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SUPERPOSED QUADRATURE AMPLITUDE MODULATION (SQAM).  
A SPECTRAL AND POWER EFFICIENT MODULATION TECHNIQUE

by

JONGSOO, SEO

A THESIS
PRESENTED TO THE SCHOOL OF GRADUATE STUDIES AND RESEARCH
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IN
ELECTRICAL ENGINEERING

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ABSTRACT

A spectral and power efficient modulation technique - Superposed Quadrature Amplitude Modulation (SQAM) is introduced.

At first, the SQAM baseband signal encoder, using both the pulse overlapping concept and the nonlinear switching filter (NLSF) technique, is described. Experimental and simulation results of eye patterns, state space diagrams and envelope fluctuations of SQAM signals are presented. The spectral characteristics and probability of error performance of SQAM signals, in an additive white Gaussian noise (AWGN) single-channel environment, are analyzed and compared to constant envelope modulation schemes such as MSK and TFM. These results indicate that our SQAM systems have comparable spectral properties, exhibit better P(e) performance and are easier to implement than TFM.

Then the performance of the SQAM system, in a nonlinearly amplified multichannel environment, in the presence of AWGN, intersymbol-interference (ISI), adjacent channel-interference (ACI) and/or cochannel interference (CCI), is investigated and compared to those of OQPSK, MSK and IQF-OQPSK (or SQORC) systems. The effects of ACI and CCI on the performance of the SQAM system are investigated by
simulation under different channel conditions, namely, (1) different channel frequency separation, (2) filter bandwidth ($B$) and symbol duration ($T_s$) product ($BT_s$), (3) flat fade depth and (4) different values of amplitude parameter ($A$) in the SQAM signal. In the simulation, the earth station high power amplifiers (HPA), operating in saturation mode, are approximated by ideal hard limiters. As a result of the performance evaluation, it has been found that the SQAM system outperforms QPSK, MSK and IIF-OQPSK systems. Further, we illustrate that the $P(e)$ performance degradation due to the ACI or CCI can be minimized by appropriately choosing the modulation technique, type of filters (transmit and receive) and their $BT_s$ values.

For future research, an application of the SQAM signal encoding technique to the premodulation filters, in digital FM or continuous phase frequency shift keying (CPFSK) scheme, is suggested. To attain even higher spectral efficiency, a duo-binary SQAM technique is also suggested and its power spectrum is studied.

These desirable performance characteristics, combined with the simple hardware implementations of SQAM modem may lead to numerous satellite and terrestrial radio system applications.
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Chapter I
INTRODUCTION

In order to improve the power and spectral efficiency of digital communication systems, various filtering and baseband signal processing techniques have been studied. To attain high spectral efficiency, the power spectrum of the signal should have a narrow mainlobe and a fast spectral roll-off. To attain high power efficiency, the transmit high power amplifier (HPA) should be operated in a saturated (nonlinear) mode.

As the presently used RF spectrum becomes highly congested, many satellite systems tend to employ higher frequency bands such as 14/11 or 30/20 GHz. At these frequencies, not only up- and down-link thermal noise, but also up- and down-link fades, caused by rain attenuations, are quite important. To minimize the impact of the uplink signal fade, the satellite traveling wave tube amplifiers (TWTA) may be overdriven into the nonlinear region. To maximize the earth station power efficiency, the earth station HPAs must be operated in the saturation region. An ideal hardlimiter presents a good first-order approximation of saturated amplifiers. [1,10].
Quadrature-phase-shift-keying (QPSK) modulation technique is widely used because of its simple hardware implementation and good $P(e)$ performance in a linearly amplified channel. However, when a QPSK signal is filtered and then amplified by HPA near saturation region, the spectral sidelobes at the output of HPA regrow due to the nonlinear amplification. These regrown spectral sidelobes cause significant interference into the adjacent channels, and degrade the probability of error performance.

In PSK-type modulation schemes, multi-interval pulse overlapping techniques have been introduced to achieve higher spectral efficiency compared to the conventional QPSK, OQPSK and MSK signals. A class of intersymbol interference and jitter-free (IIF) -OQPSK scheme studied by Feher et al. [1,4,10,11] also retains a compact mainlobe and low spectral spreading after nonlinear amplification.

In this thesis, we introduce a new class of power and spectral efficient modulation technique - Superposed Quadrature Amplitude Modulation (SQAM) which produces a fast spectral roll-off, a low out-of-band energy, and shows good $P(e)$ performance.
1.1 THESIS OUTLINE

Following this introductory chapter, in Chapter 2, two methods of generation of SQAM baseband signals are presented, namely, (1) the pulse overlapping concept and (2) the nonlinear switching filter (NLSF) technique. The NLSF method is used for our low-speed hardware prototype. For higher speed applications, binary transversal filter (BTF) implementations may be preferable [10]. The power spectral densities of SQAM signals are analyzed and compared to those of QPSK (or OQPSK), MSK, IIF-OQPSK (or SQORC) and TFM signals [1,4,6].

Chapter 3 illustrates the properties of SQAM signals, including eye patterns, signal state-space diagrams and envelope fluctuations. The probability of error performance of the SQAM modem, in an AWGN single-channel environment, is studied and compared to some well-known, constant envelope modulation schemes, such as MSK and TFM.

In Chapter 4, the performance of SQAM modem, in a nonlinearily amplified multi-channel environment, in the presence of additive white Gaussian noise (AWGN), intersymbol-interference (ISI), adjacent channel interference (ACI) and/or co-channel interference (CCI), is
investigated by computer simulation, and compared to the performance of OQPSK, MSK and IJP-OQPSK modems. The effects of ACI and CCI on the performance of SQAM modem are investigated under different channel conditions, such as different channel frequency separation, filter bandwidth and symbol duration product ($BT_s$) and flat fade depth of the desired signal. In this computer simulation, cascaded nonlinear satellite systems are simplified to a back-to-back mode (i.e., no channel input filter or TWTA), and earth station HPAs, operating in saturation mode, are approximated by the ideal hardlimiters.

Chapter 5 contains recommended further research, where applications of the SQAM signal processing technique to digital FM or duo-binary schemes are suggested.

Chapter 6 contains final conclusions.
Chapter II
A CLASS OF POWER AND SPECTRAL EFFICIENT SQAM SIGNALS

It has been demonstrated [4,6,10] that overlapping baseband pulses, where each pulse duration is wider than one symbol interval \(T_s\), have spectral advantages over non-overlapping pulses. Smooth and correlated phase transitions are required to attain good spectral properties, and low envelope fluctuations (or constant envelope) of modulated signals are preferable to attain good power efficiency.[3,20,26].

In this chapter, a new class of signals, for applications in a power and spectral efficient system, is introduced and studied. The SQAM baseband signals can be generated by using either the pulse overlapping concept or the NLSF technique.[5,10]. Conceptually, both implementations lead to identical results. For our relatively low-speed (128 kb/s) prototype hardware design, we used the NLSF method. The phase transitions and power spectrum of the SQAM signals are analyzed and compared to other modulation schemes.
2.1 GENERATION OF SQAM SIGNALS

2.1.1 SQAM Signal Encoder Using Pulse Overlapping Concept

Fig. 2.1 shows the SQAM signal encoder using a double-interval pulse overlapping concept [4,6,10]. The input signal \( x(t) \) to the encoder is given by:

\[
x(t) = \sum_{n=-\infty}^{\infty} a_n h(t-nT_s)
\]

where

\[
a_n = \pm 1 \text{ with probability } 1/2
\]

\[
h(t-nT_s) = \begin{cases} 
1 & \text{for } t=nT_s \\
0 & \text{otherwise}
\end{cases}
\]

and \( T_s \) is a data symbol duration. Let the SQAM signal encoder have an impulse response \( s(t) \), which is a double-interval \( (2T_s) \) raised-cosine pulse, superimposed with another weighted single-interval \( (T_s) \) raised-cosine pulse, i.e.,

\[
s(t) = \frac{1}{2} \left( 1 + \cos \frac{\pi t}{T_s} \right) + d(t)
\]

where

\[
d(t) = -\frac{1-A}{2} \left( 1 - \cos \frac{2\pi t}{T_s} \right)
\]

\[
0.5 \leq A \leq 1.5
\]

\[-T_s \leq t \leq T_s\]
and $A$ is an amplitude parameter of the SQAM signal. The encoded signal $y(t)$ is obtained by the convolution of the input signal $x(t)$ with the impulse response of the encoder $s(t)$, and is given by:

$$y(t) = \sum_{n=-\infty}^{\infty} a_n s(t-nT_S)$$  \hfill (2.5)

In order to obtain an intersymbol-interference and jitter-free (IJF) signal $y(t)$, the double-interval pulse $s(t)$ must satisfy the following IJF conditions.\cite{1}.

$$s(t)' = s(-t) \quad \text{for} \quad -T_S \leq t \leq T_S \quad \hfill (2.6.a)$$

$$s(T_S) = s(-T_S) = 0 \quad \hfill (2.6.b)$$

From (2.3) and (2.4), it can be readily proven that SQAM signals meet the IJF conditions. Since the input signal duration is $T_S$ second and $s(t)$ has a $2T_S$ second duration, the encoded signal can be obtained by the superposition of double-interval pulses $s(t-nT_S)$ and $s(t-(n+1)T_S)$. Therefore, the encoded signal can be rewritten as:

$$y(t) = \sum_{n=-\infty}^{\infty} y_m(t-nT_S)$$  \hfill (2.7)

where $m=1,2,3,4$. 
\[ y_1(t-nT_s) = -s(t-nT_s) - s(t-(n+1)T_s) \quad \text{for } a_n = a_{n+1} = -1 \]
\[ y_2(t-nT_s) = s(t-nT_s) - s(t-(n+1)T_s) \quad \text{for } a_n = 1 \text{ and } a_{n+1} = -1 \]
\[ y_3(t-nT_s) = -s(t-nT_s) + s(t-(n+1)T_s) \quad \text{for } a_n = -1 \text{ and } a_{n+1} = 1 \]
\[ y_4(t-nT_s) = s(t-nT_s) + s(t-(n+1)T_s) \quad \text{for } a_n = a_{n+1} = 1 \]
(2.8)

For the illustrative impulse sequence shown in Fig. 2.1(c), with input symbols \( a_0 = a_1 = 1 \), the encoded SQAM signal, in the interval \([0, T_s]\), is given by:

\[
y_4(t) = s(t) + s(t-T_s) = \left[ \frac{1}{2} \left( 1 + \cos \frac{\pi t}{T_s} \right) + d(t) \right] + \left[ \frac{1}{2} \left( 1 + \cos \frac{\pi (t-T_s)}{T_s} \right) + d(t-T_s) \right] = 1 + 2d(t) = A + (1-A)\cos \frac{2\pi t}{T_s}
\]
(2.9)

In the interval \([T_s, 2T_s]\), with input symbols \( a_1 = 1 \) and \( a_2 = -1 \), the encoded SQAM signal is given by:

\[
y_2(t-T_s) = s(t-T_s) - s(t-2T_s) = \left[ \frac{1}{2} \left( 1 + \cos \frac{\pi (t-T_s)}{T_s} \right) + d(t-T_s) \right] - \left[ \frac{1}{2} \left( 1 + \cos \frac{\pi (t-2T_s)}{T_s} \right) + d(t-2T_s) \right] = -\cos \frac{\pi t}{T_s}
\]
(2.10)
In the interval \([2T_S, 3T_S]\), with input symbols \(a_2 = -1\) and \(a_3 = 1\), the encoded SQAM signal is given by:

\[
y_3(t-2T_S) = -s(t-2T_S) + s(t-3T_S)
\]

\[
= -\left[\frac{1}{2}(1 + \cos \frac{\pi}{T_S}(t-2T_S)) + d(t-2T_S)\right]
\]

\[
+ \left[\frac{1}{2}(1 + \cos \frac{\pi}{T_S}(t-3T_S)) + d(t-3T_S)\right]
\]

\[
= \cos \frac{\pi}{T_S} t
\]  
(2.11)

Applying the same concept with input symbols \(a_n = a_{n+1} = -1\), the encoded signal is given by:

\[
y_1(t-nT_S) = -s(t-nT_S) - s(t-(n+1)T_S)
\]

\[
= -1 - 2d(t)
\]

\[
= -A - (1-A) \cos \frac{2\pi}{T_S} t
\]  
(2.12)

and also

\[
y_4(t-nT_S) = -y_2(t-nT_S)
\]  
(2.13)

\[
y_3(t-nT_S) = -y_2(t-nT_S)
\]  
(2.14)

Thus, the pulse overlapped SQAM signal, in any signaling interval, is a function of two consecutive input symbols, \(a_n\) and \(a_{n+1}\).
(a) SQAM signal encoder having an impulse response $s(t)$.

$$s(t) = g(t) + d(t)$$

(b) An example of SQAM double-interval pulse.

(c) SQAM signal encoding process using pulse overlapping concept.

Fig. 2.1. SQAM signal encoder using double-interval pulse overlapping concept.
2.1.2 SQAM Signal Encoder Using the NLSF Technique.

In this section, the SQAM baseband signal encoder, using the nonlinear switching filter (NLSF) technique [5,10], is described. We have defined four SQAM encoded signals in the $0 \leq t \leq T_s$ interval in (2.9): \ldots (2.12). (See Table 2.1.) Fig.2.2 shows the encoded SQAM signal eye pattern using the NLSF technique. For each consecutive $T_s$ symbol interval, one of these signals is switched to the transmission system. The required switching function is achieved in the channel multiplex (MUX) unit, which is controlled by a logic control unit. The selection of the $y_1(t-nT_s)$ \ldots $y_4(t-nT_s)$ signal states depends upon the states of two consecutive input symbols. In Fig.2.3, the corresponding hardware block diagram is shown, while in Fig.2.4 the detailed circuit diagram of the SQAM baseband signal processor is shown. In Fig.2.4, U1 \ldots U7 consist a logic control unit, which provides 2 bit control signals to a 4 channel MUX (U10). A monostable multivibrator (U6) compensates the time delay caused by the logic control unit. Using a 128kHz input clock pulse, U8,9 generate 64kHz sine (or cosine) components $y_2(t)$ and $y_3(t)$. (See Fig.2.2.) U13,14 generate weighted and d.c biased (i.e., $0.5 \leq A \leq 1.5$) 128kHz sine components $y_1(t)$ and $y_4(t)$. U12 compensates a time delay between the output components $y_1(t)$ \ldots $y_4(t)$. Fig.2.5 illustrates the measured waveshape and eye pattern of the SQAM signal.
Table 2.1 Encoded SQAM output signals.

<table>
<thead>
<tr>
<th>NRZ input data</th>
<th>SQAM output signal</th>
</tr>
</thead>
<tbody>
<tr>
<td>$a_{n+1}$</td>
<td>$a_n$</td>
</tr>
<tr>
<td>-1</td>
<td>-1</td>
</tr>
<tr>
<td>-1</td>
<td>1</td>
</tr>
<tr>
<td>1</td>
<td>-1</td>
</tr>
<tr>
<td>1</td>
<td>1</td>
</tr>
</tbody>
</table>

SQAM output signals $y_1(t)$ to $y_4(t)$ are defined as follows.

\[
y_1(t) = -A - (1-A) \cos(2\pi t / T_s)
\]

\[
y_2(t) = -\cos(\pi t / T_s)
\]

\[
y_3(t) = \cos(\pi t / T_s)
\]

\[
y_4(t) = A + (1-A) \cos(2\pi t / T_s)
\]

where

\[
0.5 \leq A \leq 1.5
\]

\[
0 \leq t \leq T_s
\]
Fig. 2.2 Encoded SQAM output signal eye pattern using the NLSF technique.

Fig. 2.3 Block diagram of SQAM signal encoder using the NLSF technique.
Circuit diagram of SQAM baseband signal processor.
(a). Measured waveshape. (I-channel)

Vert.: 2.5 volt/div.
Hori.: 10.4 us./div.

(b). Measured eye pattern. (I-channel)

Vert.: 2.5 volt/div.
Hori.: 2.2 us./div.

---
1
2A-1
0
-1
---

Fig.2.5 Measured baseband waveshape and eye pattern of a \( f_b = 64 \) kbaud SQAM signal. (Q-channel eye pattern is identical to I-channel.)
Amplitude parameter, \( A = 0.8 \).
Note the ICF condition is fully satisfied.
Fig. 2.6 Phase transitions of MSK, SQAM and IJP-OQPSK ($A = 1.0$).
(d) SQAM (A = 0.9)

(e) SQAM (A = 1.0, IIF-QPSK)

(f) SQAM (A = 1.1)

Fig. 2.6 Phase transitions of MSK, SQAM and IIF-QPSK (A = 1.0).
Fig. 2.6(g) Phase transitions of MSK and SQAM (A=0.8 and 1.0).

