FACULTÉ DES SCIENCES ET DE GÉNIE
FACULTY OF SCIENCE AND ENGINEERING

REPORT ON
REGENERATIVE TRANSPONDERS FOR MORE
EFFICIENT DIGITAL SATELLITE SYSTEMS:
PHASE I

BY: Dr. Kamilo FEHER
Principal Investigator
Department of Electrical Engineering
University of Ottawa
OTTAWA

FOR: DEPARTMENT OF COMMUNICATIONS
OTTAWA

UNIVERSITÉ D'OTTAWA
UNIVERSITY OF OTTAWA

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PHASE I

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REGENERATIVE TRANSPONDERS FOR MORE EFFICIENT DIGITAL SATELLITE SYSTEMS

Summary

The demand for more cost-competitive satellite capacity is increasing and recent advances in technology has encouraged an increased use of digital communication satellites. Means are being sought to utilize the available power and bandwidth more efficiently. The use of regenerative transponders in the satellite is one way that leads to the efficient use of available resources (power and bandwidth). In this report, the advantages of regenerative satellite systems are explored. Our preliminary results based on computer simulations indicate considerable potential savings in power leading to possibly more cost-competitive satellite systems. This is because the signal detection on the regenerative transponder prevents the passing on of uplink noise and interference to the downlink. Regeneration allows capabilities beyond those possible with today's non-regenerative satellites. These include interconnectivity between terminals of different types.

Conventional satellite systems are studied and the most frequently considered modulation methods described. Degradations which occur in a conventional channel (usually nonlinear) when signals are bandlimited are discussed. The regenerative system appears to have the greatest gain in situations where the conventional system has the greatest degradation.

The literature available on this subject confirm the advantages of regenerative satellites which were found during our work but a critical review of it reveals that most of the studies were somewhat pessimistic i.e. they indicated a lower gain than we expect because they considered only a simplified channel, for instance without AM/PM conversions. It appears that in nonlinear channels a reduction in the energy per-bit-to-noise density ratio of up to 5 dB in the uplink and up to 6 dB in the downlink is possible if regenerative satellites replace the conventional translating ones. This can possibly mean the use of smaller earth station antennas hence a saving in the overall cost of the system.

From our recent visit to the Comsat laboratories in the USA, we learned that a lot of research effort is already being devoted to this area of satellite communications both in the USA and abroad. We believe that this study is of benefit to Canadian telecommunications industry if they are going to keep at the forefront of developments in telecommunications, and through discussions, we have already started to create interest in regenerative satellite systems in Canadian Industry.

This most interesting study was suggested by Dr. Gerald Lo of the Communications Research Centre, Ottawa, and was carried out in close consultation with him.

1.0 INTRODUCTION

Satellite communications has proved to be a viable method of worldwide communication reaching even the remotest of places at relatively low cost.

A basic communication satellite system is shown in Figure 1.1 below. The satellite itself acts as a repeater in the sky. Earlier satellites used to be simply reflectors, but today the satellite could be very complex. The next few years will see an increased demand for more cost-competitive capacity. As this demand continues to soar, larger capacity satellites will be called for, as well as an increase in the actual number of satellites. Also, as the circuit technology continues to advance rapidly, more and more digital communication satellites will be used. This will allow techniques such as Time Division Multiple Access (TDMA) to be used. Digital transmission techniques in conjuction with TDMA have been considered a means of achieving efficient, high-capacity, flexible communications. The efficiency of TDMA is due in part to the single-carrier-per-transponder operation, which reduces the effect of the Travelling Wave Tube Amplifer (TWTA) nonlinearities by avoiding intermodulation products [1]. [However AM/AM and AM/PM nonlinearities still cause degradation in this mode of operation.]

The use of digital communication satellites and the increase in capacity and the number of satellites will necessitate several considerations.

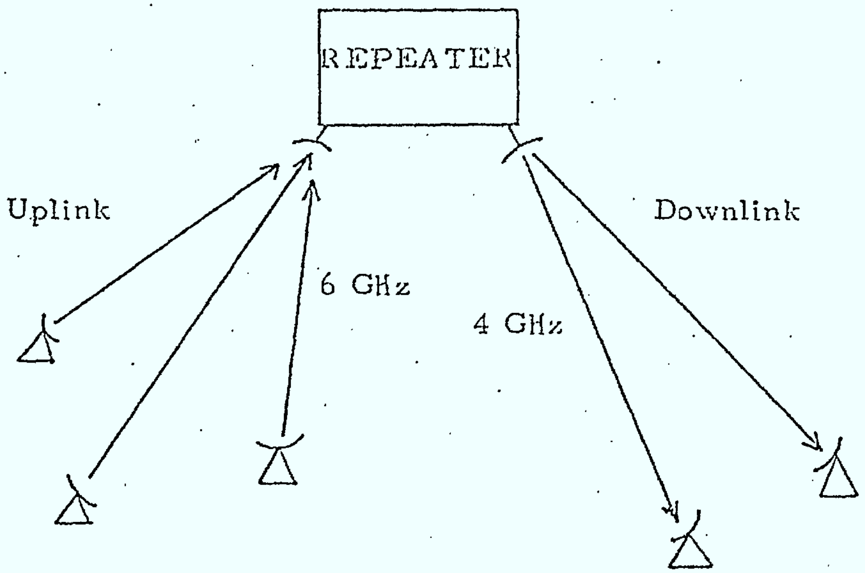

Figure 1.1 A Basic Communication Satellite System

With an increased number of users, cheaper (smaller) ground station antennas will be demanded, hence effectively the link power will be curtailed. Also, increased use of the spectrum will require more efficient modulation methods (such as 8-PSK and 16 QAM) to be used. This in turn will aggravate the problem of link power. Spectrum conservation methods will be called for. These include frequency reuse and multiple spot-beam antennas [2]. For example Intelsat V will have two spot beams. When receiving and transmitting simultaneously on many spot beams at the same carrier frequency, co-channel interference (CCI) will become an important performance limiting factor [3].

To solve this problem one may go higher up in the frequency scale say above 100 GHz where more bandwidth will be available. However, this is, of today, technology limited and also no government regulations exist for the use of such frequencies. The other solution would be to discover new modulation techniques which are power efficient and also insensitive to co-channel interference and noise. Needless to say, this is not a trivial task. A more realistic solution is to use signal processing satellite repeaters, for example *regenerative repeaters* instead of the conventional translating repeaters mostly in use these days. This (signal processing) repeater will be capable of demodulating the uplink signals into baseband data and remodulating for retransmission in the downlink. In terrestrial microwave digital systems, regeneration is usually considered advantageous when a large number of repeaters exist. If only a few repeaters are in use, it becomes more desirable to slightly increase the transmitting power in order to maintain the desired probability of error [4]. However, in a satellite system (with only one repeater) such an increase in power is far from easy due to problems of weight etc. On-board regeneration then becomes a good way to achieve the desired performance. It allows capabilities beyond those achievable with simple translating satellites. These include considerable interference protection and interconnectivity between different types of terminals [5]. Figure 1.2 indicates the type of gain achievable with regenerative repeaters. *A reduction in* E_b/N_0 *of 4-5 dB in the uplink and 4 to 6 dB in the downlink is possible.* The significance of this figure will explained more fully in Chapter 6.

This report studies the relative merits of regenerative repeaters over the conventional translating repeaters and looks briefly at the impact of implementing regenerative transponders on satellite systems.

Chapter 2 gives the system descriptions for both conventional and regenerative systems. The concepts of AM/AM and AM/PM are briefly presented and the modelling of a memoryless nonlinear device such as a TWTA is discussed. The equivalent baseband representation is also discussed. The possibility of using analytical methods is explored.

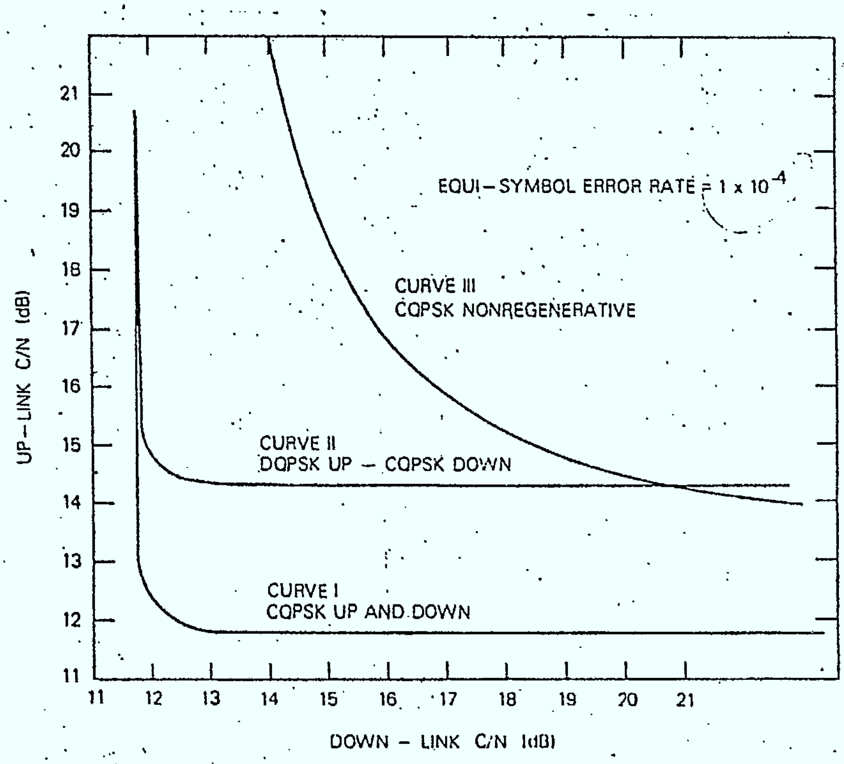

Fig. I.2 *Equi-symbol Error Rate Curves of 4-phase PSK Systems Using a Regenerative Repeater vs a Conventional Transponder*

Chapter 3 gives a more complete study of the non regenerative systems including the concept and effects of cascaded nonlinearities. A computer simulation programme developed to study a cascaded nonlinearity system is described and the simulation results are discussed. This programme was developed by Dr. Huang at Concordia University under the supervision of Dr. Feher [6]. The results obtained for conventional channel are given, as a starting point, for QPSK, OKQPSK and MSK. To be applicable to the regenerative satellite case, the programme was modified at the University of Ottawa and is being carefully scrutinised to assure reliable results.

Chapter 4 discusses various system configurations considered possible for regenerative satellite systems. Only the results of a system with coherent quadrature phase shift keying (CQPSK) on both the uplink and downlink are presented due to time constraints. *It is understood that these results are preliminary, final results and results for other system configurations will follow in the report for Phase II of the expected contract.*

Chapter 5 considers other modulation methods possible for regenerative and nonregenerative systems, notably partial response.

In Chapter 6, a comparison of regenerative and conventional systems is presented based on the results obtained from our simulations and also on literature survey. Finally recommendations, further research and considerations are included in Chapter 7.

It should be noted that this is the *first phase* of the project and a lot of time was put into adapting the programmes to the regenerative satellite systems and also into rechecking the conventional system results. Time constraints and budgetary considerations ($12,000 for this phase) have to be taken into account. The possible effects on the results, of parameter variations such as the number of samples per bit were therefore not fully considered. This reduces somewhat the confidence of the results included in this report and makes them subject for further verification which is expected to be done in Phase II of the contract.

Although the benefits of regeneration are usually considered in conjunction with TDMA, this is not treated in this initial phase. The effect of intersymbol interference (ISI) on the system was not isolated. [It should be noted however that ISI was considered because filters were used, but we did not quantify it in terms of dB's and state that so many dB'S of ISI will cause so many dB's of degration in the system E_b/N_o for a given bit error rate]. Other considerations of a system include the effect of synchronisation errors and the effect of adjacent channel interference (ACI). These have not been included in the programmes at this state. Also, the simulation relied heavily on the equivalent baseband representation of a bandpass process. This presumes symmetry of the bandpass filters etc. The case of unsymmetrical channels would be more complex to simulate.

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CHAPTER 2

COMMUNICATIONS SATELLITE SYSTEMS DESCRIPTIONS

2.1 CONVENTIONAL (NON REGENERATIVE) SATELLITE SYSTEMS

Past and some current satellite communication systems are of the analogue type with the prevalent multiplexing, modulation and multiple access techniques being FDM/FM/FDMA. Recently, however, digital transmission links using single channel per carrier (SCPC)/PSK/FDMA and PCM/PSK/TDMA are becoming more common. These systems have all been using the conventional or non regenerative mode of operation. In this mode, the satellite acts merely as a one-hop frequency translating relay repeater: the uplink signal has its carrier shifted in frequency and then amplified before being relayed on the downlink. No other processing is done on board the satellite. The major factors that inhibit the use of on board processing are weight and size considerations due to the launching operation, the power consumption limitations, and hardware reliability.

Herein, the constraints imposed on a typical non regenerative satellite system and the effects of these constraints on various types of digitally modulated signals are studied. Attention will be focussed fully on digital modulation because this is where recent emphasis is placed. Furthermore, it is only for digital satellites that on-board signal processing (baseband signal regeneration) is of practical interest.

