f+\Delta) + X_+(f-\Delta)] H_1(f) H_2(f) \qquad (C.3)$$

Where $X(f)$, $H_1(f)$ and $H_2(f)$ are the Fourier transform of the respective functions, $\delta(f)$ is the Dirac function, $\Delta = f_c - f_{IF}$ and:

$$X_+(f) = X(f)\ u(f) \qquad (C.4)$$

$$X_-(f) = X(f)\ u(-f) \qquad (C.5)$$

Where $u(f)$ is the unit-step function.

The objective of the IF bandpass filter $H_1(f)$ is to remove the higher order components of the spectrum. Therefore, $Y_1(f)$ can be written as:

$$Y_1(f) = (1/2)\ [X_-(f-\Delta) + X_+(f+\Delta)]\ H_2(f) \qquad (C.6)$$

$H_2(f)$ can be expressed as:

$$H_2(f) = e^{j\phi}\ u(f) + e^{-j\phi}\ u(-f) \qquad (C.7)$$

Hence, we can write C.6 as:

$$Y_1(f) = (1/2)\ [X_-(f-\Delta)\ e^{-j\phi} + X_+(f+\Delta)\ e^{j\phi}] \qquad (C.8)$$

And this means that, as expected, a constant phase shift (ϕ) across the IF bandwidth has been introduced.

Note that when $\phi = 0$, equation C.8 reduces to:

$$Y_1(f) = (1/2)\ [X_-(f-\Delta) + X_+(f+\Delta)] \qquad (C.9)$$

Which is the corresponding IF signal, at point (b), in a system without phase compensation.

On the other hand, in the frequency domain, equation C.2 corresponds to:

$$Y_2(f) = (1/2)\ \{X(f) * [\delta(f-\Delta)\ e^{-j\phi} + \delta(f+\Delta)\ e^{j\phi}]\} H_1(f)$$

$$= (1/2)\ [X_-(f+\Delta)\ e^{j\phi} + X_-(f-\Delta)\ e^{-j\phi}$$

$$+ X_+(f+\Delta)\ e^{j\phi} + X_+(f-\Delta)\ e^{-j\phi}\ ]\ H_1(f) \qquad (C.10)$$

Considering now that the IF bandpass filter $H_1(f)$ removes the higher order components of the spectrum, we have:

$$Y_2(f) = (1/2)\ [X_-(f-\Delta)\ e^{-j\phi} + X_+(f+\Delta)\ e^{j\phi}\ ] \qquad (C.11)$$

And, again, this means that a constant phase shift (ϕ) across the IF bandwidth has been introduced.

Equations C.8 and C.11 are identical and this implies that the received signal at point (b) is the same in both systems.

That the received signal at point (b') in both systems is the same, for the same RF main received signal, can be noted by simple inspection of figures C.3 and C.2.

Hence, IF diversity combining systems using WAPS and those with SFAPS, have the same performance. The less difficult to design should be prefered in practice.

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