Fig. 2.6(h) Phase transitions of MSK, TFM, and CCPSK.
2.3 SPECTRAL ANALYSIS OF SQAM SIGNALS

In this section, the power spectral density (PSD) of SQAM signals is derived and compared with the PSD of some other well-known signals. The PSD function of the SQAM signal is derived by using the Impulse method \[20\] and the Fourier transforms of the SQAM double-interval pulse. From (2.3) and (2.4), the SQAM double-interval pulse is expressed as:

\[ s(t) = g(t) + d(t) \]  \hspace{1cm} (2.15)

where

\[ g(t) = \frac{1}{2} (1 + \cos \pi t/T_s) \]  \hspace{1cm} (2.16)

The frequency spectrum of the SQAM signal is:

\[ S(f) = G(f) + D(f) \]  \hspace{1cm} (2.17)

where \( S(f) \), \( G(f) \) and \( D(f) \) are Fourier transforms of \( s(t) \), \( g(t) \) and \( d(t) \) respectively, and are readily obtained as:

\[ G(f) = F[g(t)] = \frac{T_s}{1 - 4T_s^2 f^2} \sin 2\pi f T_s - \frac{\sin 2\pi f T_s}{2\pi f T_s} \]  \hspace{1cm} (2.18)

and

\[ D(f) = F[d(t)] = \frac{(A-1) T_s}{1 - \frac{\pi^2}{2} f^2} \sin 2\pi f T_s - \frac{\sin 2\pi f T_s}{2\pi f T_s} \]  \hspace{1cm} (2.19)
Therefore

\[ S(f) = T_S \left( \frac{1}{1 - 4T_f^2} + \frac{\delta - 1}{1 - T_f^2} \right) \frac{\sin 2\pi f T_s}{2\pi f T_s} \]  \hspace{1cm} (2.20)

and

\[ S(0) = AT_S \]  \hspace{1cm} (2.21)

Thus, the normalized PSD function of the SQAM signal \( s(t) \) is given by:

\[ \left| \frac{S(f)}{S(0)} \right|^2 = \frac{1}{A^2} \left( \frac{1}{1 - 4T_f^2} + \frac{\delta - 1}{1 - T_f^2} \right) \frac{\sin 2\pi f T_s}{2\pi f T_s} \]  \hspace{1cm} (2.22)

(See Appendix A. for the detailed derivation of eq. (2.22).)

Fig. 2.7 shows the computed power spectra of equiprobable random SQAM baseband signals for different values of the amplitude parameter 'A'. We note that a decrease of 'A' leads to a faster spectral roll-off at higher frequencies, but at the expense of a wider main-lobe. Depending on particular system applications, desirable SQAM signals may be selected based on the trade-off between the main-lobe occupancy and side-lobe roll-off. In Figs. 2.8(a) ... (c), the power spectra of SQAM signals are compared with those of
other constant envelope modulation schemes, such as MSK and TFM, for linear and nonlinear channels [1,3,10]. The detailed computer simulation program is listed in Appendix C. From the results, we note that the SQAM signals have significant spectral advantages over QPSK and MSK signals, and comparable spectral properties to TFM signals. (The spectral region of practical interest is between 0 dB and -35 dB for most applications. ) Figs.2.9(a) ... (c) show measured power spectra of the SQAM signals with amplitude parameters $A = 0.8$, 0.9 and 1.0 in a linear channel. In this experiment, we used a random 128 kb/s equiprobable NRZ signal. The experimental, computer simulation and theoretical results indicate a close agreement. We note that the SQAM signal, designated with $A=1.0$, is identical with the conventional overlapped raised-cosine (QORC) signal and also the IJP-QQPSK (subclass n=1) signal.[1,6].
Fig. 2.7. Normalized power spectral densities of SQAM signals in a linear channel. (Computations based on equation (2.22).)
Fig. 2.8(a). Power spectra of SQAM, MSK and TFK
signals in a linear channel. (The A=1.0 case
is identical to IIF-OQPSK(n=1) described in
[1,11] or QCRC [6].)
Fig. 2.8(b). Power spectra of SQAM and TFM (with truncation length 3T, 5T and 7T) in a linear channel.
Fig. 2.8(c). Power spectra of SQAM, MSK and FSK signals in a nonlinear (hardlimited) channel.
Fig. 2.9. Measured power spectra of $f_s=64\text{kHz}$ ($f_b=128\text{kHz}$) SQAM signals in a linear channel.
A useful measure of a spectral compactness can be obtained from the fractional out-of-band power $P_{OB}$, defined as:

$$
P_{OB} = 1 - \left[ \frac{\int_{-B}^{B} S'(f) \, df}{\int_{-\infty}^{\infty} S'(f) \, df} \right] \quad (2.23)
$$

where

$$
S'(f) = \left| \frac{S(f)}{S(0)} \right|^2 \quad (2.24)
$$

and $B$ is a channel bandwidth normalized to the data symbol rate (i.e., $(f-f_c)/f_S$). Figs. 2.10(a) and (b) show the computed out-of-band to total power ratios of SQAM and MSK signals for linear and nonlinear channels. It is noticed that SQAM signals have a sharper spectral roll-off and lower out-of-band energy, compared to MSK signal. Also note that in the SQAM system, for normalized frequencies $(f-f_c)T_S > 1.0$, the out-of-band energy for $A<1.0$ is lower than that for $A \geq 1.0$. 
Fig. 2.10(a). Out-of-band to total power ratios of SQAM and MSK signals in a linear channel.
Fig. 2.19(b). Out-of-band to total power ratios of SQAM and MSK signals in a nonlinear (hardlimited) channel.
Chapter III

PERFORMANCE OF SQAM IN AWGN SINGLE-CHANNEL ENVIRONMENT

3.1 SQAM MODEM

The block diagram of a SQAM modem is shown in Fig. 3.1. The SQAM modulator consists of SQAM signal encoders and a conventional Offset-QPSK modulator structure. For this modulator, we do not require additional spectral shaping filters. Two encoded (i.e., in-phase and quadrature) SQAM baseband signals $y_{\text{Im}}(t)$ and $y_{\text{Qm}}(t)$ are quadrature-modulated. The modulated SQAM signals, prior to the hardlimiter, can be expressed as:

$$z(t) = y_{\text{Im}}(t) \cos \omega_c t + y_{\text{Qm}}(t) \sin \omega_c t$$

$$= E_m(t) \cos[\omega_c t + \phi_m(t)]$$  \hspace{1cm} (3.1)

where $E_m(t)$ and $\phi_m(t)$ are the envelope and phase of the SQAM quadrature-modulated signals, that is,

$$E_m(t) = [y_{\text{Im}}^2(t) + y_{\text{Qm}}^2(t)]^{1/2}$$  \hspace{1cm} (3.2)

and
\[
\phi_m(t) = -\tan^{-1} \frac{y_{Qm}(t)}{y_{Im}(t)}
\] (3.3)

The SQAM demodulator is the same as a conventional OQPSK demodulator. In our computer simulation and experimental set-up, we used fourth-order Butterworth low pass filters (LPF) at the demodulator. (As we found that a 4th order Butterworth LPF yields the best P(e) performance in a single-channel environment.) For the nonlinearily amplified SQAM modem, an ideal hardlimiter presents a good first-order approximation to the saturated HPA. [1,10]. Note that the hardlimited SQAM signal \( z'(t) \) has a constant envelope.

As the SQAM modem operates with an offset (staggered) logic (i.e., a \( T_b = T_s/2 \) delay-line is inserted in the Q-channel), we obtain the following 16 combinations for the I and Q-channel output signals:

<table>
<thead>
<tr>
<th>I-Channel Output Signals</th>
<th>Q-Channel Output Signals</th>
</tr>
</thead>
<tbody>
<tr>
<td>( \pm y_1(t-nT_s) )</td>
<td>( \pm y_1(t - T_s/2 -nT_s) )</td>
</tr>
<tr>
<td>( \pm y_1(t-nT_s) )</td>
<td>( \pm y_2(t - T_s/2 -nT_s) )</td>
</tr>
<tr>
<td>( \pm y_2(t-nT_s) )</td>
<td>( \pm y_1(t - T_s/2 -nT_s) )</td>
</tr>
<tr>
<td>( \pm y_2(t-nT_s) )</td>
<td>( \pm y_2(t - T_s/2 -nT_s) )</td>
</tr>
</tbody>
</table>
where

\[ y_1(t - T_s/2 - nT_s) = -A + (1-A)\cos(2\pi t/T_s) \] (3.4)

\[ y_2(t - T_s/2 - nT_s) = -\sin(\pi t/T_s) \] (3.5)

The value of the amplitude parameter \( A \), in the I and Q-channel signals respectively, is selected to control the envelope fluctuations of the modulated signal. (However, the parameter \( A \) should be same for both the channels.) We note that the IQF-QQPSK signal described in [1,10,11] is a special case of the SQAM signals for \( A=1.0 \). An example of the SQAM signal encoding process, for the in-phase and quadrature baseband channels of the SQAM modulator, is illustrated in Fig.3.2.
Fig. 3.1 The block diagram of a SQAM modem.
(Both linearly amplified and hardlimited SQAM signal transmitters are illustrated.)
Fig.3.2 SQAM I and Q-channel signal encoding process.
3.2 **EYE PATTERN, SPACE DIAGRAM AND ENVELOPE FLUCTUATION**

Fig.3.3 and 3.4 show the computer simulated eye patterns of the demodulated SQAM baseband signals with amplitude parameter \( A = 0.7, 0.85 \) and 1.0, for linear and nonlinear (hardlimited) channels. The eye patterns of these SQAM signals, for a linearly amplified case, indicate that the vertical eye opening (VEO), for different values of \( A \), have full amplitudes at every sampling instant, thus the SQAM signals would suffer negligible intersymbol interference (ISI) and P(e) performance degradation. In a nonlinearly amplified system, the VEO for \( A < 1.0 \) is wider than that for \( A \geq 1.0 \), hence we can expect that the P(e) performance for \( A < 1.0 \) will be better than that for \( A \geq 1.0 \). (See also demodulated and filtered SQAM eye patterns in Fig.3.10 and 3.11.) Fig.3.5 shows the simulated signal space diagrams of the SQAM signals with amplitude parameters \( A = 0.7, 0.85 \) and 1.0. The maximum envelope fluctuations of the SQAM signals with amplitude parameters \( A = 0.7 \) to 1.0 are shown in Table 3.1 and Fig.3.6. Note that the maximum envelope fluctuation changes from 0.7 dB (for \( A = 0.7 \)) to 3.0 dB (for \( A = 1.0 \)), depending on the amplitude parameter.
Fig. 3.3 Simulated eye patterns of the SQAM baseband signals for linear channel. Note: The $A=1.0$ case is the same as the IJP-OQPSK [1,11] or the QORC [6].
Fig. 3.4 Simulated eye patterns of the SQAM baseband signals for hardlimited channel.
Fig. 3.5. Simulated signal space diagrams of SQAM signals.

(a) $A = 0.7$ (Max. envelope fluctuation = 0.7 dB)
(b) $A = 0.85$ (Max. envelope fluctuation = 1.7 dB)
(c) $A = 1.0$ (Max. envelope fluctuation = 3.0 dB)
Table 3.1. Computed maximum envelope fluctuations of the SQAM signals.

<table>
<thead>
<tr>
<th>A</th>
<th>Max. signal amplitude</th>
<th>Min. signal amplitude</th>
<th>Max. envelope fluctuation (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.7</td>
<td>1.075</td>
<td>0.987</td>
<td>0.7</td>
</tr>
<tr>
<td>0.8</td>
<td>1.165</td>
<td>1.0</td>
<td>1.3</td>
</tr>
<tr>
<td>0.9</td>
<td>1.280</td>
<td>1.0</td>
<td>2.1</td>
</tr>
<tr>
<td>1.0</td>
<td>1.414</td>
<td>1.0</td>
<td>3.0</td>
</tr>
</tbody>
</table>

Fig. 3.6. Envelope fluctuation of the SQAM signals.
3.3 PROBABILITY OF ERROR PERFORMANCE OF SQAM SYSTEMS

In this section, we investigate an optimum receiver for the SQAM, and derive an optimum $P(e)$ performance of the SQAM system. Then the $P(e)$ performance of the SQAM modem, in the presence of an AWGN, is evaluated by computer simulation, for linear and nonlinear (hardlimited) channels.

3.3.1 Optimum Receiver for SQAM

Referring to Fig.3.2, the signals taken at every sampling instant $nT_s$, over the interval of $[-T_s/2, 0]$, form a set of 8 waveforms and are shown in Fig.3.7. The remaining four not shown in Fig.3.7 are the negative symmetricals of the ones shown.

In general, the optimum correlation receiver maximizes the signal energy at its output, and minimizes the equivalent noise bandwidth. An optimum receiver for the SQAM must be equivalent to a coherent detector in which the receiver input signal, disturbed by interference and noise, is correlated with replicas of the transmitted signal.
Fig. 3.7 Composite waveforms of SQA signal, at center duration $T_s$. 
Fig. 3.8 shows the configuration of a basic correlation receiver, which consists of multipliers followed by integrators and sampling detectors.

![Block diagram of a correlation receiver for SQAM.]

The output of the correlation receiver, at the sampling instant $t = T_s$, is obtained as:

$$v(T_s) = \int_0^{T_s} z(t) f_j(t) \, dt$$

(3.6)

where $j = 1, 2, \ldots, 8$. 
3.3.2 Optimum P(e) Performance of the SQAM Modem

Fig.3.9 shows the equivalent baseband model of a SQAM system.

-\[ A \cdot \text{Gi}(K/2) \]

NRZ Input \[ \xrightarrow{\text{SQAM Encoder}} \] Correlation Receiver \[ \xrightarrow{\text{detected NRZ signal}} \]

Fig.3.9 SQAM linear baseband model.

As the NRZ input data is an equiprobable random signal, the probability of occurrence of any SQAM output signal is identical. At every sampling instant, the signal amplitude of any output signal is equal to 'K'. (In (2.9), (2.12), the value of K is normalized to 1.0.) Therefore, the probability of error in any case is given by:

\[ P_1 = P_2 = P_3 = P_4 = 0.5 \cdot \text{erfc}(K/\sigma \sqrt{2}) \]  \hspace{1cm} (3.7)

where \( \sigma \) is the RMS value of noise power. (See Table 3.2.)
Table 3.2. Probability of error for SQAM signals.

<table>
<thead>
<tr>
<th>NRZ input data</th>
<th>Encoded SQAM output signal</th>
<th>Probability of occurrence</th>
<th>Probability of error</th>
</tr>
</thead>
<tbody>
<tr>
<td>( a_{n+1} )</td>
<td>( y_1(t) )</td>
<td>1/4</td>
<td>( P_1 = 0.5 \text{erfc}(K/\sigma\sqrt{2}) )</td>
</tr>
<tr>
<td>( a_n )</td>
<td>( y_2(t) )</td>
<td>1/4</td>
<td>( P_2 = 0.5 \text{erfc}(K/\sigma\sqrt{2}) )</td>
</tr>
<tr>
<td>1</td>
<td>( y_3(t) )</td>
<td>1/4</td>
<td>( P_3 = 0.5 \text{erfc}(K/\sigma\sqrt{2}) )</td>
</tr>
<tr>
<td>-1</td>
<td>( y_4(t) )</td>
<td>1/4</td>
<td>( P_4 = 0.5 \text{erfc}(K/\sigma\sqrt{2}) )</td>
</tr>
</tbody>
</table>

Therefore, the overall probability of error is given by:

\[
P = \frac{1}{4} \left( P_1 + P_2 + P_3 + P_4 \right)
\]

\[= \frac{1}{2} \text{erfc}(K/\sigma\sqrt{2}) \quad \text{(3.8)}\]

An ideal (optimum) curve of the SQAM P(e) performance is shown in Fig. 3.14.
3.3.3 Simulation of the P(e) Performance of SQAM System

Fig. 3.10 shows the computer simulation model, used to evaluate the P(e) performance of the SQAM modem, for linear and nonlinear (hardlimited) channels. Additive white Gaussian noise (AWGN) is added to the modulated, hardlimited signal. The SQAM baseband signal processor is the only spectral shaping element in the transmitter. The receiver is modeled as a conventional coherent OQPSK demodulator having signal shaping 4th order Butterworth LPFs with $f_{3dB} = 1.1 f_N$ ($f_N = f_s / 2 = \text{Nyquist frequency}$). In our simulation, we frequently used $f_s = 120 \text{Mb/s}$ rate, thus in this case, the 3dB bandwidth ($f_{3dB}$) is equal to 33MHz.