Fig. 2.1 shows a typical non-regenerative satellite channel model. The satellite link considered consists of the transmitting earth station, the satellite transponder and the receiving earth station. At the earth transmitting station, the transmit bandpass filter (F_1), used to bandlimit the spectrum, causes intersymbol interference (ISI); the high power amplifier (HPA), operated near saturation, causes both the AM/AM and AM/PM conversions of the signal

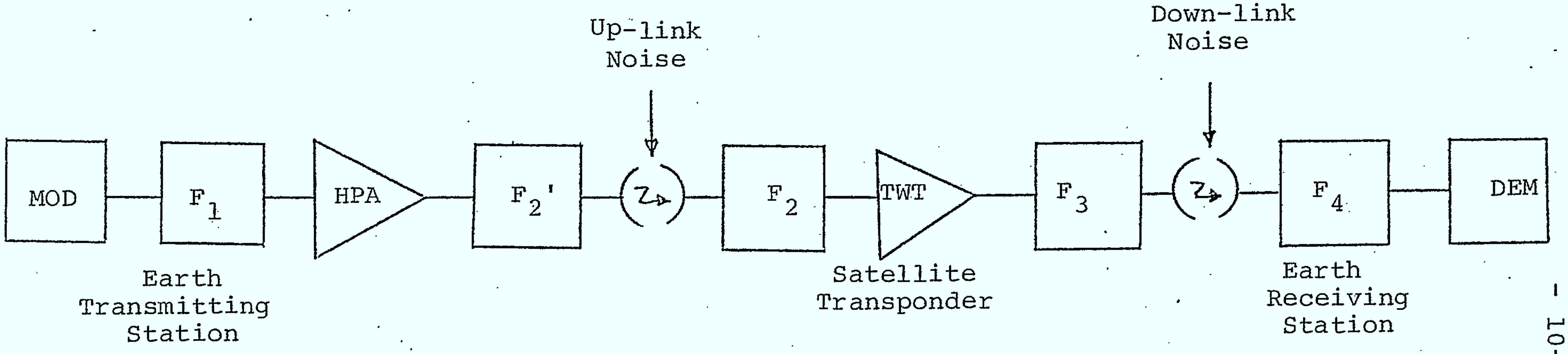

Fig. 2.1 Conventional Digital Satellite Communication System Model

and degrades the system performance. The satellite input and output MUX filters (F_2 and F_3) which are used to bandlimit the signal and reduce the spectrum spreading caused by the TWT, also induce ISI. Due to the power constraint on board the satellite transponder, the TWT is usually operated near saturation and causes AM/AM and AM/PM conversion degradation. At the receiving earth station, the receive filter (F_4) , which is used to bandlimit the thermal noise and reduce the adjacent channel interference, also degrades the system performance. For computer simulation purposes, the filter F_2' is usually lumped with on board filter F_2.

The HPA at the earth station and the TWT at the satellite transponder are operated close to saturation in order that maximum power conversion efficiency can be achieved. The DC to RF power conversion efficiency of these nonlinear devices is the highest when they are operated near saturation. In this mode of operation, these two devices exhibit the nonlinear effects of AM/AM and AM/PM conversion [1]. The AM/AM conversion causes spectral spreading of the filtered signal which will then interfere with signals in the neighbouring channels (adjacent channel interference). In the case of M-ary PSK (Phase Shift Keying) modulated signals, substantial AM/PM conversion will also cause the system performance to deteriorate.

Because of the ever increasing demand in communication traffic, the channel available for the satellite communication system is usually narrowband. For example, the Intelsat V system will be operating at 60 Mb/s using only 36 MHz of bandwidth, i.e. 1.67 b/s/Hz efficiency. To avoid the adjacent channel interference, the channel allocation is 40 MHz [2]. Tight filtering is therefore required so as to reduce the adjacent channel interference while maintaining manageable intersymbol interference so that optimum performance can still be achieved.

Digital modulation techniques suitable for the nonlinear satellite channel therefore, must meet the following two requirements.

1. Bandwidth efficient with optimal bandlimiting filters for minimal intersymbol and adjacent channel interference.

2. Optimal system gains from the HPA and TWTA with little spectral spreading and distortion degradation from their AM/AM and AM/PM conversions.

Based on these two constraints, three modulation techniques, namely, conventional QPSK, OKQPSK and MSK, have frequently been considered for such a nonlinear satellite channel [2,3,4]. The basic schemes for these three modulation techniques will be reviewed in Chapter 3.

To study and compare the performance of these three signals in a bandlimited nonlinear satellite channel, the first part of Chapter 3 will be devoted to analyzing the envelope fluctuations of these signals which arise when they are bandlimited. Understanding of the envelope fluctuations is essential in the study of the spectrum spreading which occurs when these filtered signals pass through a nonlinear device.

Two types of envelope fluctuations are considered. The first type is the overall envelope fluctuation of the filtered signal. The second type is the envelope fluctuation at the sampling instants. The latter type of envelope fluctuation is more important as the performance of the system is determined by what occurs to the signal at the sampling instants.

Next, spectrum spreading which occurs when a filtered PSK signal passes through a power amplifier will be studied. Two types of spectral spreading are experimentally demonstrated. The first type, defined as complete spectrum restoration, occurs when a jitter-free signal is hardlimited. The second type, defined as spectrum restoration with modification, is partly attributed to the timing jitter of the filtered signal.

To aid in the study of the overall system performance of these three signals through a typical non-regenerative satellite link, a computer sinulation program has been developed. Results obtained from this computer simulation will be analyzed and compared with those available in the literature.

2.2 REGENERATIVE SATELLITE SYSTEMS

The continuing growth of the demand for larger-capacity satellites imposes the requirement that satellite systems be used more efficiently. This, coupled with the need for cheaper earth terminals and the more complex mix of user types, calls for technological improvements to support a complex system structure. On-board satellite processing has been proposed as one means of increasing the satellite system capacity and providing more flexibility. Whereas conventional satellite repeaters are essentially frequency translaters and amplifiers, the signal processing repeaters carry out some other functions in addition to these.

There are three general categories of on-board processors: Radio Frequency Processors (RFP), Channel Symbol Stream Processors (CSSP), and Information Bit Stream Processors (IBSP) [5]. The radio frequency processor does all the processing at RF with no translations to lower frequencies. Such type of processors include a microwave switch matrix having say n inputs and m outputs where any input may be connected to any output. This may be used in a satellite system employing multiple spot beam antennas. The satellite, in this form, does not differentiate the type of modulation or the mode of access.

The channel symbol stream processor accepts the uplink RF signal, translates it to IF, demodulates the digital stream, and then uses this symbol stream to remodulate the downlink carrier. The remodulation is done on a symbol-by-symbol basis and detection into information bits is not carried out. This type of processor generally does not have any storage capacity. Unlike the RFP the link is not transparent; the uplink and downlink are uncoupled. Degradations on the uplink manifest themselves as quantisation errors or hard decision errors. The degradations are not passed on to the downlink but these errors are.

The information bit stream processor demodulated the incoming RF signal detect the bits does any required decoding and then encodes the baseband bit stream which is used to remodulate the downlink RF carrier. The processor may use the decoded bit stream to route the messages to any desired destination. This type of processor is clearly the most complex. Both the CSSP and IBSP are termed "regenerative" processors. In this report, regenerative repeaters shall be assumed to be of the IBSP type. A simplified block diagram of a regenerative satellite system is shown in Figure 2.2.

It is clear that all these processing transponders include a "switchboard in the sky" concept wherein different transponder input channels are switched by ground command to the appropriate downlink channel.

In conventional repeater satellite systems, the uplink performance is not independent of the downlink performance. Degrading effects occuring on the uplink path are translated to the downlink. On-board regeneration reduces greatly this uplink and downlink dependence. Thus regeneration prevents the accumulation

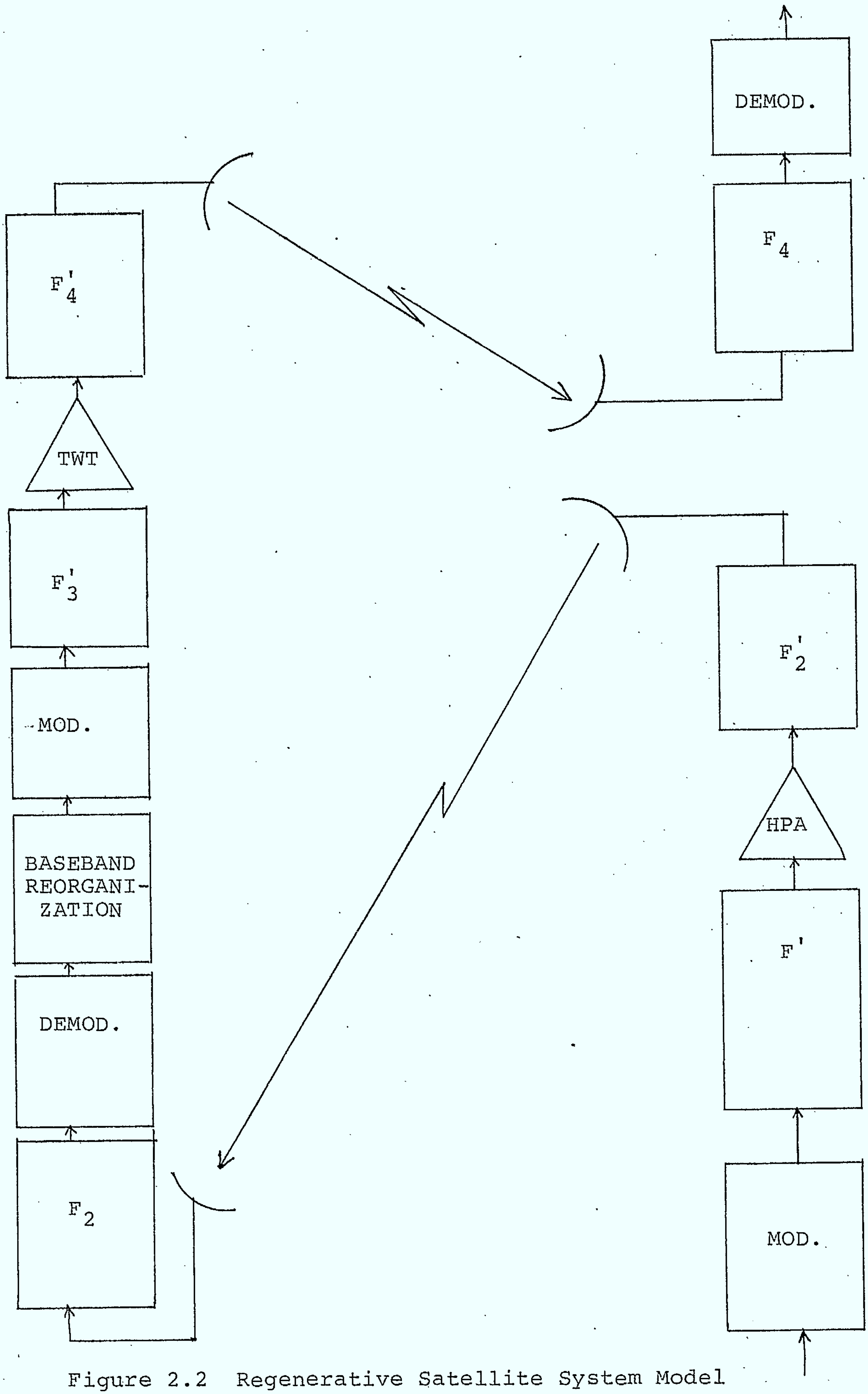

Figure 2.2 Regenerative Satellite System Model

of noise and co-channel interference [6]. To show the gain of regenerative repeaters, assume that in a given satellite system, the uplink energy per bit-to-noise density ratio is $(E_b/N_0)_u$ and the downlink energy per bit-to-noise density ratio is $(E_b/N_0)_d$. In a conventional transponder, the uplink and downlink noise will add up to result in an overall noise level. The total E_b/N_o will be given by

$$\left(\frac{E_b}{N_0}\right)_{total} = \frac{1}{\left(\frac{N_0}{E_b}\right)_u + \left(\frac{N_0}{E_b}\right)_d}$$

Note that in this equation, the quantities are ratios and not dB. The system probability of error will be a function of $(E_b/N_0)_{total}$, the function depending on the modulation method used. Let the function be as shown in Figure 2.3. Let the probability of error corresponding to $(E_b/N_0)_u$ be $P_u(e)$. If one assumes identical E_b/N_0 on up and down links, then

$$\left(\frac{E_b}{N_0}\right)_{total} = \left(\frac{E_b}{N_0}\right)_u - 3 \qquad \text{(quantities in dB).}$$

This corresponds to an error rate $P_t(e)$ shown in Figure 2.3 which is over three orders of magnitude greater than $P_u(e)$.

If a regenerative repeater is used instead, there are essentially two cascaded independent digital communication links. If $P_d(e)$ is associated with $(E_b/N_0)_d$, then the total probability of error on the satellite system is

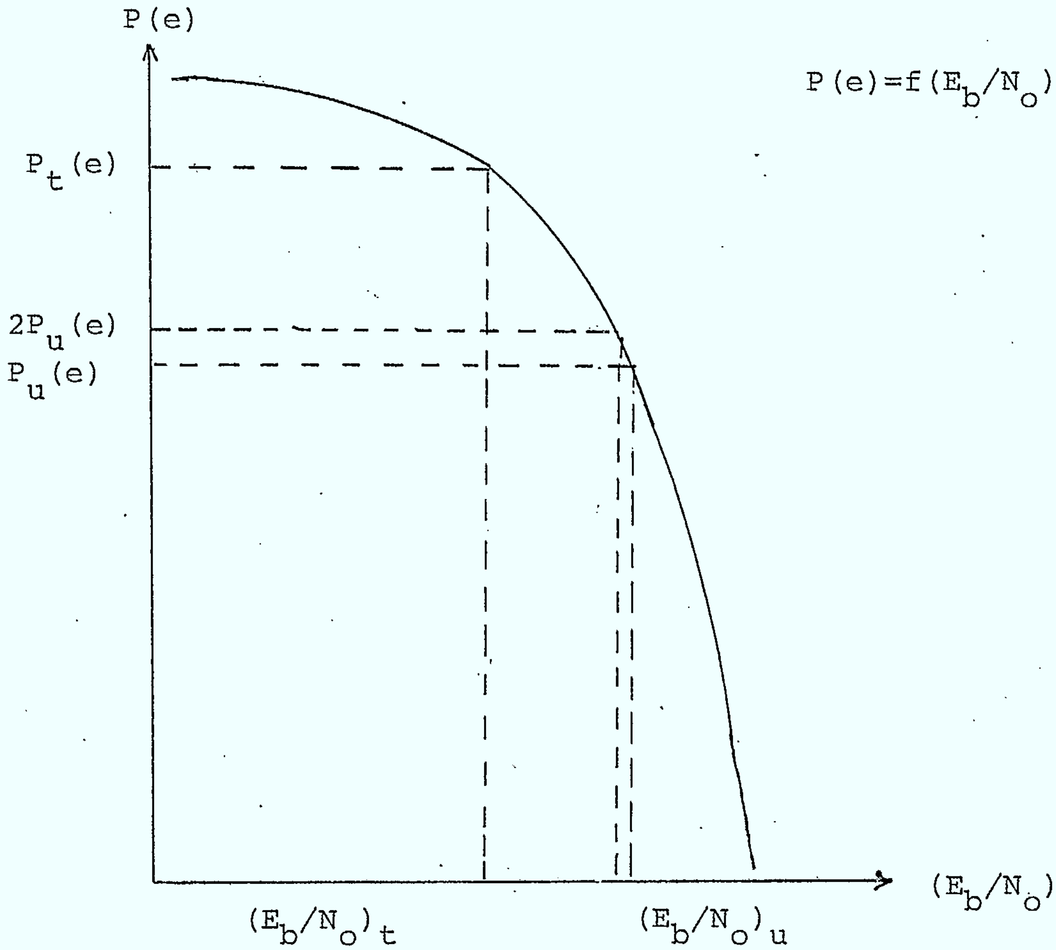

Figure 2.3 - Illustration of Regenerative Satellite Gain

$$P(e) = (1-P_d(e))P_u(e) + (1-P_u(e))P_d(e)$$

$$P_t(e) \doteq P_u(e) + P_d(e) - P_d(e)\,P_u(e) \simeq P_u(e) + P_d(e)$$

Hence if $(\frac{E_b}{N_0})_u = (\frac{E_b}{N_0})_d$, the bit error rate becomes

$$P_t(e) \doteq 2P_u(e)$$

It is clear from Figure 2.3 that in links with identical uplink and downlink E_b/N_0, there is an almost 2.6 dB gain in using a regenerative repeater. A similar gain is expected with respect to interference in the uplink.

The above applies to a linear transponder (linear amplifier or a TWT with a backoff greater than about 14 dB). In nonlinear cases, it is known that the conventional transponder suffers a considerable degradation in performance relative to the modem back-to-back performance. In such a case, it is expected that regenerative system will offer more than 3 dB gain in E_b/N_0 over the non-regenerative system. It is a part of this project to establish the magnitude of this gain. The suitability of different modulation methods shall also be considered.

2.3 NONLINEAR EFFECTS IN CONVENTIONAL AND REGENERATIVE SATELLITE LINKS

2.3.1 Introduction

The analysis of a digital satellite communication links is indeed a difficult problem. The difficulties stem from the fact that the radio frequency interference (RFI) which has a complicated probability density function (pdf) becomes even more complex after processing by the satellite which has a travelling wave tube (TWT) usually operated in a nonlinear mode. The high power amplifier (HPA) in the ground station and the TWT in the satellite need not be operated in a nonlinear mode, however, their nonlinear operation leads to a more efficient amplification of power. Thus, the usual operation of the TWT in a nonlinear fashion leads to a considerable saving in weight and hence a reduced expenditure at launch [7,8].

Although of a somewhat difficult nature, the analysis of the degradations due to the TWT nonlinearities may be undertaken with relative ease since the TWT characteristics are well known [9,10,11]. This analysis can yield an insight into the degradations due to AM/AM and/or to AM/PM conversion effects of the nonlinearity [12]. Unfortunately, the analysis quickly becomes very complicated and its mathematical tractability is lost. One must therefore resort to numerical methods as pointed out by Huang et al. [12] and demonstrated by Lindsey et al. [13].

Yet another scheme available to study degradation effects due to the TWT nonlinearity consists of a complete satellite link simulation [14,15]. This approach is used more frequently since it avoids mathematical tractability problems and affords the system designer an increased degree of flexibility. That is, with a complete system simulation it becomes quite simple to study the effects introduced by modifying any particular component of the system. In particular, a performance study comparing operation at various bit rates, with different amounts of input backoff, with different filters and various modulation schemes, can be done as demonstrated by the McMaster Research Group [14] and by Huang and Feher [16].

A fundamental necessary requirement to perform a system simulation is to have an adequate method of representing the system parameters, in particular, the filters and the nonlinear amplifiers. Since the analysis of a bandpass process can be done equivalently at baseband [17], it is therefore necessary to have the equivalent baseband representation of filters and nonlinear devices to perform a baseband simulation. The equivalent baseband representations of a bandpass nonlinearity is derived in this section. The models derived easily lend themselves to computer simulations where the signals considered are given by their complex equivalent baseband representation and have been used in [14] and [15].

Having the appropriate background, it will then be possible to consider the whole satellite link system and have a better appreciation for its performance and perhaps consider system design improvements. Of particular interest is the regenerative satellite repeaters a comparative performance evaluation between them and their counterparts non-regenerative satellite repeaters. Experimental and simulation results have shown that under certain operating conditions, regenerative satellites can achieve a probability of error similar to non-regenerative satellites with as much a 3 dB less in S/N [6,12,18].

2.3.2 AM/AM and AM/PM Conversion (TWT)

Two effects result from the passage of an amplitude and phase time varying signal through a nonlinear device such as a TWT; they are AM/AM and AM/PM conversion. These are the output amplitude and phase dependence on the input amplitude.

The TWT transponder is generally described by two characteristics; one expresses the input/output phase shift versus input power [9]. Typical characteristics of a TWT are shown in Figure 2.4. For analytical purposes however, the curves illustrated in Figure 2.4 are not very convenient. To circumvent this difficulty, a quadrature model was introduced by Eric [9] and Kaye et al. [19]. This model which is shown in Figure 2.5 illustrates how a memory less bandpass nonlinearity can be represented equivalently as the sum of two nonlinear devices exhibiting only AM/AM conversion and operating in quadrature on the input signal. This particular model is convenient for the analysis of signals when the carrier is present [9]. However, for computer simulation studies, it is preferable to have an equivalent baseband representation of a nonlinear device.

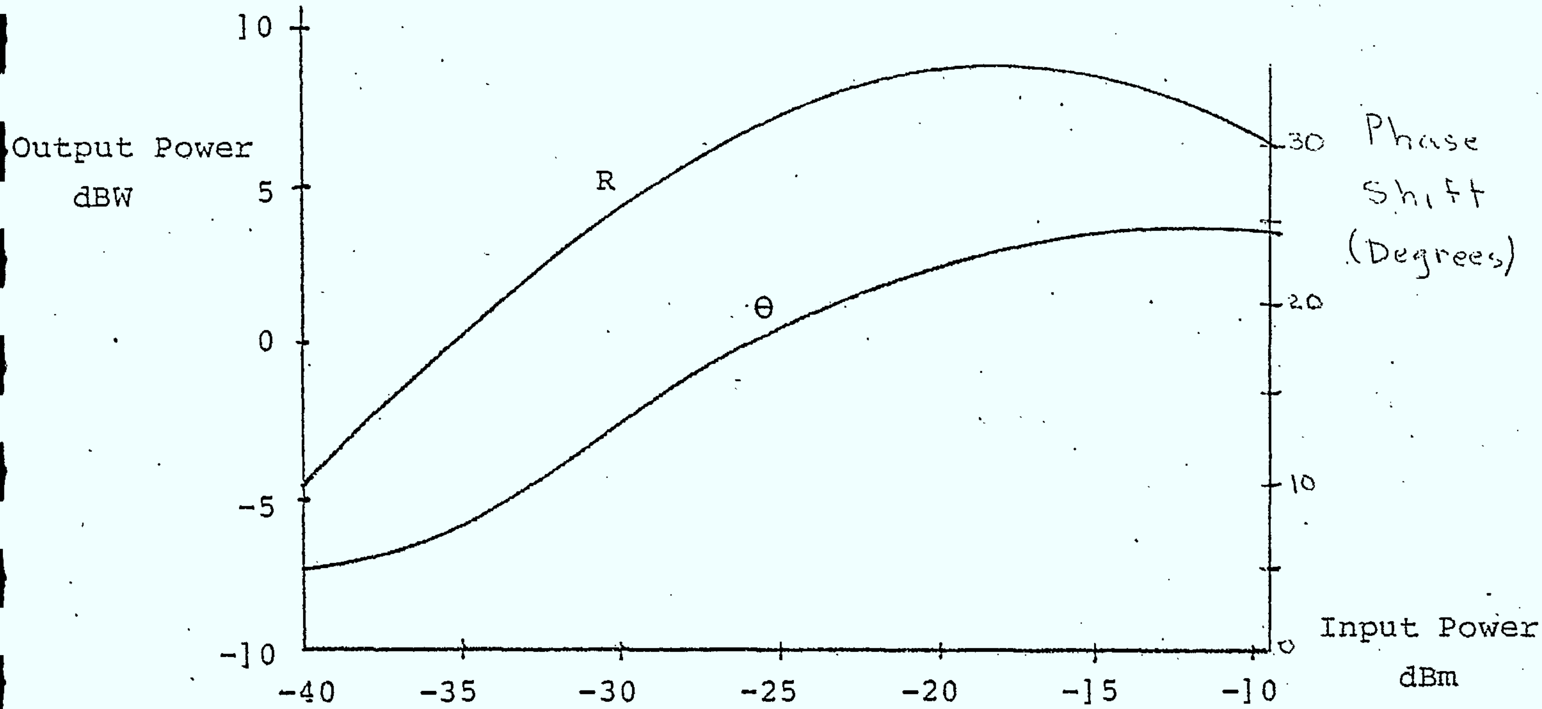

Fig. 2.4 Typical TWT characteristic.

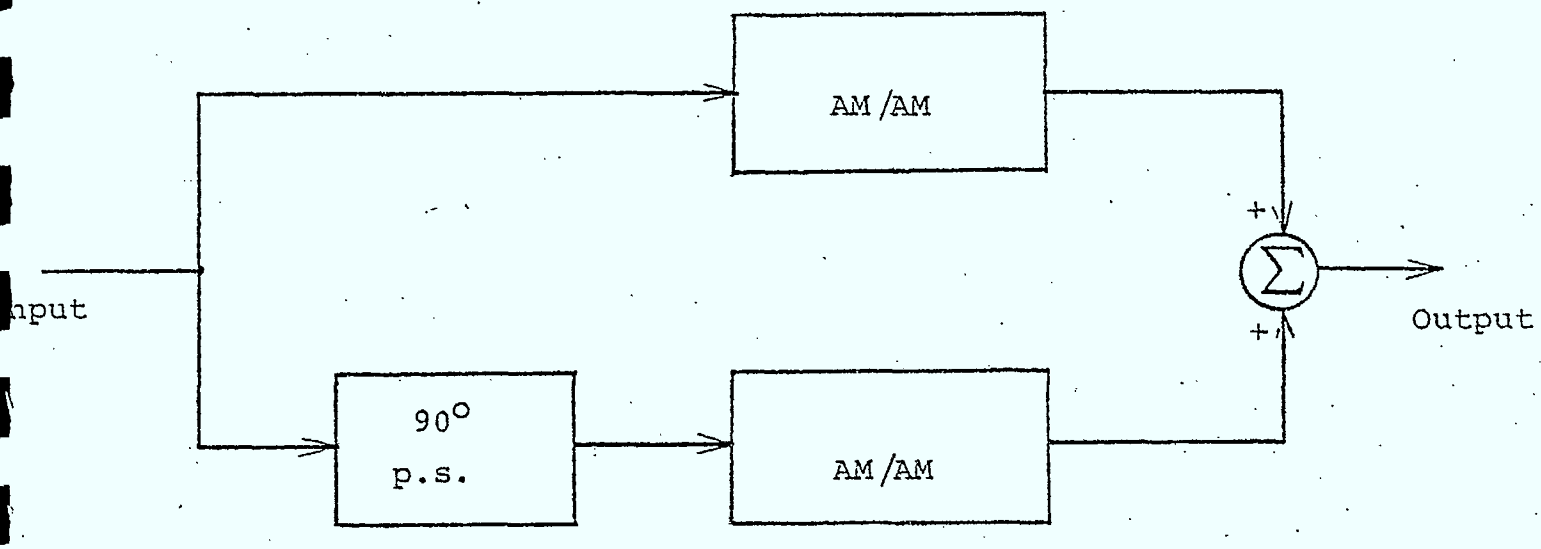

Fig. 2.5 Quadrature model of a memoryless nonlinear device.

Following the analysis given by Chan et al. [15], we now proceed to derive the equivalent baseband representation of a nonlinear device.

Let the input signal to a TWT transponder be represented by

$$V_i(t) = r(t) \cos (w_c t + \phi)$$

$$= \mathrm{Re}\{[x(t) + j\, y(t)] \exp(j\, w_c t)\} \qquad (2.1)$$

where $r(t)$ and ϕ are the envelope and phase functions respectively with $x(t)$ and $y(t)$ the real and imaginary parts of the complex baseband

$$u(t) = x(t) + j\, y(t) \qquad (2.2)$$

where

$$x(t) = r(t) \cos \phi \qquad (2.3)$$

$$y(t) = r(t) \sin \phi$$

The output signal from the TWT transponder is given by

$$V_o(t) = R[r(t)] \cos \{w_c t + \theta[r(t)] + \phi\} \qquad (2.4)$$

where

$R[r(t)]$ represents AM/AM conversion

$\theta[r(t)]$ represents AM/PM conversion

Expanding (2.4), we obtain

$$V_o(t) = R[r(t)] \{\cos \theta[r(t)] \cos \phi - \sin \theta[r(t)] \sin \phi\} \cdot \cos w_c t - R[r(t)]\{\sin \theta[r(t)] \cos \phi + \cos \theta[r(t)] \sin \phi\} \sin w_c t \qquad (2.5)$$

From (2.5), we obtain the complex baseband signal

$$V_o^*(t) = R[r(t)] \{\cos \theta[r(t)] \cos \phi - \sin \theta[r(t)] \sin \phi\} + j\, R[r(t)] \{\sin \theta[r(t)] \cos \phi + \cos \theta[r(t)] \sin \phi\} \qquad (2.6)$$

Multiplying and dividing by r(t), we can write (2.6) as

$$V_o^*(t) = p(t)\ x(t) - q(t)\ y(t) + j\{q(t)\ x(t) + p(t)\ y(t)\} \quad (2.7)$$

where

$$p(t) = \frac{R[r(t)]\cos\theta[r(t)]}{r(t)} \quad (2.8)$$

$$q(t) = \frac{R[r(t)]\sin\theta[r(t)]}{r(t)} \quad (2.9)$$

The functions p(t) and q(t) respectively represent the normalized inphase and quadrature nonlinearities. Furthermore, it has been shown by Eric [9], that the inphase and quadrature nonlinearities $Z_p(r)$ and $Z_q(r)$, respectively are odd polynomial functions of the instantaneous input amplitude r(t). Formally,

$$Z_p(r) = R[r(t)]\cos\theta[r(t)] \quad (2.10)$$

$$Z_q(r) = R[r(t)]\sin\theta[r(t)] \quad (2.11)$$

and

$$Z_p(r) = a_1 r(t) + a_3 r(t)^3 + a_5(t)^5 + \dots + a_{2i+1} r(t)^{2i+1} \quad (2.12)$$

$$Z_q(r) = b_1 r(t) + b_3 r(t)^3 + \dots + b_{2i+1} r(t)^{2i+1} \quad (2.13)$$

with $b_1=0$, $i=1,2,3,\dots$

Hence

$$p(t) = a_1 + a_3 r(t)^2 + \dots + a_{2i+1} r(t)^{2i} \quad (2.14)$$

$$q(t) = b_3 r(t)^2 + b_5 r(t)^4 + \dots + b_{2i+1} r(t)^{2i} \quad (2.15)$$

$i=0,1,2,\dots$

The sets of coefficients $\{a_i\}$ and $\{b_i\}$ can be obtained through the use of a function minimization programme such as outlined by Eric [9].