Fig. 3.10 Simulation model of the SQAM modem.
Computer simulated eye patterns of the SQAM signals, after the postdetection LPF (or before the sampling detector), are shown in Figs. 3.11 and 3.12 for linear and hardlimited channels. These eye patterns show that the SQAM signal with amplitude parameter $A=0.8$ results in the best eye opening and the least intersymbol-interference (ISI) at the sampling instant, both in linear and hardlimited channels. For comparison purpose, eye patterns of the MSK signal, after the postdetection LPF, are also shown in Figs. 3.13 and 3.14 for linear and hardlimited channels. Note that the SQAM signal with $A=0.8$ has better eye patterns than MSK signal in linear and nonlinear channels.
Fig. 3.11 Eye patterns of the SQAM signals after postdetection LPF in a linear channel (I or Q-channel).
Fig. 3.11 Eye patterns of the SQAM signals after postdetection LPF in a linear channel.
SQAM EYE DIAGRAM

(a) \( A = 0.7 \)

(b) \( A = 0.8 \)

(c) \( A = 0.9 \)

Fig. 3.12 Eye patterns of the SQAM signals after postdetection LFF in a hardlimited channel.
Fig. 3.12 Eye patterns of the 8QAM signals after postdetection LPF in a hardlimited channel.
(a) With receive filter only.

(b) With transmit and receive filters.

Fig. 3.13  Eye patterns of the MSK signal after postdetection LPF in a linear channel.

(a) With receive filter only.

(b) With transmit and receive filters.

Fig. 3.14  Eye patterns of the MSK signal after postdetection LPF in a hardlimited channel.
Figs. 3.15(a) and (b) show the simulated P(e) performance results of the SQAM modem for linear and nonlinear channels. For the illustrative case, a data bit rate of 120 Mb/s (60 Mbaud) was assumed. We obtained the best performance for A=0.8 in linearly and nonlinearly amplified channels. Figs. 3.16(a) and (b) illustrate the P(e) performance degradation, that is, the increased $E_b/N_o$ requirement of this SQAM system to maintain $P(e)=10^{-4}$. The performance of SQAM is degraded by 0.3 dB (for A=0.8) in a linear channel, and 0.5 dB (for A=0.8) in a saturated (hard-limited) channel. To facilitate the comparison with the performance of MSK and TFM, the increased $E_b/N_o$ requirements of these systems are also shown in Figs. 3.16(a) and (b). [1]. The filtering strategies, which have been used for the generation of these results, are summarized in Table 3.3. Note that SQAM (A=0.8) outperforms MSK by 0.5dB in a linear channel, and by 0.8dB in a hardlimited channel. SQAM (A=0.8) also outperforms TFM by 0.7dB in a linear channel. Further, the generation of SQAM signal is much simpler than that of TFM signal.
Note:
1. data symbol rate:
   $f_s = 60$ kHz baud.
2. receiver filter:
   4th order Butterworth
   LPF ($f_{3dB} = 33$ kHz)

Fig. 3.15(a). Simulated $P_e$ performance of SQAM modem in a linear channel.
**SQAM BER CURVE**

**Note:**
1. data symbol rate:
   \[ f_s = 60\text{kHz} \text{baud} \]
2. receiver filter:
   4th order Butterworth
   LPF \( f_{3dB} = 33\text{kHz} \)

---

![Graph](image.png)

- \( A = 1.1 \)
- \( A = 1.0 \) (IJP-OQPSK, SQORC)
- \( A = 0.9 \)
- \( A = 0.7 \)
- \( A = 0.8 \)

**Fig. 3.15(b).** Simulated \( P_e \) performance of SQAM modem in a nonlinear/hardlimited channel.
Fig. 3.16. Required increase in $E_b/N_0$ (performance degradation) to maintain $P_e = 10^{-4}$ for SQAM, MSK, and TFM.
Table 3.3 Filtering strategies for different modems.

<table>
<thead>
<tr>
<th>Modem.</th>
<th>Transmit filter</th>
<th>Receive filter</th>
</tr>
</thead>
<tbody>
<tr>
<td>MSK.</td>
<td>4th order Butterworth LPF ($f_{3dB}=33$ MHz)</td>
<td>4th order Butterworth LPF ($f_{3dB}=33$ MHz)</td>
</tr>
<tr>
<td>TFM[3]</td>
<td>Low pass filter $B(w)$</td>
<td>Optimum LPF for TFM $A(w)$</td>
</tr>
<tr>
<td>IJF-QOQPSK [1]</td>
<td>Nil</td>
<td>Aperture equalized full raised-cosine filter ($\alpha$)</td>
</tr>
<tr>
<td>SQAM</td>
<td>Nil</td>
<td>4th order Butterworth LPF ($f_{3dB}=33$ MHz)</td>
</tr>
</tbody>
</table>

### 3.4 Experimental Results

The schematics of experimental set-up of SQAM modem is shown in Fig.3.17. In this set-up, a pseudo random binary sequence (PRBS) generator provides a NRZ data of 128 kb/s, and a random noise generator (20 Hz - 20 MHz) provides an AWGN. Phase equalized 4th order Butterworth LPFs ($f_{3dB}=35.2$ kHz) are used as postdetection LPFs in the receiver. Fig.3.18 shows the measured I and Q-channel eye patterns of the SQAM signal for $A=0.8$ in a linear channel. Fig.3.19 (a) and (b) show a baseband I( or Q)-channel SQAM signal ($A=0.8$)
and a corresponding modulated (carrier frequency $f_c = 1$ MHz) $I$ (or $Q$)-channel SQAM signal, respectively. Fig. 3.20 (a) and (b) show measured signal space diagrams of SQAM signal ($A=0.8$) in a linear and in a hardlimited channel, respectively. Note that the hardlimited SQAM signal has a constant envelope. Fig. 3.21 shows measured power spectra of SQAM signal for $A=1.0$ in a linear and in a hardlimited channel, and also for $A=0.8$ in a hardlimited channel. Note that for $A=0.8$, the spectral spreading is lower than that for $A=1.0$ in a hardlimited channel.

Fig. 3.17 Experimental set-up of SQAM modem.
Fig. 3.18. Measured Eye-patterns of SQAM signal \((A=0.8)\).

Fig. 3.19. (a) I (or Q)- channel SQAM baseband signal.
(b) I (or Q)- channel modulated \((f_c=15\,\text{MHz})\)
SQAM signal.
Fig. 3.20.
Measured SQAM signal
(A=0.8) space diagram.

(a). In a linear channel.

(b). In a hardlimited channel.

Fig. 3.21.
Measured power spectra
of SQAM signals.

(b). Middle: A=0.8, hardlimited.
(c). Lower : A=1.0, linear.
Chapter IV
PERFORMANCE OF SQAM MODEM IN NONLINEAR MULTICHANNEL

4.1 ANALYSIS OF PROBABILITY OF ERROR IN MULTICHANNEL

In this section, we derive expressions for the $P(e)$ performance which include the effects of ISI, ACI [27] and/or CCI in the linearly amplified multichannel system. Based on these expressions, we also find the worst case (high bound) error probability in an ACI or CCI environment.

4.1.1 Probability of Error in Adjacent Channel Interference

The probability of error, for linearly amplified multichannel systems in the presence of AWGN, ISI and ACI, is analyzed. The block diagram of a multichannel modem is shown in Fig.4.1, where two adjacent channels (i.e., one higher frequency adjacent channel and another lower frequency adjacent channel) interfere with the main (desired) channel.

The transmitted signal is given by:

$$z_n(t) = \text{Re} \left[ v_n(t) \exp(j2\pi f_c t) \right]$$

(4.1)
where \( n=[-1,0,1] \), and 'n=0' represents the term due to the desired signal, and 'n=±1' represent terms due to the interfering signals, and \( f_0 \) is the carrier frequency of the main channel.

The corresponding complex envelope \( v_n(t) \) is given by [27]:

\[
v_n(t) = K \rho_n \exp\left[\frac{j\pi}{2} \sum_{i} a_{n,i} s(\tau-\tau_n-iT_s) \, d\tau + n\Delta F_2 t + \theta_n\right]
\]

(4.2)

where

- \( K \) = amplitude of desired signal.
- \( \rho_n \) = relative signal amplitude of adjacent channel.
  
  \( (\rho_0=1 \text{ and } \text{F.D(fade-depth)} = 20\log(\rho_{\text{adj}}/\rho_0).) \)
- \( s(t) \) = impulse response of a baseband signal encoder.
  
  (For SQAM signal encoder, see (2.3) and (2.4).)
- \( \tau_n \) = symbol timing misalignment of adjacent channels
  
  \( (\tau_0=0 \text{ sec.}) \)
- \( \Delta F \) = channel spacing. (See Fig.4.3.).
- \( \theta_n \) = carrier phase misalignment of adjacent channels
  
  \( (\theta_0=0 \text{ rad.}) \)
- \( a_{n,i} = \pm 1 \) with equal probability. ( 'i' represents terms due to intersymbol-interference. )
- \( T_0 \) = symbol duration,
Fig. 4.1 Block diagram of a modem in a multichannel interference environment.

Note:
- $z_0(t)$: Main channel.
- $z_{\pm 1}(t)$: Adjacent channels.
- $h_T(t) \triangleq h_0(t)$
The demodulated signal, at the output of the receive filter, is given by:

\[ r(t) = x_n(t) + n(t) \]  \hspace{1cm} (4.3) \]

\[ x_n(t) = x_0(t) + x_1(t) + x_{-1}(t) \]  \hspace{1cm} (4.4) \]

\[ x_n(t) = \text{Re} \{ v_n * h_n(t) * h_R(t) \} \]  \hspace{1cm} (4.5) \]

where \(*\) denotes convolution, \(h_n(t)\) and \(h_R(t)\) represent the impulse response of transmit and receive filters, respectively. (Let the main-channel transmit filter have an impulse response \(h_{\Delta}(t) = h_0(t)\) with \(n=0.\))

\(n(t)\) is a zero-mean Gaussian noise with variance

\[ \sigma_n^2 = N_0 / 2 \int_{-\infty}^{\infty} |H_R(f)|^2 df \]  \hspace{1cm} (4.6) \]

where \(N_0 / 2\) is the double-sided PSD of AWGN, and

\[ H_R(f) = F[h_R(t)] \]

Assuming \(i=0\) and desired signal \(a_0, \gamma = 1\), then the probability of error, at the sampling instant \(t=t_0\), is expressed as:

\[ P(e) = Q \left[ \left( Kf(t_0) + \text{ISI} + \text{ACI} \right) / \sigma_n \right] \]  \hspace{1cm} (4.7) \]

where

\[ f(t) = s(t) * h_{\Delta}(t) * h_R(t) \]  \hspace{1cm} (4.8) \]

\[ \text{ISI} = \sum_{i \neq 0} a_i, i f(t - iT_S) \]  \hspace{1cm} (4.9) \]
\[ ACI = x_1(t_0) + x_{-1}(t_0) \]  
\[ (4.10) \]

\[ Q(x) = \frac{1}{\sqrt{2\pi}} \int_{-\infty}^{\infty} \exp(-y^2/2)dy \]  
\[ (4.11) \]

and the bar \( \bar{\ } \) denotes the average over all random symbols contained in ISI and ACI.

In (4.9), let us assume only two intersymbol-interference components are significant, then

\[ ISI = K [ a_{0,1} f(t_{-1}) + a_{0,-1} f(t_1) ] \]  
\[ (4.12) \]

and the worst ISI occurs when \( a_{0,1} = a_{0,-1} = -1 \), that is,

\[ \text{worst ISI} = -K [ f(t_{-1}) + f(t_1) ] \]  
\[ (4.13) \]

When adjacent channel signals have opposite polarities to that of the desired signal, that is \( a_1 = 0, a_{-1} = -1 \), the worst ACI appears to be

\[ \text{worst ACI} = - \left[ |x_1(t_0)| + |x_{-1}(t_0)| \right] \]  
\[ (4.14) \]

The worst case (high bound) error probability can be calculated from (4.7), (4.13) and (4.14).
4.1.2 Probability of Error in Co-Channel Interference

The probability of error, in the presence of AWGN, ISI and CCI, is analyzed in this section. Using the same parameters as in the case of ACI in section 4.1.1, the transmitted signal is given by:

\[ z_n(t) = \text{Re} \left[ v_n(t) \exp(j2\pi f_0 t) \right] \]  \hspace{1cm} (4.1)

where complex envelope \( v_n(t) \) is given by:

\[ v_n(t) = K \rho_n \exp \left[ j \frac{t}{2} \right] \sum_{n=1}^{t} a_{n,i} s(t - \tau_n - iT_s) \, d\tau + \theta_n \]  \hspace{1cm} (4.2)

(\( \Delta F \) in (4.2) is set to zero in CCI environment.)

The demodulated output signal is:

\[ r(t) = x_n(t) + n(t) \]  \hspace{1cm} (4.3)

\[ x_n(t) = x_0(t) + x_1(t) \]  \hspace{1cm} (4.4)

\[ x_n(t) = \text{Re} \left[ v_n(t) * h_n(t) * h_R(t) \right] \]  \hspace{1cm} (4.5)

where \( n(t) \) is zero-mean Gaussian noise with variance

\[ \sigma_n^2 = \frac{N_0}{2} \int_{-\infty}^{\infty} |H_R(f)|^2 \, df \]  \hspace{1cm} (4.6)
The probability of error is expressed as:

\[ P(e) = Q \left[ \left( Kf(t_0) + \text{ISI} + \text{CCI} \right) / \sigma_n \right] \]  \hspace{1cm} (4.7)'

where

\[ f(t) = s(t) * h_T(t) * h_R(t) \]  \hspace{1cm} (4.8)

\[ \text{ISI} = \sum_{n \neq 0} \rho_n a_n, i f(t_0 - it_s) \]  \hspace{1cm} (4.9)'

\[ \text{CCI} = \sum_{n \neq 0} x_n(t_0) \]  \hspace{1cm} (4.10)'

\[ Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^{\infty} \exp(-y^2/2) \, dy \]  \hspace{1cm} (4.11)

Dominant ISI = \[ K \sum_n \rho_n \left[ a_{n,1} f(t_{-1}) + a_{n,-1} f(t_1) \right] \]  \hspace{1cm} (4.12)'

The worst ISI occurs with \( a_{n,1} = a_{n,-1} = -1 \), and the worst CCI occurs with \( a_{n,0} = -1 \) when a desired signal \( a_{0,0} = 1 \) is assumed.

Thus,

\[ \text{worst ISI} = -K \sum_n \rho_n \left[ f(t_{-1}) + f(t_1) \right] \]  \hspace{1cm} (4.13)'

\[ \text{worst CCI} = -\sum_{n \neq 0} |x_n(t_0)| \]  \hspace{1cm} (4.14)'
where the integer 'n' will be determined depending on different applications (folds) of the frequency reuse system. The worst case (high bound) error probability can be calculated from (4.7), (4.13) and (4.14).

However, it is quite difficult to analytically evaluate the average P(e) performance of quadrature modems, in the presence of AWGN, ISI and ACI [14,16,27] or CCI [13,15,17], especially for the nonlinearly amplified multichannel system. Therefore, in our next step, a computer simulation is employed to analyze the performance of the SQAM modem in a nonlinear multichannel environment.

4.2 SIMULATION ON P(E) PERFORMANCE OF MULTICHANNEL SQAM

In this section, we describe a computer simulation approach which is employed to evaluate the performance of the SQAM modem in a nonlinear channel with ACI and CCI. We use a well established computer simulation software package called COMSIM. COMSIM is a software package (primarily based on time domain operations) used for the computer aided design and analysis of digital communication systems, and was developed at Digital Communication Group of
A simulation model of a SQAM modem with two adjacent channels is shown in Fig. 4.2, while its frequency allocation is shown in Fig. 4.3.

In section 4.2.1, we discuss the assumptions made for this simulation. The performance of the multichannel SQAM modem, for different values of $A$, is evaluated in section 4.2.2, under different channel conditions. In section 4.2.3...4.2.6, effects of the receive filter bandwidth, channel spacing, fade depth on the desired signal, and effects of co-channel interference (CCI) on the performance of the SQAM multichannel systems are investigated and compared to those of other modems.
Fig. 4.2 A simulation model of a multichannel system.

Fig. 4.3 Frequency allocation for multi-channel system.

Note:

- $f_0$: main-ch carrier freq.
- $f_{\pm l}$: adj-ch carrier freq.

($f_{\pm l} = f_0 \pm \Delta f$)
4.2.1 Assumptions Used in the Simulations

(1). Two interfering adjacent channels are equally spaced. In Fig.4.3, the main channel has a carrier frequency \( f_0 \), and the adjacent channels have carrier frequencies \( f_{\pm 1} = f_0 \pm \Delta F \), where \( \Delta F \) is the channel frequency separation (or channel spacing).