Referring back to equation (2.7), which is the equivalent baseband output of the TWT transponder, we can easily produce an equivalent baseband representation of a bandpass nonlinearity. This model which is shown in Figure 2.6 is most suitable for computer simulation purposes because the complex baseband equivalent output of the nonlinear devices consists of simple functions of the complex baseband equivalent input. The most important limitation of this model is the assumption that the nonlinearity modelled is essentially memoryless. This is the case for a TWT and this model has been used extensively by Chan et al. [14] and Huang [15].

2.3.3 Nonlinear Effects on M-Ary PSK

We now proceed to consider the effects of a nonlinear satellite TWT transponder on the transmission of M-ary PSK. The basic system being considered is represented in Figure 2.7 and it illustrates a conventional non-regenerative satellite link with one nonlinearity namely the TWT in the satellite.

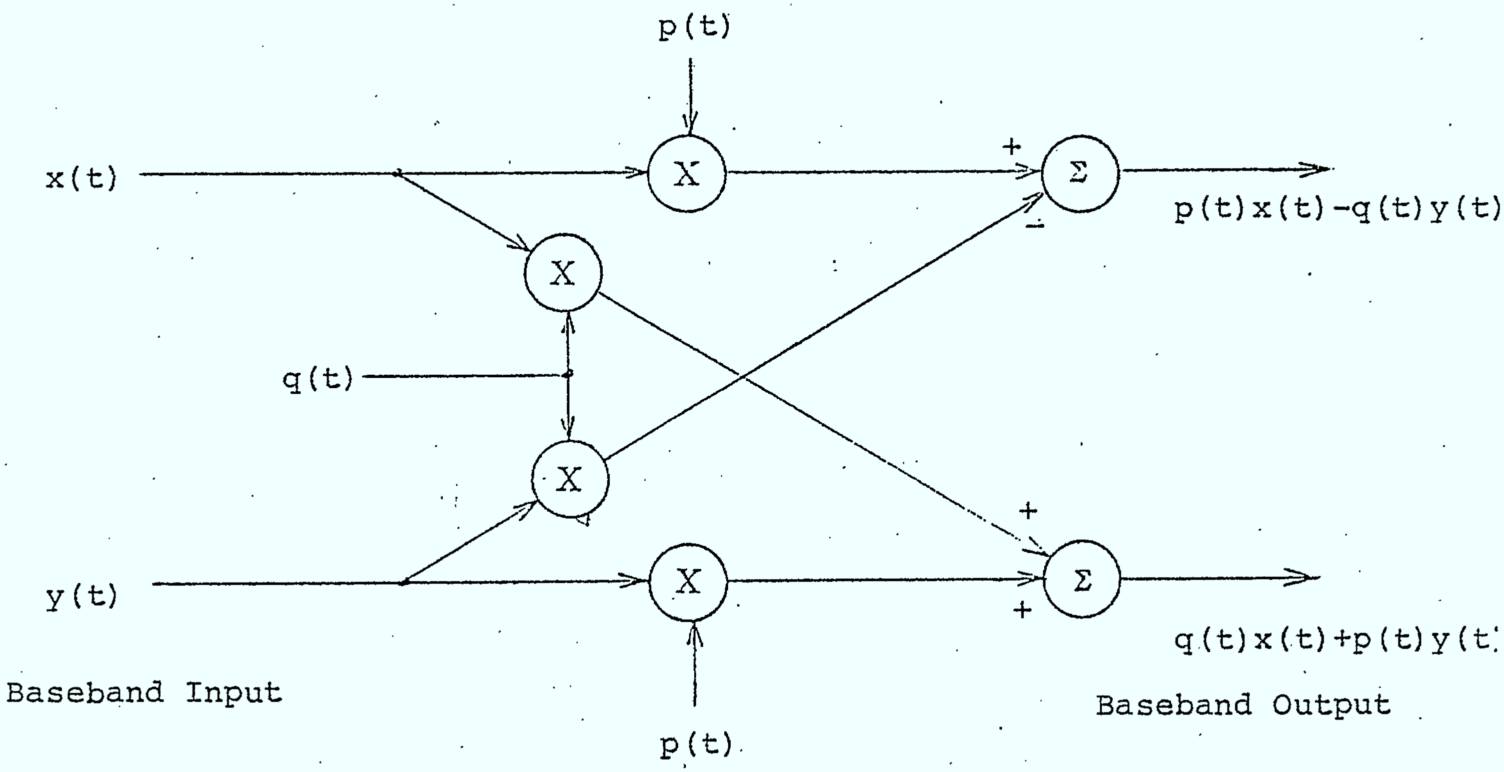

Fig. 2.6 Equivalent baseband model of a nonlinear device.

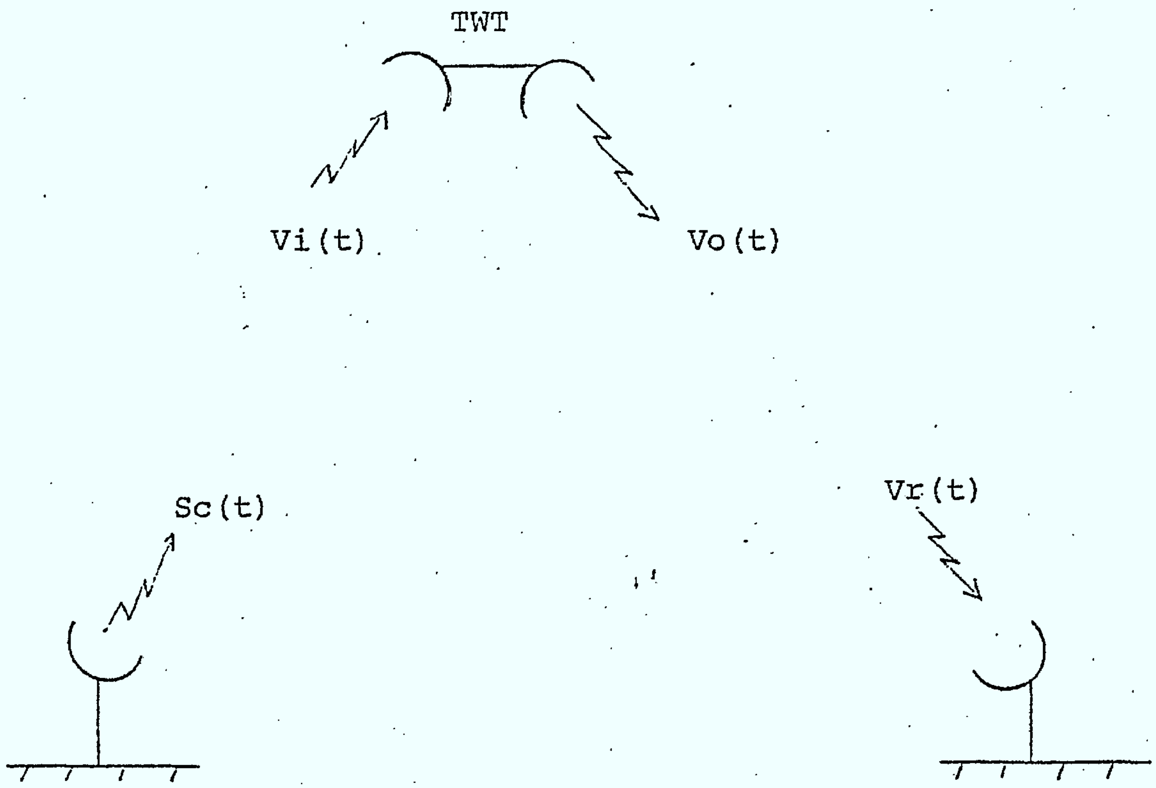

Fig.2.7 Illustration of a conventional non-regenerative satellite link.

Let the complex envelope of the M-ary PSK modulated signal transmitted by the earth station be given by:

$$S(t) = \sqrt{2Es} \left\{ \sum_{k=0}^{\infty} q(t-kt) \exp \left[j\phi_k - j\psi(t-kT) \right] \right\} \tag{2.16}$$

where $\phi_i = \frac{2\pi i}{M}$, $i = 0, 1, 2, \ldots, M-1$ (2.17)

The modulated signal with carrier is

$$Sc(t) = \mathrm{Re}\, S(t) \exp(jw_c t) \tag{2.18}$$

As the signal travels to the satellite it becomes corrupted by uplink radio frequency interference and gaussian interference. The signal received at the satellite, assuming no attenuation of the transmitted signal, is given by:

$$Vi(t) = Sc(t) + Nu(t) + J(t) \tag{2.19}$$

where $Nu(t)$: uplink channel gaussian noise

$J(t)$: uplink radio frequency interference.

Expressing the two interfering signals $Nu(t)$ and $J(t)$ by their bandpass representations and then by their complex equivalent baseband representation, we have:

$$\begin{aligned} Nu(t) &= n_c(t)\cos w_c t - n_s(t)\sin w_c t \\ &= \mathrm{Re}\left[n_c(t) + j\, n_s(t) \right] \exp(jw_c t) \end{aligned} \tag{2.20}$$

and

$$\begin{aligned} J(t) &= J_c(t)\cos w_c t - J_s(t) \sin w_c t \\ &= \mathrm{Re}\left[J_c(t) + jJ_s(t) \right] \exp(jw_c t) \end{aligned} \tag{2.21}$$

After substitution of (2.16), (2.20) and (2.21) into (2.19) we obtain:

$$\begin{aligned} Vi(t) = \mathrm{Re}\Big\{ &\Big[\sqrt{2Es} \sum_{k=0}^{\infty} q(t-kT)\cos \left[\phi_k - \psi(t-kT) \right] + n_c(t) + J_c(t) \Big] \\ &+ j \Big[\sqrt{2Es} \sum_{k=0}^{\infty} q(t-kT) \sin \left[\phi_k - \psi(t-kT) \right] + n_s(t) + J_s(t) \Big] \Big\} \cdot \\ &\exp(jw_c t) \end{aligned} \tag{2.22}$$

Thus, the expression for Vi(t) is of the simple form:

$$Vi(t) = [A + jB] \exp(jw_c t) \tag{2.23}$$

Equation (2.23) can also be written as

$$Vi(t) = E(t) \exp j [w_c t + \eta(t)] \tag{2.24}$$

where $E^2(t) = A^2 + B^2$

$$\eta(t) = \tan^{-1} \frac{B}{A} \tag{2.25}$$

After passing the signal Vi(t) through a nonlinear device we obtain the signal Vo(t) which is of the form

$$Vo(t) = R(E) \exp j [w_c t + \eta(t) + \Theta(E)] \tag{2.26}$$

where R(E) is the AM/AM conversion

and $\Theta(E)$ is the AM/PM conversion

As the signal Vo(t) travels down to the earth receiving station, downlink gaussian noise is added to the signal. Thus the received signal at the earth station is

$$Vr(t) = Vo(t) + n(t) \tag{2.27}$$

where n(t) is downlink gaussian noise.

Assuming that the demodulator consists of a correlation detector as shown in Figure 2.8, then the signals at the samplers are:

$$S = Z + n_{c2} \tag{2.28}$$

$$\tilde{S} = \tilde{Z} + n_{s2} \tag{2.29}$$

where $$Z = R(E) \cos [\eta + \bar{\phi} + \Theta(E)] \qquad (2.30)$$

$$\tilde{Z} = R(E) \sin [\eta + \bar{\phi} + \Theta(E)] \qquad (2.31)$$

for which $\bar{\phi}$ is the phase offset in the recovered carrier and n_{c2} and n_{s2} are respectively the inphase and quadrature components of the downlink gaussian noise.

Based on the signal pair $(s,\tilde{s})$, the detector determines which one of M phases was transmitted. Assume that $(Z,\tilde{Z})$ is known, and the phase reference $\bar{\phi}$ is fixed. Then, the conditional error probability is

$$Pe\ (Z,\tilde{Z},\bar{\phi}) = Pr \left\{ (s,\tilde{s}) \notin Ro \;\middle|\; Z, \tilde{Z}, \phi \right\} \qquad (2.32)$$

where Ro represents the decision cone of $2\pi/M$ radians centered at $\phi = 0$. An example for the case of M=4 is shown in Figure 2.9.

For $M = 2,4$ it is possible to obtain exact expressions for $Pe(Z,\tilde{Z},\phi)$. However, for $M = 2^k$, $K \geq 3$ numerical techniques have to be used. Bounds such as presented by Huang et al. [12] may also be used.

Notwithstanding the complexity of the analysis if we are to attempt to obtain exact expressions for the probability of error, we can obtain from the development demonstrated in this section an appreciation for the effects caused by the nonlinearity. The difficulties introduced by the nonlinearity are even more complicated than what one is led to believe by Huang et al. [12] who have assumed that the nonlinearity exhibits only AM/AM conversion. Thus a system simulation as presented by Chan et al. [14] and Huang [9] is the only viable alternative presently available.