(2). The interfering signals have the same form as the desired signal.

(3). The carrier phase and symbol timings of the interfering signals are randomized over the interval \((0, 2\pi)\) and \((-T_s/2, T_s/2)\) respectively to avoid the coherence and synchronization between the interfering and main channels.

(4). The earth station HPAs, operating in saturation mode, are approximated by ideal hardlimiters. (An ideal hardlimiter presents a good first-order approximation of a saturated amplifier. [1,12,13])

(5). Only the desired signal is attenuated due to uplink flat fading.

(6). An illustrative input data bit rate, \( f_b = 120 \text{ Mb/s} \) (or data symbol rate \( f_s = 60 \text{Mbaud} \)) is used. This is a typical bit rate for a number of high-speed systems.

(7). In the SQAM modem, only receive filters are used.
4.2.2 Effects of the Amplitude Parameter $A$ in SQAM

In this simulation, phase equalized 5th order Butterworth LPFs ($f_{3dB} = 30$ MHz) are used in the receiver. The $P(e)$ performance results of the hardlimited multichannel SQAM modem for different values of $A$ are shown in Figs.4.4(a) . . . (d). In Figs.4.4(a) and (b), the adjacent interfering channels have the equal power as the main channel, while in Figs.4.4 (c) and (d), the main channel suffers a fading by an amount of 6 dB. Note that with low levels of fading, the SQAM modem for $A = 0.8$ shows the best performance, while with high levels of fading, the SQAM modem for $A = 0.9$ has a better performance. Hereafter, the SQAM signal for $A = 0.85$ is considered for our further evaluation on the $P(e)$ performance of the multichannel SQAM systems.
SQAM BER CURVE

Note:
1. data symbol rate:
   \( f_s = 60 \text{Mbaud} \).
2. receive filter:
   5th order Butterworth
   LPF(\( f_{\text{dB}} = 30 \text{kHz} \)).
3. Equal power channels.
   (F.D = 0 dB)
4. Channel spacing:
   \( \Delta F = 90 \text{MHz} \).

Fig. 4.4(a) \( P_e \) performance of SQAM modem in a hardlimited
multi-channel system (vs. 'A').
SQAM BER CURVE

Note:
1. data symbol rate:
   \( f_s = 60 \text{ baud} \).
2. receive filter:
   5th order Butterworth
   LPF \( f_{3dB} = 30 \text{ MHz} \).
3. Equal power channels:
   \( P.D = 0 \text{ dB} \).
4. Channel spacing:
   \( \Delta f = 100 \text{ MHz} \).

Fig. 4.4(b) \( P_e \) performance of SQAM modem in a hardlimited
multi-channel system (vs. 'A').
Fig. 4.4(c)  $P_e$ performance of SQAM modem in a hardlimited multi-channel system (vs. 'A').
Fig. 4.4(d) P_e performance of SQAM modem in a hardlimited multi-channel system (vs. 'A').
4.2.3 Effects of the Receive Filter Bandwidth.

In this section, we investigate effects of a receive filter bandwidth on the performance of the SQAM modem in a multichannel environment, and try to find out an optimal filter bandwidth \( B \) and symbol duration \( T_s \) product for the 5th order Butterworth LPF. In these simulations, phase-equalized 5th order Butterworth LRFs are used. The \( P(e) \) performance results of the SQAM \( (A=0.85) \) modem for different bandwidth \( \left( f_{3dB} \right) \) of the receive filter are shown in Figs.4.5(a) and (b), with channel spacings \( \Delta F=90 \) and 100 MHz respectively. The \( P(e) \) performance degradation (that is, the increased \( E_b/N_0 \) requirement of this SQAM system to maintain \( P(e)=10^{-4} \), as a function of \( BT_s \) product, is shown in Fig.4.5(c), where \( B \) (or \( f_{3dB} \)) is the 3dB bandwidth of the LPF, and \( T_s \) \( (=1/f_s) \) is the data symbol duration. It must be noted that the filter bandwidth cannot be too large because of the thermal noise and ACI effects, and cannot be too small because of ISI effect. A trade-off exists between ISI and ACI losses, hence an optimal receive bandwidth is to be found. Using the simulation, we found that the near optimal \( BT_s \) product is about 0.5 , and the major source of the performance impairment is the ACI when \( BT_s > 0.5 \), and the ISI when \( BT_s < 0.5 \). It is also noticed that the LRF optimal bandwidth increases with increasing values of channel spacing \( (\Delta F) \), and decreases with increasing values of fade depth of the desired signal.
**SQAM BER CURVE**

**Note:**
1. $f_s = 60$ Mbndd.
2. $A = 0.85$
3. Fading depth:
   - $F.D = 6$ dB
4. Channel spacing:
   - $\Delta F = 90$ MHz.

---

**Fig. 4.5(a)** $P_e$ performance of SQAM modem in a hardlimited multi-channel system (vs. $f_{3dB}$).
SQAM BER CURVE.

Note:
1. $f_s = 60$ Mbaud.
2. $A = 0.85$
3. Fading depth:
   - F.D = 6 dB
4. Channel spacing:
   - $\Delta f = 100$ MHz.

Fig. 4.5(b) $P_e$ performance of SQAM modem in a hardlimited multi-channel system (vs. $f_{3dB}$).

- 79 -
Fig. 4.5(c) $E_b/N_0$ degradation vs. $BT_s$ products of receive filter (5th order Butterworth LPF) in a hard-limited multi-channel SQAM system.
4.2.4 Effects of Channel Frequency Separation (Spacing)

In this simulation, we assume that the interfering signals have equal power compared to the desired signal, and the 3dB bandwidth of the receive lowpass filter is 30 MHz (Nyquist minimum bandwidth). The P(e) performance results of the SQAM modem for different channel spacing are shown in Fig. 4.6(a). The P(e) performance degradation, as a function of a normalized channel spacing (or spectral efficiency), is shown in Fig. 4.6(b), where the normalized channel spacing is expressed in ΔF/f₀ (i.e. channel spacing /bit rate). The performance of the SQAM modem, for a hardlimited multichannel system, is also compared to those of OQPSK, MSK and IIF-OQPSK modems.[11,16,27]. The filtering strategies, which have been used for the generation of these results, are summarized in Table 4.1. Note that for ΔF/f₀ > 80/120, the SQAM modem outperforms all the other schemes studied here. It is also noticed that the performance of MSK modem without transmit filters is much worse than that with transmit filters. This is so because the bandlimiting filters in the MSK transmitter confine the bandwidth of the transmitted signals, and reduce the ACI. Note that for the large channel spacings, that is, when ΔF/f₀ > 110/120, the performance is getting close to the performance in a single channel environment, and the ACI is predominantly caused by the spectral spreading of the modulated adjacent carriers.
Fig. 4.6(a) $P_e$ performance of SQAM modem in a hardlimited multi-channel system (vs. $\Delta F$).

**Note:**
1. $f_s = 60$ Mbaud.
2. $A = 0.85$
3. $f_{3dB} = 30$ MHz
4. Equal power channels.
   ( $\Delta D = 0$ dB )
Fig. 4. (b) $E_b/N_0$ degradation vs. channel spacing for hardlimited multi-channel system with two equal-power ACIs. (*Compared to $E_b/N_0 = 8.4 \text{dB at } P_e = 1 \times 10^{-4}$)

- MSK (Rx filter only)
- MSK (Tx & Rx filters)
- IJP-OQPSK
- OQPSK
- SQAM ($A = 0.85$)

(normalized ch.-spacing)

(spectral efficiency)

$(b/c/\text{Hz}) \rightarrow 1.71 \quad 1.5 \quad 1.33 \quad 1.2 \quad 1.09 \quad 1.0$
Table 4.1 Filtering Strategies for Different Modems.

<table>
<thead>
<tr>
<th>Modem</th>
<th>Transmit Filter</th>
<th>Receive Filter</th>
</tr>
</thead>
<tbody>
<tr>
<td>QPSK, OQPSK</td>
<td>aperture equalizer ((x/sin x) + ) square-root raised-cosine filter</td>
<td>square-root raised-cosine filter ((\alpha=0.4))</td>
</tr>
<tr>
<td>MSK</td>
<td>5th order Butterworth LPF ((f_{3dB}=f_s/2))</td>
<td>5th order Butterworth LPF ((f_{3dB}=f_s/2))</td>
</tr>
<tr>
<td>SQAM (IJF-OQPSK)</td>
<td>Nil</td>
<td>5th order Butterworth LPF ((f_{3dB}=f_s/2))</td>
</tr>
</tbody>
</table>

4.2.5 Effects of Fade Depth on the Desired Signal

In our study, we assume that the desired signal (main channel) suffers fading due to the rain in the uplink, and the fade depth is varied from 0 dB to 14 dB with different channel spacings. The P(e) performance results of the SQAM modem for different fade depths are shown in Figs. 4.7(a)
and (b), for channel spacings $\Delta f = 90$ and $100$ MHz. The $P(e)$ performance degradation as a function of the fade depth are shown in Figs. 4.8(a), (b), and the performance of SQAM modem is compared to those of OQPSK, MSK and IIF-OQPSK modems. Note that SQAM modem followed by IIF-OQPSK modem outperforms the other modems for every fade depth. This can be explained from the fact that the out-of-band energy of SQAM signal, for linearly and nonlinearly amplified systems, is much lower than those of other signals. [Refer to Figs. 2.10(a) and (b).] Thus, the SQAM signal causes less ACI compared to other signals. It is also noticed that in a narrow channel spacing or at low levels of fade depth, the OQPSK modem outperforms the MSK modem. This is so because in MSK the sidelobe roll-off is faster, but its mainlobe bandwidth is 50% wider than that of OQPSK. Note that in the ACI environment, the performance of MSK modem without transmit filter is much worse than that with transmit filter, especially at high levels of fade depths or with narrow channel spacing. Simulated eye patterns of the demodulated and filtered MSK, IIF-OQPSK and SQAM signals are shown in Figs. 4.9(a), (b) and (c) respectively.
SQAM BER CURVE

Note:
1. $f_s = 60$ kHz.
2. $A = 0.85$
3. $f_{3dB} = 30$ MHz
4. Channel spacing:
   $\Delta F = 100$ MHz.
5. F.D : Fading depth.

Fig. 4.7(a) $P_e$ performance of SQAM modem in a hardlimited multi-channel system (vs. F.D).
Fig. 4.7(b) $P_e$ performance of SQAM modem in a hardlimited multi-channel system (vs. $F_d$).

F.D = 12 dB

Note:
1. $f_s = 60$ Mbaud.
2. $A = 0.85$
3. $f_{3dB} = 30$ MHz
4. Channel spacing:
   $\Delta F = 90$ MHz
Fig. 4.8(a) $E_b/N_0$ degradation vs. fade depth of the desired channel for hardlimited multichannel system. (Channel spacing: $\Delta F = 100$ MHz)
Fig. 4.8(b) $E_b/N_0$ degradation vs. fade depth of the desired channel for hardlimited multichannel system.

( Channel spacing : $\Delta f = 90$ MHz)
Fig. 4.9 Eye patterns of demodulated and filtered MSK, IJP-QPSK, and SIAM signals in a hard-limited multi-channel system \((\Delta F = 100\,\text{kHz}, \, F.D = 6\,\text{dB})\).
4.2.6 Effects of CCI on P(e) Performance of SQAM Modems

In a frequency-reuse communication system, co-channel interference (CCI) is one of the major sources of the performance impairment. [13, 15, 17]. A simulation model of a co-channel system is shown in Fig. 4.10, while its frequency allocation is shown in Fig. 4.11. In this simulation, an interfering channel, centered at the same carrier frequency as the main channel (i.e., ΔF = 0), is introduced, and 5th order phase-equalized Butterworth LPFs are used in the SQAM receiver. The P(e) performance results of the SQAM modem at different levels of main-channel carrier-to-cochannel interference (C/I) ratio, are shown in Fig. 4.12. The P(e) performance degradation of the SQAM modem, as a function of C/I ratio for different BTs, product, is shown in Fig. 4.13. It is noticed that ISI, in addition to CCI, is also a major source of the performance impairment. The P(e) performance degradation of the SQAM modem as a function of C/I ratio is compared to those of OQPSK, MSK, and IIF- OQPSK modems in Fig. 4.14. Note that SQAM outperforms all the other modems studied here. It is also noticed that MSK modem without transmit filter outperforms the one with a transmit filter. This is so because as the bandlimiting of the filter becomes severe, the ISI increases, thus, the performance is degraded in a CCI environment. [Refer to Fig. 3.13 and eq.(4.12)].
Fig. 4.10 A simulation model of co-channel system.

Fig. 4.11 Frequency allocation for co-channel system.
SQAM BER CURVE

Note:
1. $f_s = 60$ MHz.
2. $A = 0.85$
3. receive filter:
   5th order Butterworth
   LPF ($f_{3dB} = 33$ MHz).

Fig. 4.12 $P_e$ performance of SQAM modem in a hardlimited co-channel system (vs. $C/I$).
Fig. 4.13 $E_b/N_0$ degradation vs. C/I ratio for different $f_{3dB}$ of receive filter in a hardlimited co-channel SQA system.
Fig. 4.14 $E_b/N_0$ degradation versus C/I in a hardlimited CCI and AWGN environment.

(* compared to $E_b/N_0 = 8.4$ dB at $P_0 = 1 \times 10^{-4}$)

**Note:**
1. SQAM receive filter:
   5th order Butterworth
   LPF ($f_{3dB} = 30$ MHz).

- SQAM ($A = 0.85$)
- DSK (Tx & Rx filter)
- IJPSK-OQPSK
- DSK (Rx filter only)
Chapter V
RECOMMENDED FURTHER RESEARCH

For a good power efficiency, a constant envelope modulation technique may be preferable, and this leads to the use of frequency modulations. In FM schemes, to narrow the power spectrum, a pulse shaping filter (i.e. premodulation filter) prior to the frequency modulator is required. In TFM (or CORPSK), a premodulation filter, composed of a correlative encoder and an aperture-equalized raised-cosine filter, is used to achieve smooth phase transitions, and thus good spectral properties. However, the implementation of the filter is complicated. The impulse response of the premodulation filter in TFM is shown in Fig. 5.1, and the block diagram of TFM transmitter is shown in Fig. 5.2.

In SQAM, the phase is a smooth function of time, and correlated between the consecutive symbols. (Refer to section 2.2.) The impulse response of SQAM signal encoder $s(t)$ is shown in Fig. 2.1.(b), where its generation is quite simple as shown in Fig. 2.3.
Fig. 5.1 Impulse response of TFM premodulation filter.

![Impulse response graph]

Fig. 5.2 Block diagram of TFM transmitter.

In SQAM, the phase behavior is very similar to TFM for \( \lambda = 1.0 \), and similar to CCPSK [20] for \( \lambda < 1.0 \). Hence, if we use the SQAM signal encoder as a premodulation filter in digital FM scheme, the resulting frequency modulated signal will retain a constant envelope and good spectral properties. Fig. 5.3 shows the suggested block diagram.
Fig. 5.3 Digital FM transmitter combined with SQAM encoder.

In order to improve further the spectral properties of the SQAM signals, 'duo-binary SQAM technique', where SQAM baseband signal processing is combined with the duo-binary (1+D; class I) technique [21], is suggested. A duo-binary SQAM encoder is composed of a SQAM encoder and a duo-binary encoder. The block diagram of a duo-binary SQAM encoder is shown in Fig. 5.4. The power spectral density function of the duo-binary SQAM signal is derived and plotted in Appendix B. This result shows that duo-binary SQAM signals provide a lower out-of-band energy and a faster spectral roll-off, compared to conventional SQAM signals.
Fig. 5.4 Block diagram of Duo-binary SQAM Encoder.

A further study of modulated duo-binary SQAM signals, being a similar approach to the previous chapters, could lead to additional valuable results.

We also suggest an extension of the pulse overlapping interval (i.e. multi-symbol interval) in SQAM systems, or a partial response[25] SQAM for a better power spectrum.
Chapter VI
CONCLUSION

A spectral and power efficient modulation technique—SQAM has been introduced. In SQAM, by superposing two baseband waveshapes which have different symbol-intervals and weighting factors, a good power and spectral efficiency was achieved.

In the first part of this thesis, a power spectral density function of the SQAM signal was derived, and the spectral characteristics and the P(e) performance in a single-channel environment were compared with those of other schemes. Experimental and simulation results showed that SQAM signals have significant spectral advantages over QPSK, OQPSK and MSK signals and better P(e) performance than other constant envelope modulation schemes, such as MSK and TFM. In SQAM, the envelope fluctuation and power spectrum can be controlled to minimize the P(e) performance degradation in a particular operating environment, by adjusting the amplitude parameter A.