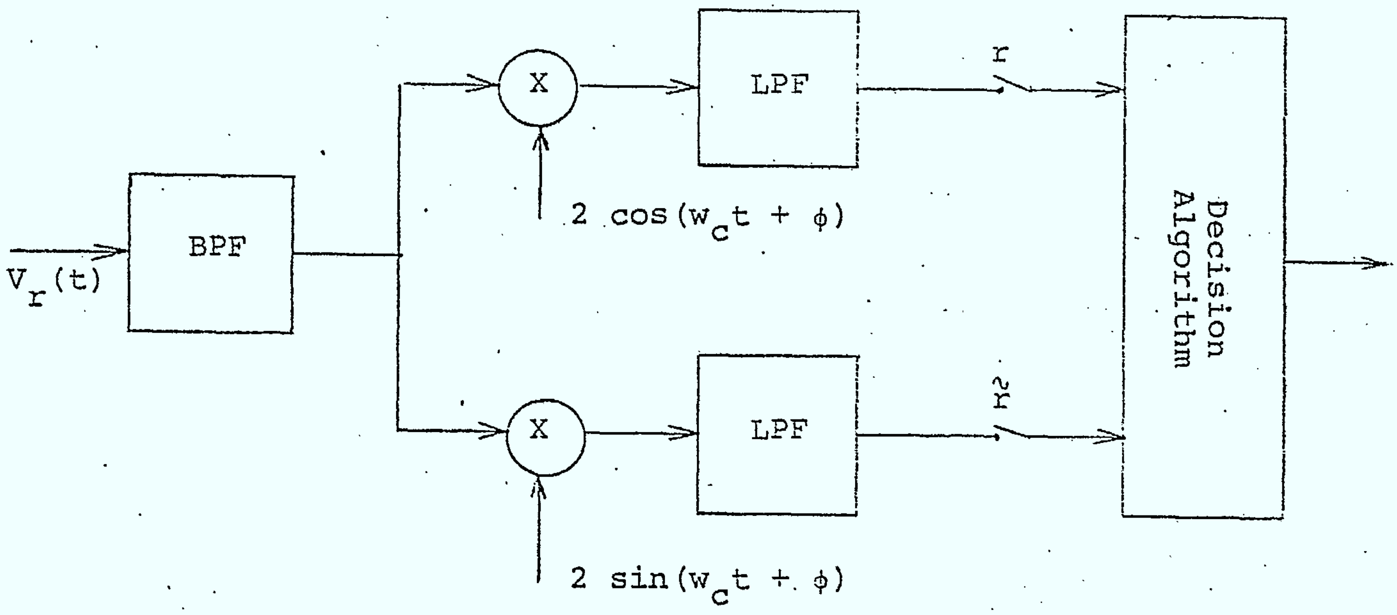

Fig. 2.8 Correlation detection demodulator

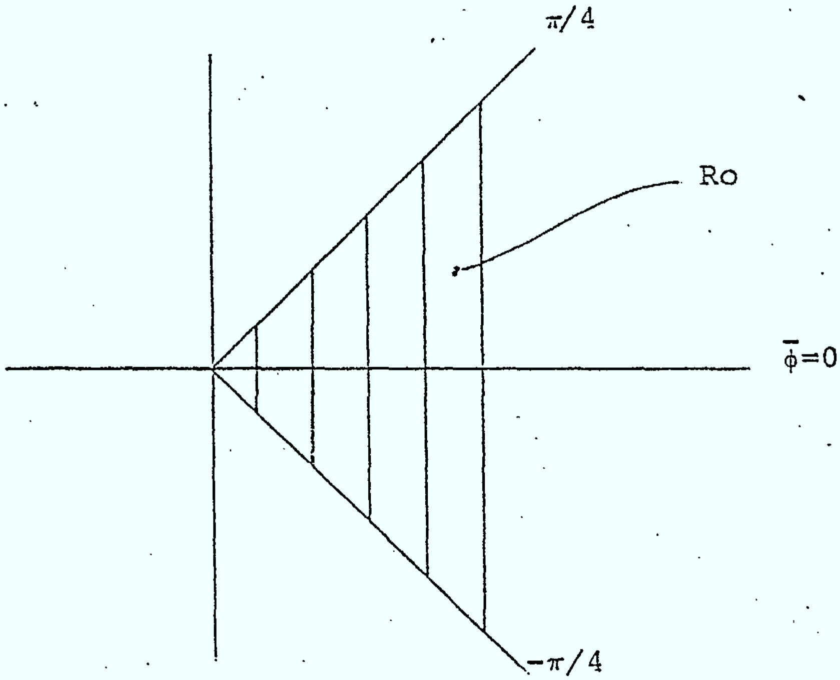

Fig. 2.9 Decision region for M=4

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CHAPTER 3

NON-REGENERATIVE SYSTEMS STUDY

3.1 INTRODUCTION

In the previous chapter, the conventional (non regenerative) satellite system was described briefly. It was memtioned that the three most frequently considered modulation techniques for operation in a nonlinear channel are QPSK, OKQPSK and MSK. The basic schemes for these three modulation techniques will be reviewed in the next section. A study and performance comparison of these three signals in a bandlimited satellite channel is undertaken in this chapter. To this end a computer simulation programm that was used is briefly described, and the results of these simulations are analyzed and compared with some of those available in the literature.

The chapter is based on and follows closely the work done by J. Huang in his Ph.D. thesis [1].

The typical conventional satellite system model is redrawn in Figure 3.1 for reference purposes.

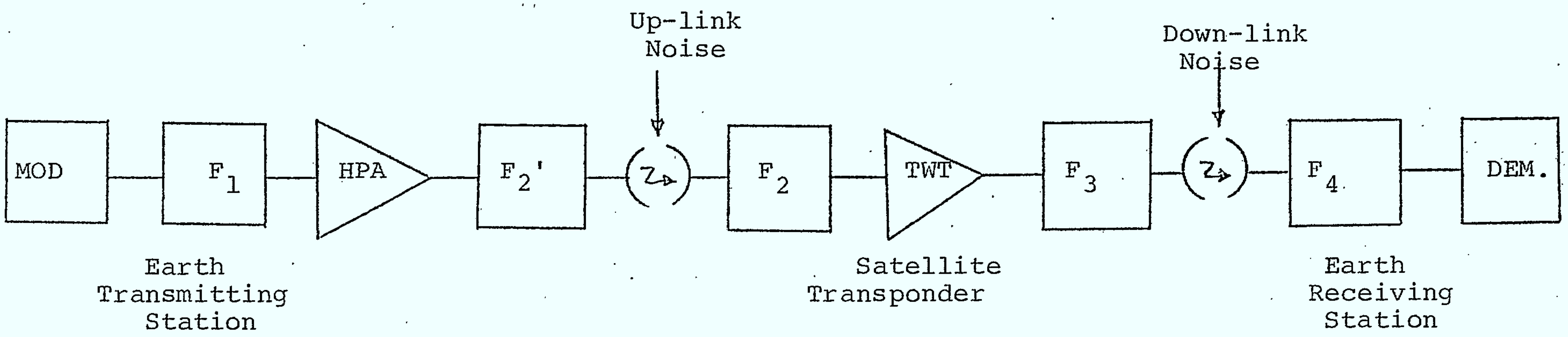


Figure 3.1 - Conventional Digital Satellite Communication System Model

3.2 CONVENTIONAL QPSK OFFSET QPSK AND MSK

As mentioned in the previous section, three modulation techniques, namely, conventional QPSK, offset QPSK and MSK, have been frequently considered for signal transmission through a nonlinear satellite channel. In this section, the basic schemes of these three modulation techniques are described.

In conventional QPSK (Quaternary Phase-Shift Keying), the in-phase and the quadrature phase data streams, as shown in Figure 3.2a, are aligned coincidently.

In OKQPSK (Offset QPSK) and MSK(Minimum Shift Keying), one data stream is delayed by half a symbol interval in comparison with the other data stream (Figs. 3.2.b and 3.2c). The difference between OKQPSK and MSK lies in the extra pulse shaping networks which shape the rectangular pulses of OKQPSK into the half cosine pulses of MSK.

Simplified block diagrams of these three modulation schemes are shown in Fig.3.3. The binary input random data stream having a bit interval T is first serial to parallel converted into two data streams. These two data streams have a symbol duration of 2T as shown in Fig. 3.2.

In the case of conventional QPSK, they are then quadrature modulated by a carrier. To obtain OKQPSK from QPSK, one of the two parallel data streams is first delayed by a bit interval relative to the other data stream. Afterwards, they are then quadrature modulated

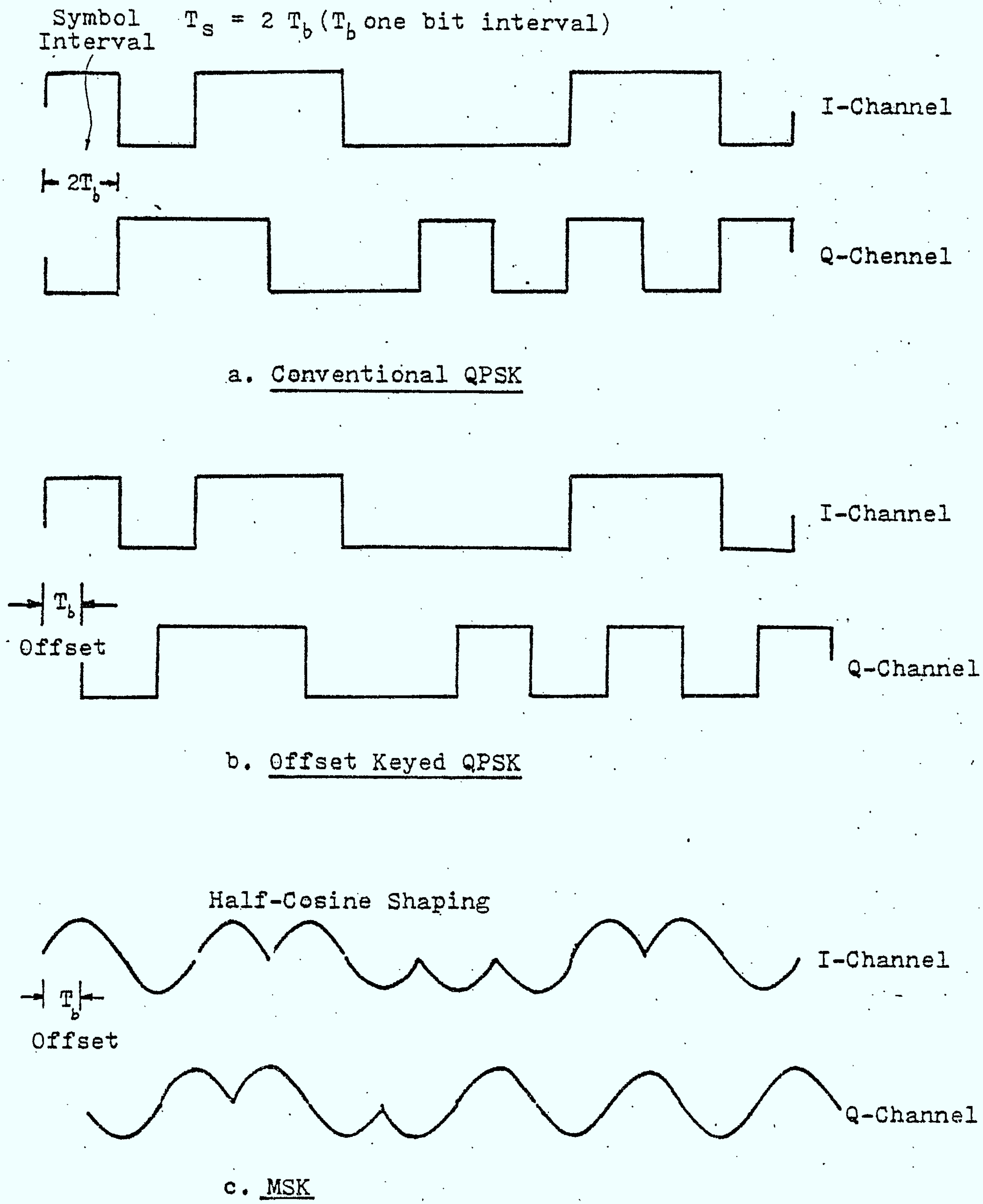
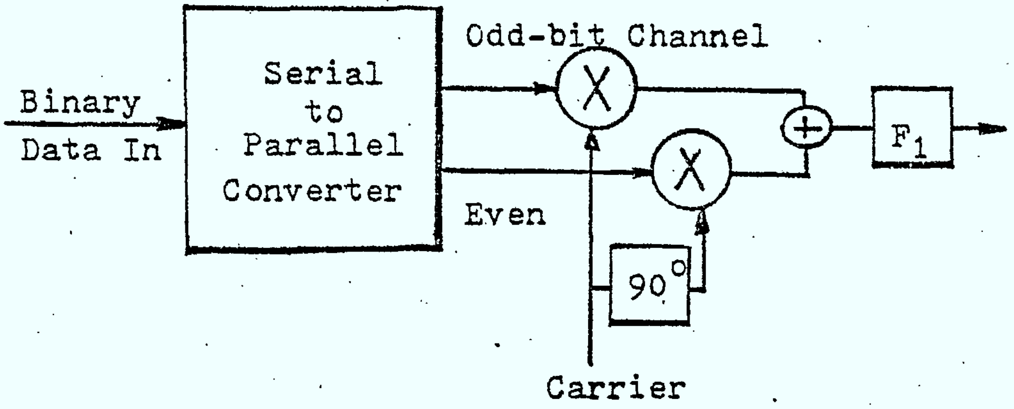
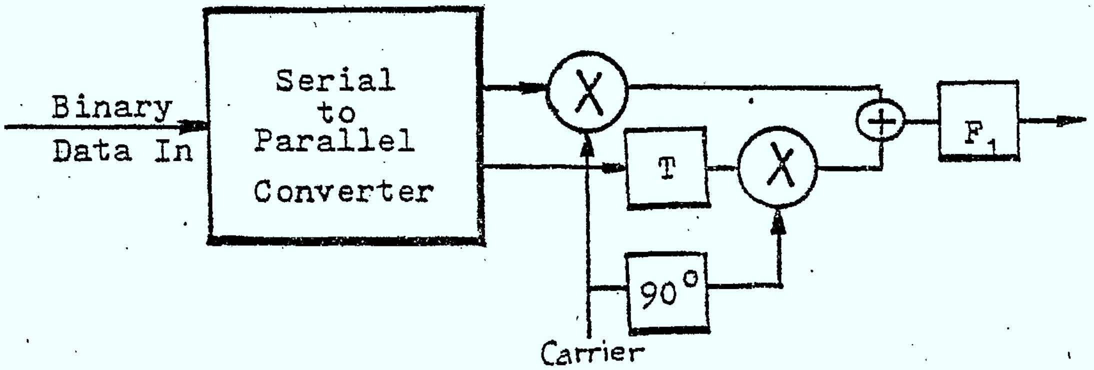