In the second part of this thesis, the performance of a SQAM modem, in a nonlinearly amplified multichannel environment, in the presence of AWGN, ISI, ACI and/or CCI,
was investigated and compared to those of OQPSK, MSK and IIF-OQPSK modems. The effects of ACI and CCI on the performance of SQAM modem were investigated under different channel conditions, such as different channel spacing, filter bandwidth symbol duration product (\(BT_s\)) and flat fade depth. It has been found that our SQAM modem outperforms OQPSK, MSK and IIF-OQPSK modems for certain system environments, and that ACI is one of the major sources of the performance impairment of multichannel systems, especially under flat fadings. It was also found that the proper selection of modulation techniques, type of modem filters (transmit and receive) and their \(BT_s\) values can minimize the degradation of \(P(e)\) performance caused by ACI or CCI.

These desirable performance characteristics, combined with the simple hardware implementation of SQAM modems, may lead to numerous satellite and terrestrial radio systems applications.
Appendix A

DERIVATION OF THE PSD OF SQAM SIGNAL

*(Equation (2.22) in Chapter 2.)*

The impulse method [19] is applied to derive the frequency spectrum of the SQAM baseband signals. This method uses the differentiation and the time domain shift properties of the Fourier transform. In (2.3) and (2.4), a SQAM double-interval pulse is defined as:

\[ s(t) = g(t) + d(t) \quad (A.1) \]

First, consider the double-interval raised-cosine pulse \( g(t) \) of the SQAM signal:

\[ g(t) = \begin{cases} 1/2 (1 + \cos \pi t/T_S) & |t| \leq T_S \\ 0 & |t| > T_S \end{cases} \quad (A.2) \]

(See Fig.A.1.)

Now differentiate the pulse \( g(t) \) a sufficient number of times to produce impulse functions. The derivatives of \( g(t) \) with respect to \( 't' \) are given by:
Fig. A.1. A double-interval raised-cosine pulse and its first three derivatives.
\[
\frac{dg(t)}{dt} = \begin{cases} \frac{-\pi}{2T_s} \cdot \sin \frac{\pi t}{T_s} & |t| \leq T_s \\ 0 & |t| > T_s \end{cases} \quad (A.3)
\]

\[
\frac{d^2g(t)}{dt^2} = \begin{cases} -\frac{\pi^2}{2T_s} \cdot \cos \frac{\pi t}{T_s} & |t| \leq T_s \\ 0 & |t| > T_s \end{cases} \quad (A.4)
\]

\[
\frac{d^3g(t)}{dt^3} = \begin{cases} \frac{\pi^3}{2T_s} \cdot \sin \frac{\pi t}{T_s} - \frac{\pi^2}{2T_s^2} [\delta(t-T_s) - \delta(t+T_s)] & |t| \leq T_s \\ 0 & |t| > T_s \end{cases} \quad (A.5)
\]

From eq. (3) and (5)

\[
\frac{d^3g(t)}{dt^3} = \begin{cases} -\frac{\pi^2}{2T_s} \cdot \frac{dg(t)}{dt} + \frac{\pi^2}{2T_s^2} \delta(t+T_s) - \delta(t-T_s) & |t| \leq T_s \\ 0 & |t| > T_s \end{cases} \quad (A.6)
\]

From differential and time shift properties of the Fourier transforms (i.e., \(F\{\frac{d^n}{dt^n}G(t)\} = (j2\pi f)^nG(f)\) and \(F\{a(t-t_0)\} = A(f)e^{-j2\pi ft_0}\)), we can rewrite (6) as:

\[
(j2\pi f)^3 G(f) = \frac{-\pi^2}{2T_s} (j2\pi f)G(f) + \frac{\pi^2}{2T_s^2} (e^{j2\pi fT_s} - e^{-j2\pi fT_s})
\]

\[
= \frac{-\pi^2}{2T_s} (j2\pi f)G(f) + \frac{\pi^2}{2T_s^2} (2\pi \sin 2\pi fT_s) \quad (A.7)
\]
Solving for \( G(f) \), we find the frequency spectrum of the double-interval raised-cosine pulse to be:

\[
G(f) = \frac{\frac{2}{T_s^2}}{\frac{2}{T_s^2} - (2\pi f)^2} \cdot \frac{\sin 2\pi f T_s}{2\pi f T_s}
\]  

\[
= \frac{T_s}{1 - 4\pi^2 f^2 T_s^2} \cdot \frac{\sin 2\pi f T_s}{2\pi f T_s}
\]  

(A.8)

(A.9)

Now let's derive the frequency spectrum of the weighted single-interval raised-cosine pulse \( d(t) \) shown in Fig. A.2.

\[
d(t) = \begin{cases} 
-\frac{1-A}{2} (1 - \cos \frac{2\pi t}{T_s}) & |t| \leq T_s \\
0 & |t| > T_s 
\end{cases}
\]  

(A.10)

The derivatives of \( d(t) \) with respect to \( t \) are given by:

\[
\frac{d}{dt} d(t) = \begin{cases} 
-\frac{1-A}{2} \cdot \frac{2\pi}{T_s} \cdot \sin \frac{2\pi t}{T_s} & |t| \leq T_s \\
0 & |t| > T_s 
\end{cases}
\]

(A.11)

\[
\frac{d^2}{dt^2} d(t) = \begin{cases} 
-\frac{1-A}{2} \left(\frac{2\pi}{T_s}\right)^2 \cdot \cos \frac{2\pi t}{T_s} & |t| \leq T_s \\
0 & |t| > T_s 
\end{cases}
\]

(A.12)
Fig. A.2. A weighted single-interval raised-cosine pulse and its first three derivatives.
\[
\frac{d^3 d(t)}{dt^3} = \begin{cases} \\
\frac{1-A}{2} \left( \frac{2\pi}{T_s} \right)^2 \sin \frac{2\pi t}{T_s} + \left[ \delta(t-T_0) - \delta(t+T_0) \right] \text{if } |t| \leq T_s \\
0 \text{ if } |t| > T_s \end{cases} 
\]

From eq. (11) and (13)

\[
\frac{d^3 d(t)}{dt^3} = \left( \frac{2\pi}{T_s} \right)^2 \frac{d}{dt} \frac{d}{dt} d(t) - \frac{1-A}{2} \left( \frac{2\pi}{T_s} \right)^2 \left[ \delta(t+T) - \delta(t-T) \right] 
\]

(A.14)

From properties of Fourier transform, eq. (14) can be rewritten as:

\[
(j2\pi f)^3 D(f) = -\left( \frac{2\pi}{T_s} \right)^2 (j2\pi f) D(f) - \frac{1-A}{2} \left( \frac{2\pi}{T_s} \right)^2 (j2\pi f \sin 2\pi f T_s) 
\]

(A.15)

Solving for \( D(f) \)

\[
D(f) \left[ -(2\pi f)^3 + \left( \frac{2\pi}{T_s} \right)^2 \right] = -(1-A) \left( \frac{2\pi}{T_s} \right)^2 \sin 2\pi f T_s \\
D(f) = \frac{-1-A}{(2\pi f)^3 - \left( \frac{2\pi}{T_s} \right)^2} \frac{\sin 2\pi f T_s}{2\pi f} 
\]

(A.16)

As the frequency spectrum of \( s(t) \) is given by:

\[
S(f) = G(f) + D(f) 
\]

(A.17)

\[
S(f) = \frac{T_s}{1 - 4f^2 \frac{T_s^2}{2}} + \frac{(A-1)T_s}{1 - f^2 \frac{T_s^2}{2}} \frac{\sin 2\pi f T_s}{2\pi f T_s} 
\]

(A.18)

and

\[
S(0) = A T_s 
\]

(A.19)
Therefore the normalized PSD function of the SQAM signal $s(t)$ is:

$$
\left| \frac{S(f)}{S(0)} \right|^2 = \frac{1}{A^2} \left( \frac{1}{1 - 4f^2T_s^2} + \frac{A-1}{1 - f^2T_s^2} \right)^2 \left( \frac{\sin 2\pi fT_s}{2\pi fT_s} \right)^2 \quad (A.20)
$$
Appendix B

POWER SPECTRUM OF DUOBINARY SQAM

B.1 DUO-BINARY ENCODER

The block diagram of the duo-binary encoder is shown in Fig.B.1.

Fig.B.1. Block diagram of the Duo-binary encoder.

where

\[ a_k = \pm 1 \] with equal probability

\[ T_s = \text{one symbol interval delay-line (D flip/flop)} \]

and

\[ b_k = a_k + a_{k-1} \quad (B.1) \]
B.2 POWER SPECTRUM OF DUO-BINARY SQAM SIGNALS

Let the Duo-binary encoder have an impulse response $h_1(t)$ shown in Fig.B.2.

\[ a_k \rightarrow h_1(t) \rightarrow b_k \rightarrow s(t) \rightarrow c_k \]

Fig.B.2. Impulse response of the Duo-binary SQAM encoder

From (B.1), $h_1(t)$ is given by:

\[ h_1(t) = \delta(t) + \delta(t-T_s) \quad \text{(B.2)} \]

where

$\delta(t)$: Dirac delta function.

Therefore, the transfer function $H_1(f)$ is expressed as:

\[ H_1(f) = \mathcal{F}\{h_1(t)\} \]

\[ = 1 + e^{-j2\pi f T_s} \quad \text{(B.3)} \]

The impulse response $s(t)$ and transfer function $S(f)$ of the SQAM encoder are given in (2.3), (2.4) and (2.23). Therefore, the overall transfer function $V(f)$ is given by:

\[ V(f) = H_1(f) S(f) \]
\[ V(0) = 2A T_s \]

The normalized PSD function of the Duo-binary SQAM signal is given by:

\[
\frac{|V(f)|^2}{|V(0)|^2} = \frac{1 + \cos 2\pi f T_s}{2A^2} \left( \frac{1}{1 - 4T_s^2 f^2} + \frac{A - 1}{1 - T_s^2 f^2} \right) \times \left( \frac{\sin 2\pi f T_s}{2\pi f T_s} \right)^2
\]

\[
= \left( \frac{\cos f T_s}{A} \right)^2 \left( \frac{1}{1 - 4T_s^2 f^2} + \frac{A - 1}{1 - T_s^2 f^2} \right) \times \left( \frac{\sin 2\pi f T_s}{2\pi f T_s} \right)^2
\]

(B.6)

where \( A \) is the amplitude parameter of SQAM signal.

The normalized PSD is plotted in Fig.B.3. It is noticed that Duo-binary SQAM signal has a lower out-of-band energy and a faster spectral roll-off, compared to the conventional SQAM signal. Note that Duo-binary SQAM signals have a 1st spectral null at the Nyquist frequency (i.e. \( f_c = f_c/2 \)).
Fig. B.3. Normalized PSD of Duo-binary SQAM signals.
Appendix C

PROGRAMS TO SIMULATE PERFORMANCE OF SQAM MODEM

The following programs are used to simulate the performance of the SQAM modems, in an AWGN single-channel (linear and nonlinear) environment and in a nonlinear multichannel environment. The data source is an equiprobable NRZ signal with a bit rate of 120 Mb/s. The simulation here is based on the equivalent baseband concepts of the modulated system. In the simulation, the earth station HPA (a high power amplifier), operating in the saturation mode, is approximated by an ideal hardlimiter. Using this simulation, the P(e) performance of the required system model can be evaluated, and power spectrum, eye-pattern, waveshape, phase transition and signal state-space diagram can be obtained. The simulation models, in a single-channel environment and in a multi-channel environment, are shown in Fig.3.9 and Fig.4.2(or Fig.4.10) respectively.

This program was developed from our available software computer simulation systems called COMSIM, which was developed at the Digital Communication Group of the University of Ottawa.
// EXEC PLOTG, FORM=0111, PARM.GO='SIZE=250K', TIME=5
// FORT.SYSIN DD

*********************************************************************************

PROGRAM TO SIMULATE THE P(E) PERFORMANCE OF SQAM MODEM IN
AN AWGN SINGLE-CHANNEL ENVIRONMENT.

*********************************************************************************

COMMON LDIM, IOFF
COMMON/ PARA/LSAMPL, NSYMB, NO1, NO2, BAUD, SBANDW
COMPLEX TF(2048), TF1(2048), TF2(2048), TF3(2048), TF4(2048)
COMPLEX DATA(2048), DATA1(2048)
DIMENSION PEI(25), EBN0(25), IPNI(128), IPNQ(128)
DIMENSION IWK(12), PO(65), IPN1(128), IPNQ1(128)
DIMENSION XARRAY(27), YARRAY(27), APEI(25)
COMMON /AA/A

INITIALIZE PROGRAM.

IOFF=8
ICEI=4
NSYMB=128
LSAMPL=16
BAUD=60.
POF=0.
DATA NSNR/25/
DATA NRUNS/4/
DATA FBW1, FBW2, FBW3, FBW4/33.0, 30.0, 40.0, 40.0/
DATA ALPHAL, ALPHAL2/0.4, 0.4/
SBANDW=FLOAT(LSAMPL)*BAUD
LDIM=LSAMPL*NSYMB
NO1=LDIM/2+1
NO2=NO1+1

END OF INITIALIZATION.

START OF COMPUTATIONS.

A=0.60
DO 500 500
     KM=1, NRUNS
     A=A+.10
     CALL LOAD4(DATA, IPNI, IPNQ)
     CALL BUT(TF, FBW1, ICEI)
     CALL HLIM(DATA)
     CALL RCOS(SBANDW, FBW2, TF, ALPHAL, LDIM, POF)
     CALL ENERGY(DATA, EB, BAUD)
     CALL FILTER(DATA, TF)
     CALL HHGG(TF, PNOISE)
     CALL DECOD4(DATA, PNOISE, IPNI, IPNQ)
     #EBNO, PEI, NSNR, EB
     DO 5 I=1, NSNR
     5 WRITE(6, 150)

      FORMAT(5X, 'EB/NO', 10X, 'PROB. OF. ERROR', /)
WRITE(6,172) (EBNO(I),PEI(I),I=1,NSNR)  
FORMAT(5X,F5.1,10X,E13.6) I=1,NSNR  
YARRAY(I)=PEI(I)  
DRAW(XARRAY,YARRAY,NSNR,KM,NRUNS)  
CONTINUE

C***********************************************
C THE FOLLOWING SUBROUTINE PERFORMS THE FILTERING
C PROCESS ON THE DATA SEQUENCE.
C***********************************************
SUBROUTINE FILTER(SIGNAL,TF)  
COMMON /LDIM,IOFF
COMPLEX SIGNAL(I),TF(1)  
DIMENSION IWK(12)
CALL FFT2C(SIGNAL,11,IWK) I=1,LDIM  
1 SIGNAL(I)=CONJG(SIGNAL(I)*TF(I))  
CALL FFT2C(SIGNAL,11,IWK) I=1,LDIM  
2 SIGNAL(I)=CONJG(SIGNAL(I))/FLOAT(LDIM)  
RETURN
END

C***********************************************
C THIS SUBROUTINE COMPUTES THE EFFECTIVE NOISE (PNOISE)
C AT THE OUTPUT OF THE RECEIVE FILTER
C***********************************************
SUBROUTINE HHGG(TF,PNOISE)  
COMMON LDIM
COMMON/ PARA/ LSAMPL, NSYM, NO1, NO2, BAUD, SBANDW
COMPLEX SUM=0.0  
DO 1 L=1,LDIM
HH=(ABS(TF(L)))**2  
SUM=SUM+HH
1 PNOISE=SUM*SBANDW/FLOAT(LDIM)/2.  
PP=PNOISE*2./BAUD
RETURN
END
C*******************************************************************************
C BIT ENERGY COMPUTATION
C*******************************************************************************

SUBROUTINE ENERGY(DATA, EB)
COMMOM
COMMON/ PARA/ LSAMPL, NSYM, NO1, NO2, BAUD, SBANDW
COMPLEX
WATTS=0.
DO 1, I=1, LDIM
1 WATTS=WATTS+((CABS(DATA(I)))*2.)
CONTINUE
WATTS=WATTS/FLOAT(LDIM)
EB=WATTS/(2.*BAUD)
RETURN
END

C*******************************************************************************
C SUBROUTINE HLIM(DATA)
C*******************************************************************************

COMMON
COMMON/ PARA/ LSAMPL, NSYM, NO1, NO2, BAUD, SBANDW
COMPLEX
DO 10, I=1, LDIM
10 DATA(I)=DATA(I)/CABS(DATA(I))
RETURN
END

C*******************************************************************************
C THIS SUBROUTINE DECODES THE PROCESSED DATA
C*******************************************************************************