Figure 3.2 <u>Signaling Format of Conventional QPSK, OKQPSK and MSK</u>

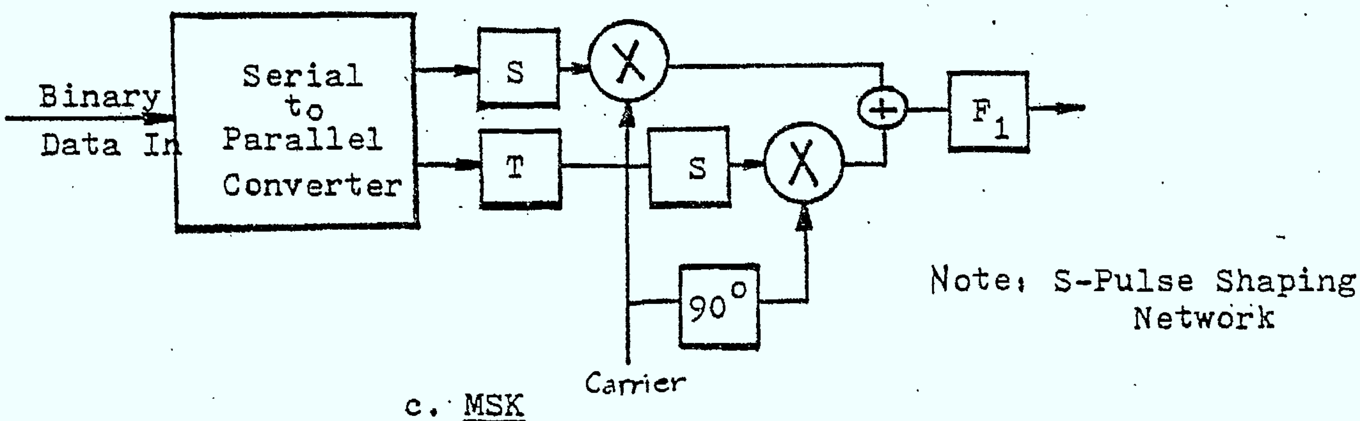
as in conventional QPSK. MSK is similar to OKQPSK except that a pulse shaping network modifies the rectangular pulses of OKQPSK into half cosine pulses in these two parallel data streams before they are quadrature modulated. In all three cases, a bandpass filter F_1 after the modulator is used for spectrum shaping purpose.



a. Conventional QPSK



b. Offset Keyed QPSK



c. MSK

Figure 3.3 Block Diagrams of Conventional QPSK, OKQPSK and MSK

3.2.1 FFSK View of MSK

MSK can also be viewed as one type of the general Continuous Frequency Shift Keying (CFSK) modulation schemes. In this sense, it is called Fast Frequency Shift Keying (FFSK) [2]. As indicated in Fig. 3.4, the frequency deviation in FFSK $\Delta f = \frac{f_1 - f_2}{2}$ is exactly equal to $\pm \frac{1}{4} T$, where f_1 is the frequency of the signal representing the 1 symbol and f_2 is the frequency of the signal representing the 0 symbol.

This view of looking at MSK as FFSK can easily be demonstrated as follows:

For a frequency shifted keying signal,

$$s(t) = A \cos\left[2\pi(f_c \pm \Delta f)t\right] \tag{3.1}$$

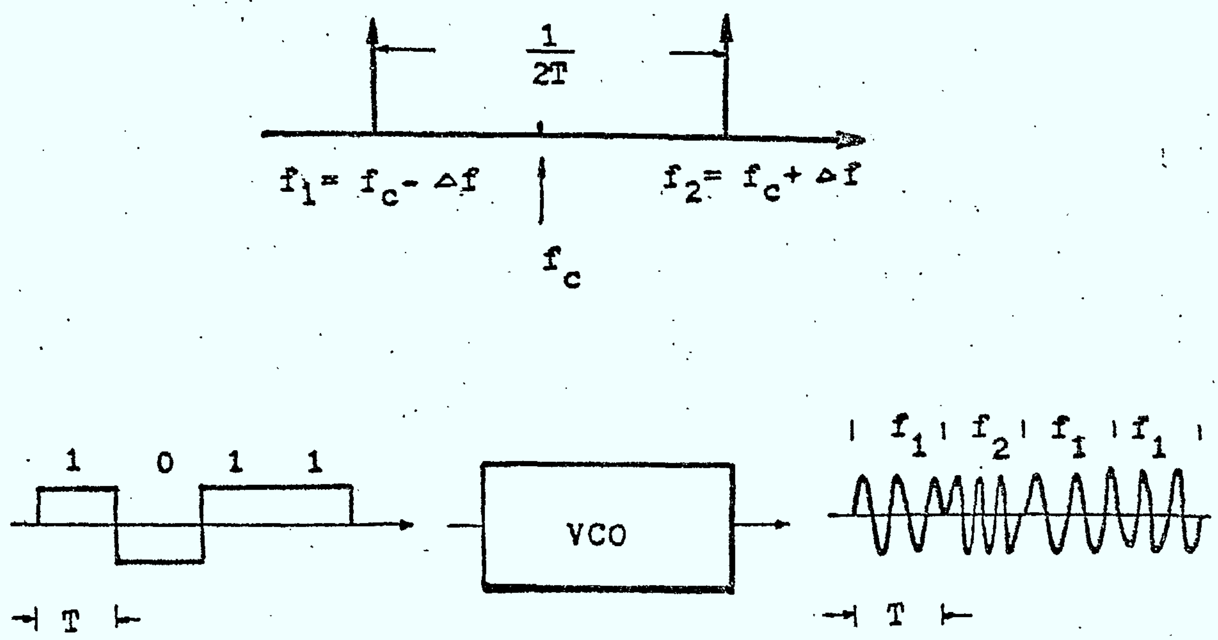


Fig. 3.4 FFSK View of MSK

which can be considered as the transmission of a sinusoid whose frequency is shifted between the following two frequencies:

$$f_1 = f_c - \Delta f \tag{3.2}$$

and

$$f_2 = f_c + \Delta f$$

The FSK signal s(t) in Eq. (3.1) can then be expanded as:

$$A \cos (\pm 2\pi \Delta f t) \cos 2\pi f_c t - A \sin (\pm 2\pi \Delta f t) \sin 2\pi f_c t \tag{3.3}$$

Now, if the frequency deviation is

$$\Delta f = \tfrac{1}{2}(f_2 - f_1) = \pm 1/4T \tag{3.4}$$

where T is the unit bit duration of the input random data stream.

Then,

$$s(t) = A \cos (\pm \pi t/2T) \cos 2\pi f_c t - A \sin (\pm \pi t/2T) \sin 2\pi f_c t \tag{3.5}$$

Thus, s(t) is an offset quadrature carrier signal with half a cosine pulse shaping in its baseband signals. The baseband signals are:

$$x(t) = A \cos (\pm \pi t/2T)$$

and (3.6)

$$y(t) = A \sin (\pm \pi t/2T).$$

It is important to notice that,

$$x^2(t) + y^2(t) = A^2 \tag{3.7}$$

This feature is characteristic of the baseband signaling waveform suitable for applications in CFSK. This signaling format is also known as MSK-formatted signal [3], or MSK-type signal [4].

3.2.2 Spectra of QPSK, OKQPSK and MSK

The spectra of conventional QPSK (hereafter abbreviated as QPSK), OKQPSK and MSK are shown in Fig. 3.5. The power spectrum of OKQPSK is identical to that of the QPSK. Both have the (sin x/x) shape centered at the carrier frequency. Mathematically, the spectra of both the QPSK and OKQPSK are given by [5]:

$$S_{\substack{QPSK \\ OKQPSK}}(f) = 2\, P_c T \left(\frac{\sin 2\pi f T}{2\pi f T}\right)^2 \tag{3.8}$$

where f = frequency offset from carrier, P_c = power of the modulated waveform, T = unit bit duration.

The spectrum of a MSK signal is given also by [5]

$$S_{MSK}(f) = \frac{8\, P_c\, T(1 + \cos 4\pi f T)}{\pi^2 (1 - 16 f^2 T^2)^2} \tag{3.9}$$

Eqs. (3.8) and (3.9) can easily be obtained from the well-known fact that the power spectral density of a PSK signal with a baseband signaling waveform s(t) has the same spectral shape as that of s(t) itself. The only difference is that the spectrum of the PSK signal is double-sided and centered at the carrier frequency [6].

Comparing the spectrum of MSK with that of QPSK (or OKQPSK) we find that the main lobe width of MSK is 50% wider than that of QPSK (or OKQPSK). The sidelobes of MSK, on the other hand, are much lower than those of QPSK (or OKQPSK).

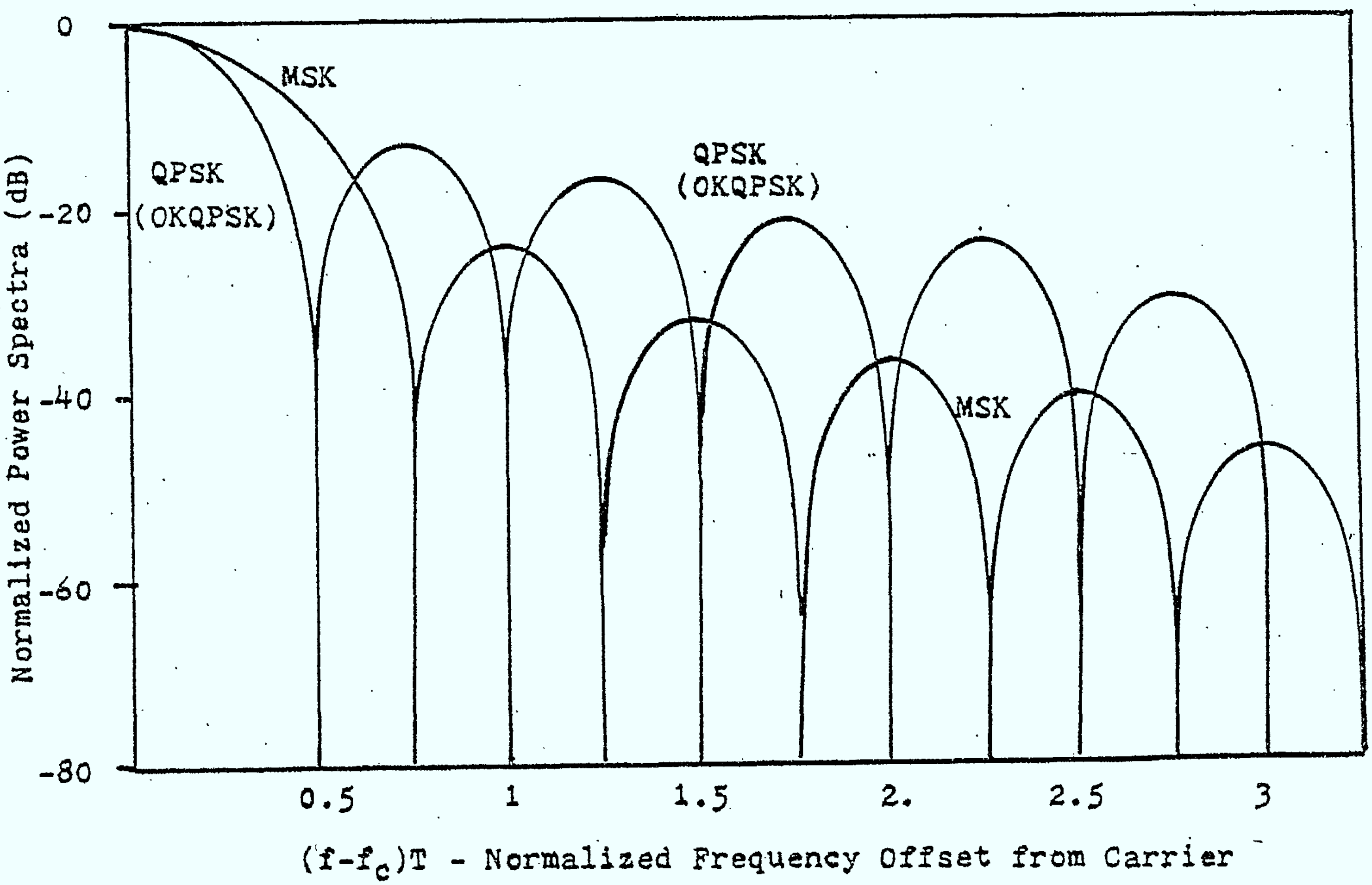


Fig. 3.5 <u>Power Spectra for MSK and QPSK (OKQPSK)</u>

(T = 1 sec. P_c = 1 W.)

3.3 OVERALL ENVELOPE FLUCTUATIONS OF QPSK, OKQPSK AND MSK SIGNALS

Due to the time coincidence of the two data streams in QPSK, phase transitions of 0°, $\pm 90^{\circ}$, and 180° can occur. Fig. 3.6.a shows the RF amplitudes and phases of the filtered QPSK signal. At the 180° instantaneous phase transitions, which arise when both the in-phase channel data and the quadrature channel data change phase simultaneously, the envelope goes through zero amplitude. At the $\pm 90^{\circ}$ phase transitions, which occur when only one channel data change phase at the keying instant, there is a 3dB envelope fluctuation.

In OKQPSK, because of the one bit (half a symbol) duration delay between the two data channels, the 180° phase transitions are avoided. Only $\pm 90^{\circ}$ phase transitions arise with maximum 3 dB envelope fluctuations as shown in Fig. 3.6.b. The overall envelope fluctuations of OKQPSK are thus smaller than those of QPSK.

In MSK, the data waveforming pulses are of half cosine shape as shown in Fig. 3.6.c. According to Eq.(3.7), the signal has a constant envelope. Another important feature of MSK is that the phase transitions are linear and continuous [2].

The overall envelope fluctuations of QPSK, OKQPSK and MSK can also be studied from the signal space diagrams, as shown in Fig.3.7. Fig.3.7.a shows the signal space diagrams of these three signals using computer simulation.