SUBROUTINE DECOD4(DATA, PNOISE, IPNI, IPNQ, EBN0, PEI, NSNR, EB)
COMMOM
COMMON/ PARA/ LSAMPL, NSYM, NO1, NO2, BAUD, SBANDW
COMPLEX
DIMENSION
DATA NERROR, NOLD, NOF, NN, MI, MQ/0, 0, 0, 0, 0, 1, 1/
IF (IOFF.EQ.0) GO TO 111
DO 9, K=1, IOFF
9 XX=AIMAG(DATA(1))
LD=LDIM-1
DO 5, KK=1, LD
5 DATA(KK)=CMPLX(REAL(DATA(KK)), AIMAG(DATA(KK+1)))
9 DATA(LDIM)=CMPLX(REAL(DATA(LDIM)), XX)
C SYNCHRONIZE THE RECEIVED DATA
111 LFU=2*NSYM
300 DO 10, K=1, LDIM
10 - 116 -
NEW=0
DO 200 J=1,NSYMB
    J1=K+(J-1)*LSAMPL
    IF(J1.GT.LDIM)
       XB=REAL(DATA(J1))
       YB=AIMAG(DATA(J1))
       IF(SIGN(1.,XB).EQ.IPNI(J)) NEW=NEW+1
       IF(SIGN(1.,YB).EQ.IPNO(J)) NEW=NEW+1
       CONTINUE 399,12,13
       NN=NN+1
       NN0=0
200    IF(NEW-NOLD) GO TO 10
12     NOLD=NEW
13     NOF=K
399    IF(NOLD.EQ.LFU) GO TO 10
10     CONTINUE
400    WRITE(6,60)
60     FORMAT(5X,'SAMPLING TIME IS DELAYED TO ',I2,
          $' SIXTEENTHS OF THE SYMBOL INTERVAL',/)
      IF(NOF1.EQ.0) GO TO 230
      LO=LDIM-1
      DO 250 I=1,NOF1
         AMP=DATA(1)
         DO 240 J=1,LO
            DATA(J)=DATA(J+1)
240            DATA(LDIM)=AMP
250         CONTINUE
230         CONTINUE
         EOI=FLOAT(NSYMB)+1.
         SNR=NOISE*EB/5.
         SIGMA=SQRT(SNR)
         DO 110 J=1,LSAMPL
            EI=0.
            EQ=0.
            DO 100 K=1,NSYMB
               J1=(K-1)*LSAMPL+J
               AXBAR=(REAL(DATA(J1))+REAL(DATA(J1+1)))/2.
               AYBAR=(AIMAG(DATA(J1))+AIMAG(DATA(J1+1)))/2.
               IF(SIGN(1.,AXBAR).NE.IPNI(K)) EI=EI+1
               IF(SIGN(1.,AYBAR).NE.IPNO(K)) EQ=EQ+1
               ARG=ABS(AXBAR)/SIGMA
               ARGQ=ABS(AYBAR)/SIGMA
               EI=EI+ERFC(ARG)/2.
               EQ=EQ+ERFC(ARGQ)/2.
100            CONTINUE
110            IF(EOI.LE.EI) GO TO 120
         EOI=EI
         MI=J
120            CONTINUE
         IF(EQ.LE.EQ) GO TO 110
EOQ=EQ
MQ=J
CONTINUE

MOFF=IABS(MI-MQ)
IF (MOFF.NE.0) WRITE(6,160) MOFF
FORMAT(5X,'SAMPLING POINTS FOR I AND Q CHANNELS DIFFER BY',
       '12, SIXTEENTHS OF THE SYMBOL INTERVAL',/)
MOFF=(MQ+MI)/2-1
OFF=(FLOAT(MOFO)+FLOAT(NOF1))/FLOAT(LSAMPL)
WRITE(6,161) OFF

FORMAT(5X,'RECEIVED DATA IS DELAYED BY ',E9.3,' SYMBOLS',/)
DO 2 K=1,NSYMB
JI=(K-1)*LSAMPL+MI
JQ=(K-1)*LSAMPL+MQ
AXBAR=(REAL(DATA(JI))+REAL(DATA(JI+1)))/2.
AYBAR=(AIMAG(DATA(JQ))+AIMAG(DATA(JQ+1)))/2.
INDEXX=0
INDEXY=0
IF(SIGN(1.,AXBAR).NE.IPNI(K)) INDEXX=1
IF(SIGN(1.,AYBAR).NE.IPNQ(K)) INDEXY=1
IF((INDEXX.EQ.1).OR.(INDEXY.EQ.1)) NERROR=NERROR+1
C COMPUTE THE PROBABILITY OF ERROR FOR THIS SYMBOL
DO 4 M=1,NSNR
XM=FLOAT(M)/10.
SNR=PNOISE*EB/(10.**XM)
SIGMA=SQR(2/SNR)
ARG=(ABS(AXBAR))/(SIGMA*SQR(2.))
IF (ARG.GT.12.) ARG=12.
EX=ERFC(ARG)/2.
IF (INDEXX.EQ.1) EX=1.-EX
ARG=(ABS(AYBAR))/(SIGMA*SQR(2.))
IF (ARG.GT.12.) ARG=12.
EY=ERFC(ARG)/2.
IF (INDEXY.EQ.1) EY=1.-EY
IF (EX.LT.1.1E-15) EX=0.
IF (EY.LT.1.1E-15) EY=0.
4 PEI(M)=PEI(M)+((EX+EY)/2.)
2 CONTINUE
C DO 5 I=1,NSNR
C PEI(I)=PEI(I)/FLOAT(NSYMB)
C5 EBN0(I)=FLOAT(I)
WRITE(6,150)
FORMAT(5X,'EB/NO',10X,'PROB. OF. ERROR',/)
WRITE(6,151) (EBNO(I),PEI(I),I=1,NSNR)
C WRITE 6,152 NERROR
FORMAT(5X,F5.1,10X,E13.6)
C RETURN
END

- 118 -
SUBROUTINE BUT(TF,FN,ICEL)  
COMMON LDIM,IOFF  
COMMON/ PARA/LSAMPL,NSYM,N01,N02,BAUD,SBANDW  
COMPLEX TF(1)  
IF(1)=CMPLX(1.0,0.0)  
DO 10 I=2,N01  
J=I-1  
A2=1.0/CMPLX(1.+(A1*FLOAT(J)/FBNOR)**(2*ICEL))  
DO 20 TF(I)=CMPLX(A2,0.)  
20 RETURN  
END

C THIS SUBROUTINE GENERATES TWO CHANNELS OF SQAM SIGNAL.

SUBROUTINE LOAD4(DATA,IPNI,IPNQ) COMMON LDIM,IOFF  
COMMON/ PARA/LSAMPL,NSYM,N01,N02,BAUD,SBANDW  
COMPLEX DATA(1)  
DIMENSION NX(7),NY(7),IPNI(1),IPNQ(1),RI(16),RQ(16)  
DO 3 I=1,7  
NX(I)=-1  
NY(I)=-1  
3 DATA  
I=0  
J=6  
KKK=2**JLAST  
DO 1 C GENERATING ONE SYMBOL.  
IF (I.GE.JLAST) I=0  
IF (J.GE.JLAST) J=0  
I=I+1  
J=J+1  
NX(I)=NX(J)*NX(I)  
NY(I)=NY(J)*NY(I)  
IPNI(K)=NX(I)  
1 IPNQ(K)=NY(I)  
- 119 -
C LOADING INTO THE ARRAY TO BE FOURIER TRANSFORMED:
DO 10 K=1,KKK
   K=K-1
   IF (K.KK) KML=KKK
   MI=IPNI(KML)
   MQ=IPNQ(KML)
   CALL SIG(MI,IPNI(K),RI)
   CALL SIG(MQ,IPNQ(K),RQ)
   J1=(K-1)*LSAMPL
   DO 10 I=1,16
      DATA(J1-I)=CMPLX(RI(I),RQ(I))
   END
   IF (IOFF.EQ.0) GO TO 20
   DO 15 I=1,IOFF
      A=AIMAG(DATA(LDIM))
      DO 13 L=2,LDIM
         K=LDIM+2-L
         DATA(K)=CMPLX(REAL(DATA(K)),AIMAG(DATA(K-1)))
      END
      DATA(I)=CMPLX(REAL(DATA(I)),A)
      CONTINUE
   END
   RETURN
END

C ********************************************************

SUBROUTINE SIG(M,IP,R)
C ********************************************************

COMMON /AA/A
REAL P=3.14159265/16.
IF(M.EQ.IP) GOTO 20
IF(IP.EQ.1) GOTO 10
DO 1 I=1,16
   R(I)=COS((I+0.5)*P)
10  GOTO 11
11  R(I)=-COS((I+0.5)*P)
20  IF(IP.EQ.1) GOTO 30
21  R(I)=-A-(1.0-A)*COS(2.*(I+0.5)*P)
30  GOTO 31
31  R(I)=A+(1.0-A)*COS(2.*(I+0.5)*P)
40  CONTINUE
RETURN
END
THIS SUBROUTINE DRAWS PROBABILITY OF ERROR CURVES FOR
P(E) AS LOW AS 1.0E-8 AND A C/N RATIO AS HIGH AS 30 DB.

SUBROUTINE DRAW(XARRAY, YARRAY, NSNR, JCURV, JLAST)
DIMENSION XARRAY(27), YARRAY(27), X(9), Y(9)

IF(JCURV.GT.1) GOTO 2

ESTABLISH THE SURFACE AREA.

CALL PLOTS(30.0, 27.5)

ESTABLISH THE ORIGIN.

CALL PLOT(3.0, 3.0, -3)

DRAW THE LOGARITHMIC Y-AXIS.

CALL LGAXS(0.0, 0.0, 20H_PROBABILITY OF ERROR, 20, 18,
+90., 1.0E-8, .4)

DRAW THE LINEAR X-AXIS.

CALL AXIS(0.0, 0.0, 11HEB/NO IN DB, -11,
+14.0, 0.0, 6.0, 1.0)

WRITE THE 'TITLE OF THE GRAPH.

CALL SYMBOL(4.0, 23.0, 0.49, 14HSQAMBER CURVE, 0.0, 20)

PLOT THE IDEAL CURVE.

DO JJ=1,7

J(J)=FLOT(JJ)

Y(J)=.5*ERFC(SQRT(1.0**(.1*JJ)))

X(8)=6.0
X(9)=1.0
Y(8)=1.0E-8
Y(9)=0.4

CALL NEWPEN(2)

CALL LGLIN(X, Y, 7, 1, 0, 0, 1)

CONTINUE

CALL LGLIN(X, Y, 7, 1, 0, 0, 1)

PLOT THE DATA IN LOG-LINEAR MODE.

M=NSNR
DO 1 K=1, M

KK=K+5
IF(KK.GT.M) KK=M

XARRAY(K)=XARRAY(KK)
YARRAY(K)=YARRAY(KK)
IF(YARRAY(K).LE.1.0E-8) GOTO 3

GOTO 3

2 M=K-1
GOTO 4
I=M+1
II=I+1
XARRAY(I)=6.0
XARRAY(II)=1.0
YARRAY(I)=1.E-8
YARRAY(II)=0.4
CALL NEWPEN(1)
CALL LGLIN(XARRAY,YARRAY,M,1,0,0,1)
IF(JCURV.EQ.JLAST) CALL PLOT(0.0,0.0,0,999)
RETURN
END

C**********************************************************************************************
C RAISED COSINE FILTER WITH ARBITRARY ALPHA
C**********************************************************************************************

SUBROUTINE RCOS(SBANDW,FBANDW,TF,ALPHA,LDIM,EOF)
COMPLEX
COMPLEX
COMMON
C******** FF(X) IS THE INVERSE BASEBAND SPECTRUM
FF(X)=2.*X*(1.-((X/3.141592)**2)/(SIN(2.*X))
C
B(X)=(1.-A)/(X/(3.141592)**2-1.)C # 1./(1.-4.*(X/3.141592)**2)
C
FF(X)=2.*X/(B(X)*SIN(2.*X))
C
QQ=2.
NO=LDIM/2
NO1=NO+1
IF (ALPHA.EQ.0) ALPHAC=0.0001
FN=LDIM*(FBANDW/2)/SBANDW
F1=(1.-ALPHA)*FN
F2=(1.+ALPHA)*FN
IFN=1IFIX(FN)
IF1=1IFIX(F1)+1
IF2=1IFIX(F2)+1
A1=3.141592/(2.*FLOAT(IFN))
TF(1)=CMPLX(1.0,0.0)
DO 8 I=2,IF1
   J=I-1
   A2=FF(FLOAT(J)*A1)
   TF(I)=CMPLX(A2,0.0)
8 CONTINUE

JK=IF1+1
DO 9 I=J-1
   A3=QQ
   IF(I.NE.FN) A3=FF(FLOAT(I)*A1)
   A=(3.141592/(2.0*ALPHA))*(FLOAT(I)/FLOAT(IFN))-1.
   TF(J)=CMPLX((0.5*(1.0-SIN(X)))*A,0.0)
   TF(J)=TF(J)*CMPLX(A,0.0)
9

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9
JH=IF2+1
DO
10
TF(I)=CMPLX(0.0,0.0)

CONTINUE

I=JH,NO1

CONTINUE

10
NO2=NO1+1
DO
5
TF(I)=TF(LDIM+2-I)

CONTINUE

I=NO2,LDIM

CONTINUE

5
IFOF=IFIX(FOF*FLOAT(LDIM)/SBANDW)
IF(IFOF.EQ.0)
IF(IFOF.LT.0) GO TO 50
DO
20
XX=TF(LDIM)
JJ=LDIM-I
DO
30
TF(LDIM+I-I)=TF(LDIM-I)

CONTINUE

I=1,II

RETURN

50
JJ=LDIM-I
DO
70
XX=TF(1)
DO
60
TF(LDIM)=XX

CONTINUE

70
RETURN

END

***********************************************************************
C                      GENERATES  TWO CHANNELS  OF  MSK SIGNAL
C                      ***********************************************************************

SUBROUTINE LOAD7 (DATA,IPNI,IPNQ)
COMMON
COMMON/ PARA/LSAMPL, NSYMB, NO1, NO2, BAUD, SBANDW
COMPLEX
DIMENSION IPNI(1), IPNQ(1), NX(7), NY(7), RI(16), RQ(16)
DATA NX(1), NX(2), NX(3), NX(4), NX(5), NX(6), NX(7)/-1,-1,-1,
# -1,-1,-1,-1,
# DATA NY(1), NY(2), NY(3), NY(4), NY(5), NY(6), NY(7)/-1,-1,-1,
# 1,1,-1,-1/
IOFF=8
DATA
DO
1
JLAST,J,1/7,6,0/
K=1,128
1=0
J=0
I=I+1
J=J+1
NX(I)=NX(J)*NX(I)
NY(I)=NY(J)*NY(I)
IPNI(K)=NX(I)

1  DO 10 K=1,128
   CALL MSK(IPNI(K),RI)
   CALL MSK(IPNI(K),RQ)
   JI=(K-1)*16
10  DO 10 I=1,16
    IF(IOFF.EQ.0) THEN
       DATA(J1+I)=CMPLX(RI(I),RQ(I))
       GOTO 20
    END IF
    DO 15 I=1,IOFF
    A=AIMAG(DATA(2048))
    DO 13 L=2,2048
    K=2050-L
    13  DATA(K)=CMPLX_REAL(DATA(K)),AIMAG(DATA(K-1))
    DATA(1)=CMPLX_REAL(DATA(1)),A)
15  CONTINUE
20  RETURN

C*************************************************************************
C*************************************************************************
C*************************************************************************

SUBROUTINE MSK(IP,R)

REAL 
R(16)

P=3.14159265/16.

IF(IP.EQ.1) THEN
   GOTO 10
1   DO 1 I=1,16
    R(I)=-SIN((I+0.5)*P)
10  CONTINUE

GOTO 11
11  DO 11 I=1,16
    R(I)=SIN((I+0.5)*P)
11  CONTINUE

RETURN
END
PROGRAM TO PLOT EYE-PATTERNS, WAVEFORMS, STATE-SPACE
DIAGRAMS AND PHASE TRANSITIONS OF SQAM BASEBAND SIGNALS.

COMMON/LDIM,IOFF
COMMON/Para/LSAMPL,NSYMB,NO1,NO2,BAUD,SBANDW
COMPLEX TF(2048),TF1(2048),TF2(2048),TF3(2048),TF4(2048)
COMPLEX DATA(2048),DATA1(2048)
DIMENSION PEI(25),EBNO(25),IPNI(128),IPNQ(128)
DIMENSION IWK(12),PO(65),IPNIL(128),IPNQL(128)
DIMENSION XARRAY(27),YARRAY(27),APEI(25)
REAL TR(514)
COMMON /AA/A

IOFF=8
ICEL=4
NSYMB=128
LSAMPL=16
BAUD=60.
DATA FBW1,FBW2,FBW3,FBW4,33.0,30.0,40.0,40.0/NRUNS/1/
DATA ALPHA1,ALPHA2/0.4,0.4/
SBANDW=FLOAT(LSAMPL)*BAUD
LDM=LSAMPL*NSYMB
NO1=LDM/2+1
NO2=NO1+1

START OF COMPUTATIONS.

A=0.70
DO 500 500
A=A+1.0
CALL LOAD4(DATA,IPNI,IPNQ)
CALL TRAN(DATA,TR)
CALL (DATA(I),TR(I),I=1,512)
WRITE(6,100) FORMAT(5X,F12.7,20X,F12.7)
100 CALL HLIM(DATA)
CALL BUT(TF,FBW1,ICEL)
CALL FILTER(DATA,TF)
CALL EYEQQ(DATA,16,128,KM,NRUNS)
CALL WAVSH2(DATA)
CALL SPACE(DATA,KM,NRUNS)
CALL CONTINUE

500 STOP
END
SUBROUTINE WAVSH2(DATA)
COMMON LDIM, LMOD, LOFF
COMMON/PARA/LSAMPL, NSYM, NO1, NO2, BAUD, SBANDW
DIMENSION DATA2(259), DATA3(259), MOD(2)
COMPLEX DATA(1)
C ESTABLISH THE SURFACE AREA.
C CALL ESTABLISH THE PLOTS(36.0, 27.5)
C CALL WRITE THE TITLE OF THE ORIGIN.
C WRITE THE TIME AXIS.
C CALL DRAW THE PLOT(2.5, 19.5, -3)
C DRAW THE AMPLITUDE AXIS.
C CALL AXIS(0.0, -1.325, 0.0, 0.0, 0.5)
C ESTABLISH THE SURFACE AREA.
C CALL AXIS(0.0, -0.2, 0.1H, 1.4, 0, 90.0, -1.0, 0.5)
C PLOT THE DATA.
DATA2(258) = 0.0
DATA2(259) = 0.5
DATA3(258) = 0.5
DATA3(259) = 8.
DO 4 KK = 1, 257
DATA2(KK) = REAL(DATA(KK))
DATA3(KK) = FLOAT(KK)
4 CALL ESTABLISH LINE(DATA3, DATA2, 257, 1, 0, 0)
C CALL DRAW THE ORIGIN.
C CALL DRAW THE TIME AXIS.
C CALL DRAW THE AMPLITUDE AXIS.
C CALL AXIS(0.0, -1.325, 0.0, 0.17, 0.5)
C CALL AXIS(0.0, -0.2, 0.1H, 1.4, 0, 90.0, -1.0, 0.5)
C CALL PLOT THE DATA.
DO 7 KK = 1, 256
DATA2(KK) = REAL(DATA(KK + 256))
DATA3(KK) = FLOAT(KK)
7 CALL LINE(DATA3, DATA2, 257, 1, 0, 0)
CALL PLOT(0.0, 0.0, 999)
RETURN
END
**THIS SUBROUTINE DRAWS THE EYE-PATTERNS FOR A SYMBOL**

**DURATION**.

**SUBROUTINE EYEQQ(DATA,LSAMPL,NSYMB,JCURV,JLAST)**

**DIMENSION DATA2(64),DATA3(64)**

**COMPLEX**

**IF(JCURV.GT.1)**

**GOTO 3**

**CALL ESTABLISH THE SURFACE AREA.**

**CALL PLOTS(20.0,27.5)**

**CALL THE ORIGIN.**

**CALL PLOT(3.0,7.0,-3)**

**CALL WRITE THE TITLE OF THE GRAPH.**

**CALL SYMBOL(0.0,18.0,0.49,16HSQAM EYE DIAGRAM,0.0,16)**

**CALL DRAW THE TIME AXIS.**

**CALL AXIS(0.0,-6.0,1H,-1,12,0.0,0.0,0.3,0)**

**CALL DRAW THE AMPLITUDE AXIS.**

**CALL AXIS(0.0,-6.0,1H,1,1,90.0,0.0,0.60)**

**CALL PLOT THE DATA.**

**P=0.0**

**JJ=2*LSAMPL**

**DATA2(JJ+1)=P**

**DATA2(JJ+2)=0.6**

**DATA3(JJ+1)=1.0**

**DATA3(JJ+2)=3.0**

**M2=NSYMB/2**

**DO 4 I=1,JJ**

**DO 2 I=(KK-1)*2*LSAMPL+1**

**DATA2(I)=REAL(DATA(I))**

**CALL KK=1,M2**

**2 CALL DATA3(I)=FLOAT(I)**

**CALL LINE(DATA3,DATA2,JJ,1,0,0)**

**CONTINUE**

**P=P-4.0**

**IF(JCURV.EQ.JLAST)**

**CALL PLOT(0.0,0.0,0.999)**

**RETURN END**

**THIS SUBROUTINE DRAWS SIGNAL STATE-SPACE DIAGRAMS.**

**SUBROUTINE SPACE(DATA,JCURV,JLAST)**

**REAL X(2050),Y(2050)**

**COMPLEX DATA(1)**

**IF(JCURV.GT.1)**

**GOTO 2**

**CALL ESTABLISH THE SURFACE AREA**

**CALL PLOTS(30.0,27.5)**

**CALL THE ORIGIN.**

**CALL PLOT(15.0,12.5,-3)**

**TITLE**
CALL SYMBOL(-6.0,12.5,0.49,13) HSPACE DIAGRAM,0.0,13
CALL AXI$(-8.0,0.0,0.1H,1.16,0.0,0.0,-2.0,0.25)
CALL AXI$(0.0,-8.0,1H,1.16,0.90,0.0,-2.0,0.25)
X(2049)=0.0
X(2050)=0.25
Y(2049)=0.0
Y(2050)=0.25

C PLOT
DO 21 I=1,2048
X(I)=REAL(DATA(I))
Y(I)=AIMAG(DATA(I))

21 CONTINUE
CALL LINE(X,Y,2048,1,0,0)
IF(JCURV.EQ.JLAST) CALL PLOT(0.0,0.0,999)
WRITE(6,61)
FORMAT('X',SPC-SPACE-DIAGRAM IS REQUESTED ---')
RETURN

END

C ************************************************************************************************************
C DRAW PHASE TRANSITIONS OF SQAM SIGNAL.
C ************************************************************************************************************

SUBROUTINE TRAN(DATA,TR)
REAL X(514),Y(514),TR(514),B(514)
COMPLEX DATA(1)

C ESTABLISH THE SURFACE AREA.
CALL PLOTS(36.0,27.5)
C ESTABLISH THE ORIGIN.
CALL PLOT(2.5,13.5,-3)
C DRAW THE TIME AXIS.
CALL AXIS(0.0,0.0,1H,-1.32,5,0.0,0.0,1.0)
C DRAW THE AMPLITUDE AXIS.
CALL AXIS(0.0,-2.0,1H,1.40,90.0,-2.0,1.0)
TR(513)=0.0
TR(514)=1.0
B(513)=0.5
B(514)=16.
DO 4 I=1,512
X(I)=REAL(DATA(I))
Y(I)=AIMAG(DATA(I))
TR(I)=ATAN(Y(I)/X(I))
B(I)=FLOAT(I)

4 CONTINUE
CALL LINE(B,TR,512,1,0,0)
CALL PLOT(0.0,0.0,999)
RETURN
END
JOB 'SUN2',CLASS=B
EXEC PLOTG,FORM=0111,PARM.GO='SIZE=500K',TIME=5

****** PROGRAM TO COMPUTE AND PLOT THE P.S.D. AND THE OUT-OF-BAND ENERGY OF SQAM BASEBAND SIGNALS. ******

COMMON LDIM,IOFF
COMMON/ PARA/LSAMPL,NSYMB,NO1,NO2,BAUD,SBANDW
COMPLEX DATA(16384)
DIMENSION PEI(25),EBNO(25),IPNI(1024),IPNQ(1024)
DIMENSION IWK(12),PO(129),IPNI1(1024),IPNQ1(1024)
DIMENSION XARRAY(27),YARRAY(27),APEI(25),TPO(129)

IOFF=8
NSYMB=1024
LSAMPL=16
BAUD=60.
DATA SBANDW=FLOAT(LSAMPL)*BAUD
LDIM=LSAMPL*NSYMB
NO1=LDIM/2+1
NO2=NO1+1

START OF COMPUTATIONS.

A=0.60
DO 500 A=A+.20
CALL LOAD4(DATA,IPNI,IPNQ)
CALL HLIM(DATA)
CALL SPECT(DATA,PO,TPO)
CALL DRAW5(TPO,KM,NRUNS)

500 STOP
END

*****************************************************************************
C THIS SUBROUTINE GENERATES BASEBAND SQAM SIGNALS.
*****************************************************************************

SUBROUTINE LOAD4(DATA,IPNI,IPNQ)
COMMON LDIM,IOFF
COMMON/ PARA/LSAMPL,NSYMB,NO1,NO2,BAUD,SBANDW
COMPLEX DATA(1)
DIMENSION NX(10),NY(10),IPNI(1),IPNQ(1),RI(16),RQ(16)
DO 3 I=1,10

3

NY(1)=-1
NY(2)=-1
NY(3)=1
NY(4)=1
NY(5)=1
NY(6)=1
NY(7)=1
NY(8)=-1
NY(9)=1
NY(10)=1
DATA
I=0
J=9
KKK=2**JLAST
DO 1 GENERATING 1 ONE SYMBOL.
   K=1,KKK
   I=0
   J=0
   IF (I.GE.JLAST)
     I=I+1
   IF (J.GE.JLAST)
     J=J+1
   NX(I)=NX(J)*NX(I)
   NY(I)=NY(J)*NY(I)
   IPNI(K)=NX(I)
   IPNQ(K)=NY(I)
1   C

C LOAD INTO THE ARRAY TO BE FOURIER TRANSFORMED.
DO 10 K=1,KKK
   KM1=K-1
   IF (KM1.EQ.0) KM1=KKK
   MI=IPNI(KM1)
   MQ=IPNQ(KM1)
   CALL SIG(MI,IPNI(K),RI)
   CALL SIG(MQ,IPNQ(K),RQ)
   J1=(K-1)*LSAMPL
   DO 10 I=1,16
      DATA(J1+I)=CMPLX(RI(I),RQ(I))
   IF (IOFF.EQ.0) GO TO 20
   DO 15 I=1,IOFF
      A=AIMAG(DATA(LDIM))
   DO 13 L=2,LDIM
      K=LDIM+2-L
      DATA(K)=CMPLX(REAL(DATA(K)),AIMAG(DATA(K-1)))
   13      DATA(1)=CMPLX(REAL(DATA(1)),A)
   15 CONTINUE
   20 RETURN

END
C**************************************************************************
C GENERATES TWO 'CHANNELS' OF MSK SIGNAL
C**************************************************************************

SUBROUTINE LOAD7  (DATA, IPNI, IPNQ)
COMMON
COMMON/ PARA/ LSAMPL, NSYMB, NO1, NO2, BAUD, SBANDW
COMPLEX
DIMENSION NX(10), NY(10), IPNI(1), IPNQ(1), RI(16), RQ(16)
DO 3  I=1,10
NX(1)= -1
3 NY(I) = -1
NY(2) = -1
NY(3) = 1
NY(4) = -1
NY(5) = 1
NY(6) = -1
NY(7) = -1
NY(8) = -1
NY(9) = 1
NY(10) = 1
DATA JLAST/10/
I=0
J=9
KKK=2**JLAST
DO 10 1  K=1, KKK
1  GENERATING ONE
K=1, KKK
IF (I.GE.JLAST) GO TO 10
IF (J.GE.JLAST) GO TO 20
I=I+1
J=J+1
NX(I) = NX(J)*NX(I)
NY(I) = NY(J)*NY(I)
IPNI(K) = NX(I)
10 IPNQ(K) = NY(I)

C LOAD INTO THE ARRAY TO BE FOURIER TRANSFORMED.
DO 10 10  K=1, KKK
CALL MSK(IPNI(K), RI)
CALL MSK(IPNQ(K), RQ)
J1 = (K-1)*LSAMPL
DO 10 10  I=1,16
DATA(J1+I) = CMPX(RI(I), RQ(I))
GO TO 20
10  IF (IOFF.EQ.0) GO TO 20
DO 15 15  I=1, IOFF
A = AIMAG(DAT(A(LDIM)))
DO 13 13  L=2, LDIM
K = LDIM + 2-L
DATA(K) = CMPX(REAL(DAT(A(K))), AIMAG(DAT(A(K-1))))
13 DATA(1) = CMPX(REAL(DAT(A(1))), A)
15 CONTINUE
20 RETURN
END

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C* THIS SUBROUTINE COMPUTES THE P.S.D. AND OUT-OF-BAND ENERGY.
C*---------------------------------------------------------------------
SUBROUTINE SPECT(SIGNAL, PO, TPO)
COMMON LDIM, IOFF
COMMON PARA, LSAMPL, NSYM8, NO1, NO2, BAUD, SBANDW
COMPLEX SIGNAL(1)
COMPLEX CWK(258)
DIMENSION X(32768), PO(1), IWK(8), WK(128), TPO(129)
S=0.
DO 10 I=1, LDIM
L=I-IOFF
IF(L.LE.0) L=L+LDIM
X(I)=REAL(SIGNAL(I))
X(I+16384)=AIMAG(SIGNAL(L))
10 S=S+X(I)+X(I+16384)
S=S/FLOAT(32768)
DO 20 I=1, 32768
20 CALL FTFPS(X, Y, 32768, 256, 0, PO, PSY, XPS, IWK, WK, CWK, IER)
P MAX=PO(1)
TPO(129)=PO(129)
C
C**** CALCULATE P.S.D. ***************
C
C30 DO 30 I=1, 129
30 TPO(I)=-10.*ALOG10(PO(I)/P MAX)
C WRITE (6, 1) TPO(I), I=1, 129
C1 FORMAT (2X, 'POWER SPECTRUM'/, 10(2X, F6.1))
C
C**** CALCULATE OUT-OF-BAND ENERGY ************
C
C40 DO 40 I=1, 127
40 TPO(129-I)=TPO(130-I)+2.*PO(129-I)
C TPO(1)=TPO(2)+PO(1)
TP MAX=TPO(1)
DO 50 I=1, 129
50 TPO(I)=-10.*ALOG10(TPO(I)/TP MAX)
C WRITE (6, 2) TPO(I), I=1, 129
C2 FORMAT (2X, 'OUT OF BAND TO TOTAL POWER RATIO'/
& 10(2X, F6.1))
C
RETURN
END
C******************************************************************************
C THIS SUBROUTINE DRAWS THE NORMALIZED P.S.D.
C******************************************************************************

SUBROUTINE DRAW5(TPO, JCURV, JLAST),
COMMON LDIM, IOFF
COMMON PARA/LSAMPL, NSMB, NO1, NO2, BAUD, SBANDW
DIMENSION TPO(1), DATA2(131), DATA3(131)
COMPLEX DATA(1)
C
IF(JCURV.GT.1) GOTO 2
C
CALL ESTABLISH THE SURFACE AREA.
C
CALL PLOTS(20, 0, 27, 5)
C
CALL ESTABLISH THE ORIGIN.
C
CALL PLOT(3, 0, 7, 0, -3)
C
WRITE THE TITLE OF THE GRAPH.
C
CALL SYMBOL(5, 0, 17, 0, 0, 35, 19, HSQAM, POWER SPECTRUM, 0, 0, 24)
C
DRAW THE FREQ AXIS.
C
CALL AXIS(0, 0, 0, 0, 20, HNORMALIZED FREQUENCY, -20, 5, 0, 0, 0, 0, 1, 0)
C
DRAW THE NORMALIZED PSD AXIS.
C
CALL AXIS(0, 0, 0, 0, 18, HNORMALIZED PSD DB, 18, 9, +90.,
#-80, 0, 10.)
DATA2(66) = -80.
DATA2(67) = 10.
DATA3(66) = 0.0
DATA3(67) = 1.0

2 DO 4 KK = 1, 165
DATA2(KK) = -TPO(KK)
DATA3(KK) = (FLOAT(KK) - 1.0)*16./256.

4 CONTINUE

CALL LINE(DATA3, DATA2, 65, 1, 0, 0)

IF(JCURV.EQ.JLAST) CALL PLOT(0, 0, 0, 0, 999)
RETURN
END
**EXEC**  
PLOTG,FORM=0111,PARM.GO='SIZE=500K',TIME=5  

*----------------------*
| COMMON              |
| COMMON/getParam/LSAMPL,NSYMB,NO1,NO2,BAUD,SBANDW |
| COMPLEX TF(2048),TF1(2048),TF2(2048),TF3(2048),TF4(2048) |
| COMPLEX DATA(2048),DATA1(2048),BUFF(2048) |
| DIMENSION PEI(25),EBNO(25),IPNI(128),IPNQ(128) |
| DIMENSION IWK(12),PO(65),IPNI1(128),IPNQ1(128) |
| DIMENSION XARRAY(27),YARRAY(27),APEI(25) |
| REAL MP,MP1,MP2,MP0,MPF */A/A |

**INITIALIZE**  

**PROGRAM.**

IOFF=8  
NSYMB=128  
LSAMPL=16  
BAUD=60.  
DATA NSNR/25/  
DATA NRUNS/8/  
DATA DATA ALPHA1,ALPHA2/0.4,0.4/  
DATA DATA ALPHA3,ALPHA4/0.1,0.2/  
DATA DATA BAKOFH,BAKOF/7.0,2.0/  
DATA SBANDW=FLOAT(LSAMPL)*BAUD  
LDIM=LSAMPL*NSYMB  
NO1=LDIM/2+1  
NO2=NO1+1  
DATA PSHIF1,PSHIF2/1.70,3.14/  
DATA DATA ITSH1,ITSH2/672,1344/  

**SET**  
**SYSTEM**  
**PARAMETERS**

A=0.85  
DATA AT1,AT2/-0.1,0.1/  
DATA FOF,FOF1,FOF2/0.,100.,-100./  
DATA DATA NOA/5/  
DATA DATA ICEL/5/  
DATA DATA FBW1,FBW2,FBW3,FBW4/30.0,30.0,40.0,35.0/  

**PRINT**  
1 PRINT(2X,'**SIMULATION OF SQAM MODEM IN ACI **')  
2 WRITE(6,2)  
6 WRITE(6,2)  

1 FORMAT(2X,'** SIMULATION OF SQAM MODEM IN ACI **')  
WRITE BAUD  
6 FORMAT(5X,'SYMBOL RATE:',F7.2,' MBAUD')  
WRITE(6,2)  
2 FORMAT(5X,'RX FILTER CUTOFF FREQ=',F4.1,' MHz')  

- 134 -
START OF COMPUTATIONS.