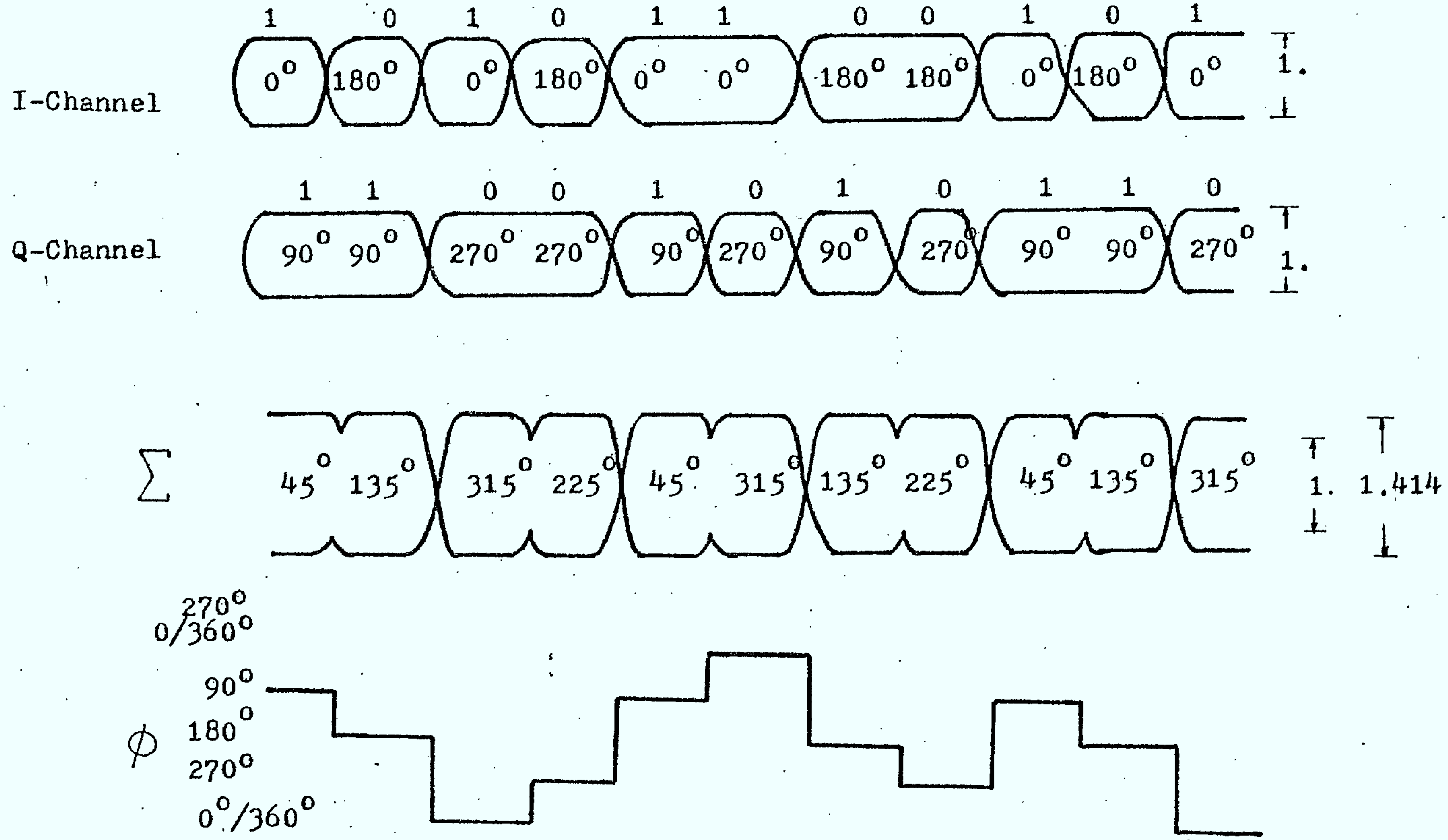


Fig. 3.6.a RF Amplitude and Phase of Conventional QPSK Signal

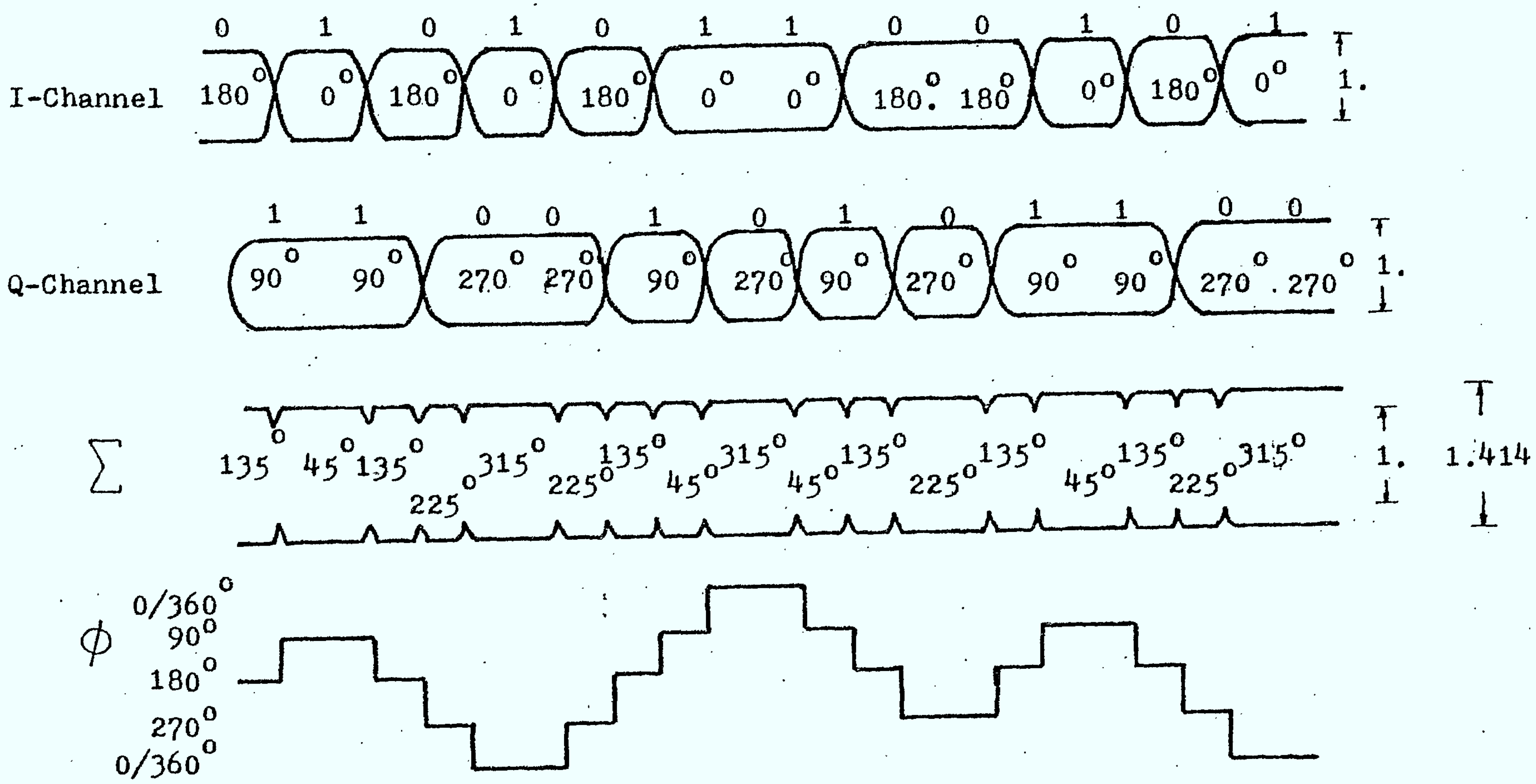


Fig. 3.6.b RF Amplitude and Phase of OKQPSK Signal

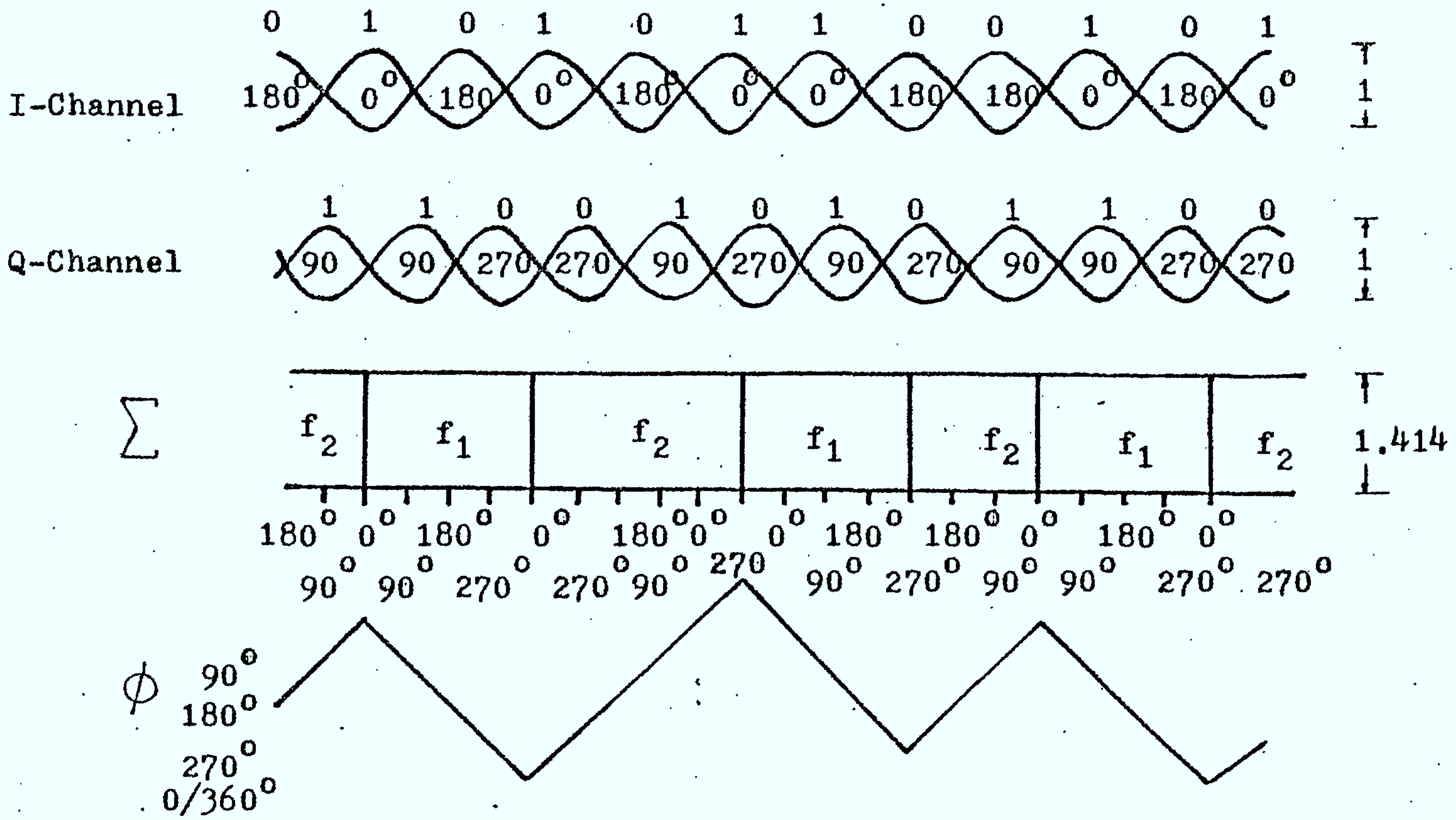
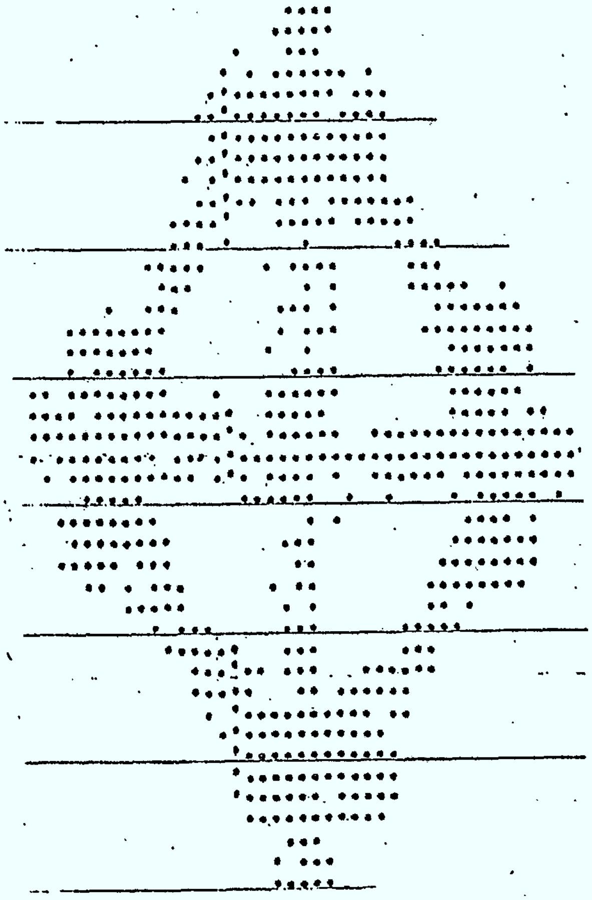