DO 600
PSH1F1 = 1.7
PSH1F2 = 3.14
ITSH1 = 672
ITSH2 = 1344
DO 19
APEI(I) = 0.0
DO 500
CALL LOAD4(DATA, IPNI, IPNQ)
BUT(TF, FBW1, ICFL, FOFL)
RCOS(FBW1, TF, ALPHA1, FOFL)
HLIM(DATA)
POWER(DATA, MP, PF)
ENERGY(DATA, EB, BAUD)
FILTER(DATA, TF, FOFL)
CALL LOAD4(DATA1, IPN11, IPNQ1)
HLIM(DATA1)
PHASE(DATA1, PSHIF1)
CALL FSM1(DATA1, ITSH1)
ATT(DATA1, AT1)
POWER(DATA1, MP1, PF1)
FILTER(DATA1, TF, FOFL1)
I = 1, LDIM
DATA(I) = DATA(I) + DATA1(I)
CALL LOAD4(DATA1, IPN11, IPNQ1)
HLIM(DATA1)
PHASE(DATA1, PSHIF2)
CALL FSM1(DATA1, ITSH2)
ATT(DATA1, AT2)
POWER(DATA1, MP2, PF2)
FILTER(DATA1, TF, FOFL2)
I = 1, LDIM
DATA(I) = DATA(I) + DATA1(I)
CALL POWER(DATA, MPF, PFF)
HHGG(TF, PNOISE)
DECOD4(DATA, PNOISE, IPNI, IPNQ,
#EBNO, PEI, NSNR, EB)
DO 5
PEI(I) = PEI(I) / FLOAT(NSYM)
EBNO(I) = FLOAT(I)
APEI(I) = APEI(I) + PEI(I)
IF((PEI(I), LT..1E-03.AND.PEI(I), GT..1E-04).AND.(KM, EQ, 1)) KL = I
CONTINUE
PO1 = (PEI(KL)**2.)*PO1

ITSH1 = ITSH1 - 1
PShif1=PShif1+.22
PShif2=PShif2+.46
ITSh2=ITSh2+1

500 CONTINUE
WRITE(6,4) FORMAT(5X,'A=',F3.2)
4 N=1
WRITE(6,3) N,POF1,AT1,AC1
N=2
WRITE(6,3) N,POF2,AT2,AC2
3 FORMAT(///,2X,'CHANNEL ',II,' CHARACTERISTICS',/,,2X,
&'OFFSET FREQ.',',F6.1,' MHZ',', ATTENUTATION',',F4.1,' DB',
&' CARRIER TO INT. RATIO': ',F5.1,' DB')
DO 64 I=1,NSNR
64 APEI(I)=APEI(I)/FLOAT(NRUNS)
CONTINUE
WRITE(6,150) FORMAT(5X,'EB/NO','10X,'PROB. OF. ERROR',1)
WRITE(6,172) (EBNO(I),APEI(I),I=1,NSNR)
172 DO 99 I=1,NSNR
99 XARRAY(I)=EBNO(I)
CALL DRAW(XARRAY,YARRAY,NSNR,KA,NOA)
CALL EYEQQ(DATA,16,128,KA,NOA)
C CHANGE SYSTEM PARAMETERS
AT1=AT1-3.
AT2=AT2-3.
C
600 STOP
END

C******************************************************************************
C THIS SUBROUTINE ATTENUATES THE SIGNAL POWER.
C******************************************************************************

SUBROUTINE ATT(DATAT,AT)
COMMON LDIM
COMMON/LSAMPL,NSYMB,N01,N02,BAUD,SBANDW
COMPLEX DATA(I)
ATT=10.**(-AT/20.)
DO 10 I=1,LDIM
10 DATA(I)=DATA(I)*ATT
RETURN
END
SUBROUTINE TIME(DATA, IT)
COMMON LDIM
COMMON/PARA/LSAMPL, NSYM, NO1, NO2, BAUD, SBANDW
COMPLEX DATA(1), DAT
DO I=1, IT
  DO 20 J=2, LDIM
    DATA(J-1) = DATA(J)
  20 DATA(LDIM) = DAT
10 RETURN
END

SUBROUTINE POWER(DATA, MP, PF)
COMMON LDIM
COMMON/PARA/LSAMPL, NSYM, NO1, NO2, BAUD, SBANDW
COMPLEX DATA(1)
REAL MP
PF=0.
MP=0.0
DO 10 I=1, LDIM
  MP = MP + ((CABS(DATA(I)))**2.)
10 IF(PF .LT. CABS(DATA(I))) THEN
   PF = CABS(DATA(I))
   CONTINUE
PF = PF**2./MP
PF = 10.* ALOG10(PF)
RETURN
END

SUBROUTINE FILTER(SIGNAL, TF, FOF)
COMMON LDIM, IOFF
COMMON/PARA/LSAMPL, NSYM, NO1, NO2, BAUD, SBANDW
COMPLEX SIGNAL(1), TF(1), XX
DIMENSION INK(12)
CALL FFT2C(SIGNAL, 11, INK)
IFOF = IFIX(FOF*FLOAT(LDIM)/SBANDW)
CALL FSMI(SIGNAL, IFOF)
DO 1 I=1,LDIM
   SIGNAL(I)=CONJG(SIGNAL(I)*TP(I))
   FFT2C(SIGNAL,11,1W)
1    I=1,LDIM
RETURN
END

C****************************************************************************************************
C THIS SUBROUTINE COMPENSATES THE PHASE SHIFT DUE TO THE CARRIER OFFSET.
C****************************************************************************************************

SUBROUTINE PHASE(DATA,PSHIFT)
   COMMON LDIM
   COMMON COMPLEX EPS,DATA(1)
   EPS=CMPLX(COS(PSHIFT),-SIN(PSHIFT))
   DO 10 I=1,LDIM
      DATA(I)=DATA(I)*EPS
10    CONTINUE
RETURN
END

C****************************************************************************************************
C THIS SUBROUTINE DECODES THE PROCESSED DATA
C****************************************************************************************************

SUBROUTINE DECOD4(DATA,PNOISE,IPNI,IPNQ,#EBNO,PEI,NSNR,EB)
   COMMON LDIM,IOFF
   COMMON COMPLEX PARA/LSAMPL,NSYMB,NO1,NO2,BAUD,SBANDW
   COMMON DIMENSION EBNO(1),PEI(1),IPNI(1),IPNQ(1)
   DATA NOLD,NOF/0,0/,111
   IF (IOFF.EQ.0) GO TO 111
   DO 9 K=1,IOFF
      XX=AIMAG(DATA(1))
      LD=LDIM-1
      DO 5 KK=1,LD
         DATA(KK)=CMPLX(REAL(DATA(KK)),AIMAG(DATA(KK+1)))
      5    DATA(LDIM)=CMPLX(REAL(DATA(LDIM)),XX)
   9     CONTINUE
   C SYNCHRONIZE THE RECEIVED DATA
   C111   IF(KN.GT.1) GOTO 210
   111   LFU=2*NSYMB
   C NERROR=0
   K=1
   CONTINUE
   C
   NEWM=0
   - 138 -
DO 200 J=1, NSYM
J1=K+(J-1)*LSAMPL
IF(J1.GT.LDIM) J1=J1-LDIM
XB=REAL(DATA(J1))
YB=AIMAG(DATA(J1))
IF (SIGN(1.,XB).EQ.IPNI(J)) NEW=NEW+1
IF (SIGN(1.,YB).EQ.IPNQ(J)) NEW=NEW+1
CONTINUE
GOTO 12
IF(NOLD.GE.NEW) GOTO 12
NOLD=NEW
NOF=K
K=K+1
GOTO 300
IF(NOLD.LT.LFU.AND.K.LE.LDIM) GOTO 300
NOS=NOF-1
ND=NOS
CALL FSM1(DATA,ND)
C 230
IF(KN.GT.1) GOTO 110
MI=1

MQ=1
EOI=FLOAT(NSYMB)+1.
EOQ=FLOAT(NSYMB)+1.
SNR=PNOISE*EB/10.
SIGMA=SQR(SNR)
DO 110 J=1, LSAMPL
EI=0.
EQ=0.
DO 100 K=1, NSYMB
J1=(K-1)*LSAMPL+J
AXBAR=(REAL(DATA(J1))+REAL(DATA(J1+1)))/2.
AYBAR=(AIMAG(DATA(J1))+AIMAG(DATA(J1+1)))/2.
IF(SIG(1.,AXBAR).NE.IPNI(K)) EI=EI+1
IF(SIG(1.,AYBAR).NE.IPNQ(K)) EQ=EQ+1
ARGI=ABS(AXBAR)/SIGMA
ARGQ=ABS(AYBAR)/SIGMA
EI=EI+ERFC(ARGI)/2.
EQ=EQ+ERFC(ARGQ)/2.
100 IF (EOI.LE.EI) GO TO 120
EOI=EI
MI=J
120 IF (EOQ.LE.EQ) GO TO 110
EOQ=EQ
MQ=J
110 CONTINUE
MOFF=IABS(MI-MQ)
MOFF=(MQ+MI)/2
OFF=(FLOAT(MOFF)+FLOAT(NOFS)-8.0)/FLOAT(LSAMPL)
DJPP=((8.0-FLOAT(MOFF))/FLOAT(LSAMPL))*2
DO 11 M=1, NSNR
PEI(M)=0.0
K=1, NSYMB
XI
DO 2 J1=(K-1)*LSAMPL+MI
JQ=(K-1)*LSAMPL+MQ
CONTINUE
AXBAR = (REAL(DATA(JI))+REAL(DATA(JI+1)))/2.
AYBAR = (AIMAG(DATA(JQ))+AIMAG(DATA(JQ+1)))/2.
INDEXX = 0
IF(SIGN(1.,AXBAR).NE.IPNI(K)) INDEXX = 1
INDEXY = 0
IF(SIGN(1.,AYBAR).NE.IPQ(K)) INDEXY = 1
IF((INDEXX.EQ.1).OR.(INDEXY.EQ.1)) NERROR = NERROR +1

C COMPUTE THE PROBABILITY OF ERROR FOR THIS SYMBOL
DO 4 M = 1, NSNR
X = FLOAT(M)/10.
SNR = PNOISE*EB/(10.**XM)
SIGMA = SQRNT(SNR)
ARG = (ABS(AXBAR))/(SIGMA*SQR(2.))
IF (ARG.GT.12.) ARG = 12.
EX = ERFC(ARG)/2.
IF (INDEXX.EQ.1) EX = 1.-EX
ARG = (ABS(AYBAR))/(SIGMA*SQR(2.))
IF (ARG.GT.12.) ARG = 12.
EY = ERFC(ARG)/2.
IF (INDEXY.EQ.1) EY = 1.-EY
IF (EX.LT.1.E-15) EX = 0.
IF (EY.LT.1.E-15) EY = 0.
PEI(M) = PEI(M) + ((EX+EY)/2.)
CONTINUE
4 CONTINUE
WRITE (6,152) NERROR
C152 FORMAT(/5X,'ERRORS=',I5,/) RETURN
END

C*************************************************************************
C APERTURE EQUALIZED RAISED COSINE FILTER WITH ARBITRARY ALPHA
C*************************************************************************

SUBROUTINE RCOS(FBANDW,TF,ALPHA,FOF)
COMMON LDIM,IOFF
COMMON PARA,LSAMPL,NSYMB,NO1,NO2,BAUD,SBANDW
COMPLEX FF(X) IS THE INVERSE BASEBAND SPECTRUM

FF(X) = 2.*X*(1.-((4.*FBANDW**2)/(3.141592**2))*(X**2))/SIN(2.*X)
QQ = 2.

IF (ALPHA.EQ.0)
FN = LDIM*(FBANDW/2.)
F1 = (1.-ALPHA)*FN
F2 = (1.+ALPHA)*FN
IFN = IFIX(FN)
IF1 = IFIX(F1)+1
IF2 = IFIX(F2)+1
A1 = 3.141592/(2.*FLOAT(IFN))
TF(1) = CMPLX(1.0,0.0)

C*************************************************************************
DO 8 I=2,IF1
   J=I-1
   A2=FF(FLOAT(J)*A1)
   TF(I)=CMPLX(A2,0.0)
8     CONTINUE
   JK=IF1+1
   DO 9 J=JK,IF2
      I=J-1
      A3=QQ
      IF(I.NE.FN) A3=FF(FLOAT(I)*A1)
      A=(3.141592/(2.0*ALPHA))*((FLOAT(I)/FLOAT(IFN))-1.)
      TF(J)=CMPLX(.5*(1.0-SIN(A)),0.0)
      TF(J)=TF(J)*CMPLX(A3,0.0)
9     CONTINUE
   JH=IF2+1
   DO 10 J=JH,NO1
      TF(I)=CMPLX(0.0,0.0)
10    CONTINUE
   I=NO2,LDIM
   DO 5 I=NO2,LDIM
      TF(I)=TF(LDIM+2-I)
      CONTINUE
      IFOF=IFIX(FOF*FLOAT(LDIM)/SBANDW)
      CALL FSM1(TF,IFOF)
      RETURN
   END

***********************************************************************
C PHASE EQUALIZED BUTTERWORTH FILTER
***********************************************************************
SUBROUTINE BUT(TF,FB,ICEL,FOF)
COMMON LDIM,IOFF
COMMON/PARA/LSAMPL,NSYMB,NO1,NO2,BAUD,SBANDW
COMPLEX A1=1./NSYMB
      FBNOR=FB/BAUD
      TF(1)=CMPLX(1.0,0.0)
      DO 10 I=2,NO1
         J=I-1
         A2=1./SQRT(1.+(A1*FLOAT(J)/FBNOR)**(2*ICEL))
10      TF(I)=CMPLX(A2,0.)
      DO 20 I=NO2,LDIM
         TF(I)=TF(LDIM+2-I)
20      IFOF=IFIX(FOF*FLOAT(LDIM)/SBANDW)
      CALL FSM1(TF,IFOF)
      RETURN
   END
SUBROUTINE
COMMON
COMMON/ PARA/ LSAML, NSYMB, NO1, NO2, BAUD, SBANDW
COMPLEX
J=LDIM-JABS(IOF)
K=J+1
IF (IOF.EQ.0) GOTO 9
IF (IOF.LT.0) GOTO 4
DO 1 I=1, J
1 DO 2 I=K, LDIM
2 DO 3 I=1, J
3 DATA(I-J)=DATA(I)
GOTO 9
4 DATA(I+IOF)=BUFF(I)
5 IOF=-IOF
6 I=L, LDIM
7 DATA(I+J)=DATA(I)
8 DATA(I)=BUFF(I)
9 RETURN
REFERENCES


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