Fig. 3.6.c RF Amplitude and Phase of MSK Signal



Bit Rate 90 Mb/s

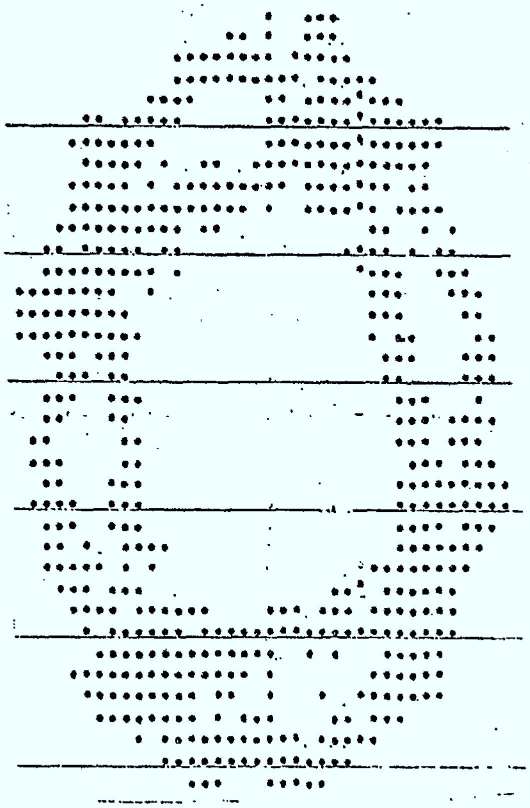
$F_1 = F_4 = 54$ MHz

$F_2 = F_3 = 125$ MHz

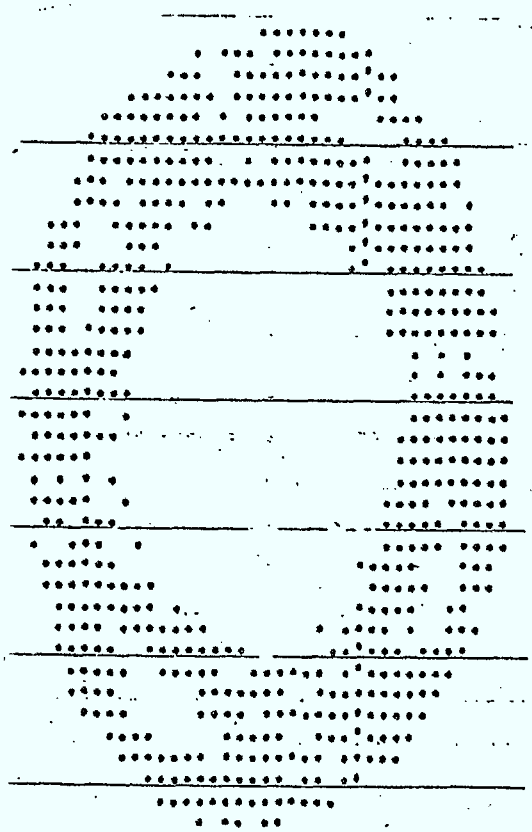
(f_{3dB} of the 4 pole Chebychev filters)

HPA and TWT at 1 dB input backoff

a. QPSK



b. OKQPSK



c. MSK

Fig. 3.7 Computer-Simulated Signal Space Diagrams

In computer simulation, the received quadrature signals were assigned to quadrature components of a two-dimensional matrix and stored in the memory. At the end of the simulation, the signal space diagram was then plotted from the computer memory of the final matrix.

The signal space diagrams show that whenever the in-phase and quadrature channel data change states simultaneously, the QPSK signal goes through the origin (zero amplitude). In OKQPSK and MSK, because of the time delay between the two parallel channel data, this does not occur.

The inherent envelope fluctuation of QPSK therefore causes distortion and more spectrum spreading when it passes through a nonlinear amplifier. The spectrum spreading of QPSK and OKQPSK will be demonstrated and compared experimentally in section 3.5.

In the following we will analyze the overall envelope variations of the filtered QPSK and OKQPSK signals. This analysis, in conjunction with the pictorial description of Fig. 3.6, will also be used in our discussion on the envelope fluctuations of the filtered QPSK and OKQPSK signals at the sampling instants.

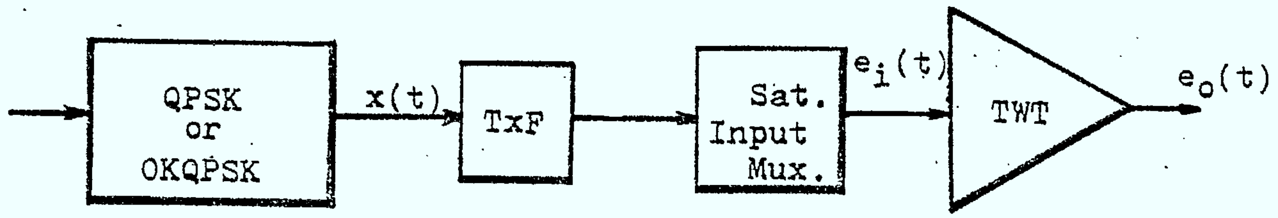


Fig. 3.8 Simplified Digital Satellite Channel Model

To simplify our discussion, the quadrature carrier communication system model shown in Fig. 3.1 will be assumed to include only the transmit filter and the TWT input MUX filter and the TWT itself as depicted in Fig. 3.8. This simplified model is adopted because we are only interested in the envelope fluctuations caused by the bandlimiting filters and one nonlinear device. A similar model has also been used by Robinson, et al in their theoretical analysis on spectrum spreading of QPSK [7] and by McMaster University in their comparative performance study by computer simulation. [8]. The initial part of the analysis on QPSK follows [7] which is then extended to OKQPSK.

The QPSK or OKQPSK signal x(t) at the transmit filter TXF input can be written as:

for QPKS

$$x(t) = \sum_{-\infty}^{\infty}\left[a_k\, s(t - kT_s) \cos 2\pi f_c t\right] + \sum_{-\infty}^{\infty}\left[b_k\, s(t - kT_s) \sin 2\pi f_c t\right] \tag{3.10}$$

for OKQPSK,

$$x(t) = \sum_{-\infty}^{\infty}\left[a_k\, s(t - kT_s) \cos 2\pi f_c t\right] + \sum_{-\infty}^{\infty}\left[b_k\, s(t - T_s/2 - kT_s) \sin 2\pi f_c t\right] \tag{3.11}$$

where a_k, $b_k = \pm 1$ with equal probability of occurrence.

f_c is the carrier frequency, T_s is the symbol interval (T_s=2T) and

$$s(t) = \begin{cases} \text{rectangular pulse for } |t| \leqslant T_s/2 \\ 0 \qquad\qquad\qquad \text{otherwise} \end{cases}$$

x(t) can be rewritten as:

QPSK,

$$x(t) = \text{Re} \left\{ e^{j2\pi f_c t} \cdot (a_k + j\, b_k)\, s(t - kT_s) \right\} \tag{3. 12}$$

OKQPSK,

$$x(t) = \text{Re} \left\{ e^{j2\pi f_c t} \left[a_k\, s(t - kT_s) - j\, b_k\, s(t - T_s/2 - kT_s) \right] \right\} \tag{3. 13}$$

where Re stands for the real part of a complex variable.

Assuming h(t) is the low-pass equivalent of the impulse response of the filters between the quadrature carrier modulator and the TWT, then, the filtered equivalent baseband signal at the input of the TWT is [9]:

for QPSK,

$$\begin{aligned} e_i(t) = & \sum_K \overbrace{\left[a_k\, R(t - kT_s) - b_k\, I(t - kT_s) \right]}^{y(t)} \cos 2\pi f_c t \\ & - \sum_K \underbrace{\left[a_k\, I(t - kT_s) + b_k\, R(t - kT_s) \right]}_{z(t)} \sin 2\pi f_c t \end{aligned} \tag{3 . 14}$$

and

for OKQPSK,

$$\begin{aligned} e_i(t) = & \sum_K \overbrace{\left[a_k\, R(t - kT_s) - b_k\, I(t - T_s/2 - kT_s) \right]}^{y(t)} \cos 2\pi f_c t \\ & - \sum_K \underbrace{\left[a_k\, I(t - kT_s) + b_k\, R(t - T_s/2 - kT_s) \right]}_{z(t)} \sin 2\pi f_c t \end{aligned} \tag{3. 15}$$

where $R(t) = \frac{1}{2} \operatorname{Re}\left[s(t) * h(t) \right]$

$I(t) = \frac{1}{2} \operatorname{Im}\left[s(t) * h(t) \right]$

and (3.16)

$$R(t - kT_s) = \frac{1}{2} \operatorname{Re}\left[s(t - kT_s) * h(t) \right]$$

$$I(t - kT_s) = \frac{1}{2} \operatorname{Im}\left[s(t - kT_s) * h(t) \right]$$

with * denoting convolution.

The IF signal $e_i(t)$ can be written for both QPSK and OKQPSK as

$$\begin{aligned} e_i(t) &= y(t) \cos 2\pi f_c t + z(t) \sin 2\pi f_c t \\ &= \sqrt{y^2(t) + z^2(t)} \cos\left(2\pi f_c t + \tan^{-1}\left[y(t)/z(t)\right] \right) \end{aligned} \tag{3.17}$$

The amplitude of $e_i(t)$ is:

for QPSK,

$$\begin{aligned} A(t) &= \sqrt{y^2(t) + z^2(t)} = \sqrt{2\sum_K \left[R^2 (t - kT_s) + I^2 (t - kT_s)\right]} \\ &= \sqrt{\frac{1}{2}\sum_K v^2 (t - kT_s)} \end{aligned} \tag{3.18}$$

and

for OKQPSK,

$$\begin{aligned} A(t) &= \sqrt{y^2(t) + z^2(t)} \\ &= \sqrt{2\sum_K \left\{\left[R^2(t - kT_s) + I^2(t - kT_s)\right] + \left[R^2(t - T_s/2) + I^2(t - T_s/2 - kT_s)\right]\right\}} \\ &= \sqrt{\frac{1}{4}\sum_K \left[v^2 (t - kT_s) + v^2 (t - T_s/2 - kT_s)\right]} \end{aligned} \tag{3.19}$$

where $v(t)$ is the envelope of the filtered signal and $v^2(t) = R^2(t) + I^2(t)$.

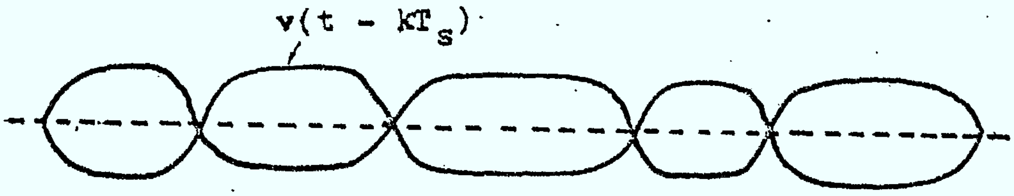
The envelope signal $v(t - kT_s)$ generally looks as shown in Fig. 3.9. a. So for QPSK, the envelope has zero amplitudes when both data channels change transitions simultaneously. The overall envelope fluctuations of QPSK can change from maximum to zero.

For OKQPSK, because of the time delay between $v(t - T_s/2 - kT_s)$ and $v(t - kT_s)$, there are no zero crossings and the overall envelope fluctuations are smaller than QPSK as shown in Fig. 3.9. b.

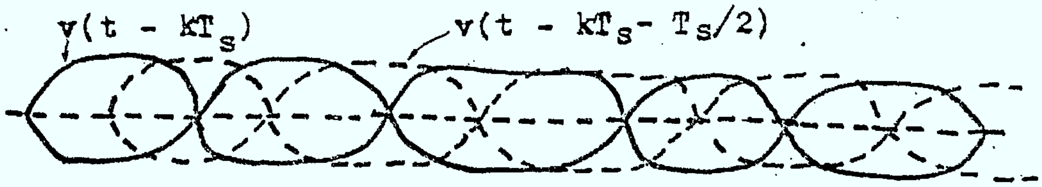
If the amplitude distortion (AM/AM) and the phase distortion (AM/PM) functions are assumed to be g(.) and f(.), respectively, then the signal at the output of the TWTA is:

$$e_o(t) = g(A(t)) \cos (2\pi f_c t + \tan^{-1} y/z + f(A(t))) \qquad (3.20)$$

Large amplitude variations will hence induce more signal distortions due to the nonlinearity effect of the TWTA. Depending on the g(.) and f(.) functions, the spectrum spreading effect can be computed using the methods described by Robinson et al, [7] and Palmer et al. [11].



a. QPSK



b. OKQPSK

Fig. 3.9 Overall Envelope Fluctuations of QPSK and OKQPSK Signals

3.4 ENVELOPE FLUCTUATIONS OF FILTERED QPSK, OKQPSK AND MSK SIGNALS AT THE SAMPLING INSTANTS

It was shown in the previous section that by delaying one channel data by a bit interval in respect to the other channel data, OKQPSK and MSK have much smaller overall envelope fluctuations than those of QPSK. They therefore suffer less spectrum spreading than QPSK when passing through a nonlinear power amplifier. The subject of spectrum spreading will be covered in a later section.

On the other hand, in a bandlimited satellite channel, the OKQPSK (and MSK) signal will be shown to have larger envelope fluctuation at the sampling instants than those of QPSK. As these three modulation schemes belong to four phase modulation, a multiply by four method is frequently employed for carrier recovery [12].
The larger amplitude fluctuations of OKQPSK (and MSK) at the sampling instants imply much larger variations in the fourth powered signal for carrier recovery. If baseband impulses are assumed instead of the rectangular pulse, then the QPSK signal and its signal space diagram are as shown in Fig. 3.10. a, while the OKQPSK signal and its signal space diagram are shown in Fig. 3. 10. b. Clearly, because of the offsetting in the two parallel data streams in OKQPSK, when the in-phase channel (I-channel) is sampled, there is no signal in the quadrature channel (Q-channel) and vice versa. In the case of QPSK, there are always signals at the sampling instants at both channels.

When both QPSK and OKQPSK are bandlimited, the impulses become shaped pulses as shown in Fig. 3.11.b. For QPSK, the signal amplitudes at the sampling points do not change too much if no large intersymbol interference effect from the filter is assumed. For OKQPSK, the signal envelope at the sampling points is no longer only a function of the sampled channel. This is because sampling points in one channel are transition points in the other channel and vice versa. The signal states of OKQPSK(and MSK), as shown in Fig. 3.11, lie on a line.

The signal states of QPSK on the other hand, as shown in Fig. 3.11, are still located near the four points of Fig. 3.9.a without too much variation.

To verify that this is indeed the case, computer simulation was used to plot the signal scatter diagrams at the sampling instants. The signal scatter diagrams of QPSK, OKQPSK and MSK, as shown in Fig. 3.12, are obtained by sampling the received QPSK signal once per symbol and the received OKQPSK and MSK signal twice per symbol. These scatter diagrams agree with those described in Fig. 3.11.

When fourth powered, the four signal states of QPSK will converge to one point as shown in Fig. 3.13. This is because the received signal,

$$s(t) = \cos (2\pi f_c t + \theta) \qquad (3.21)$$

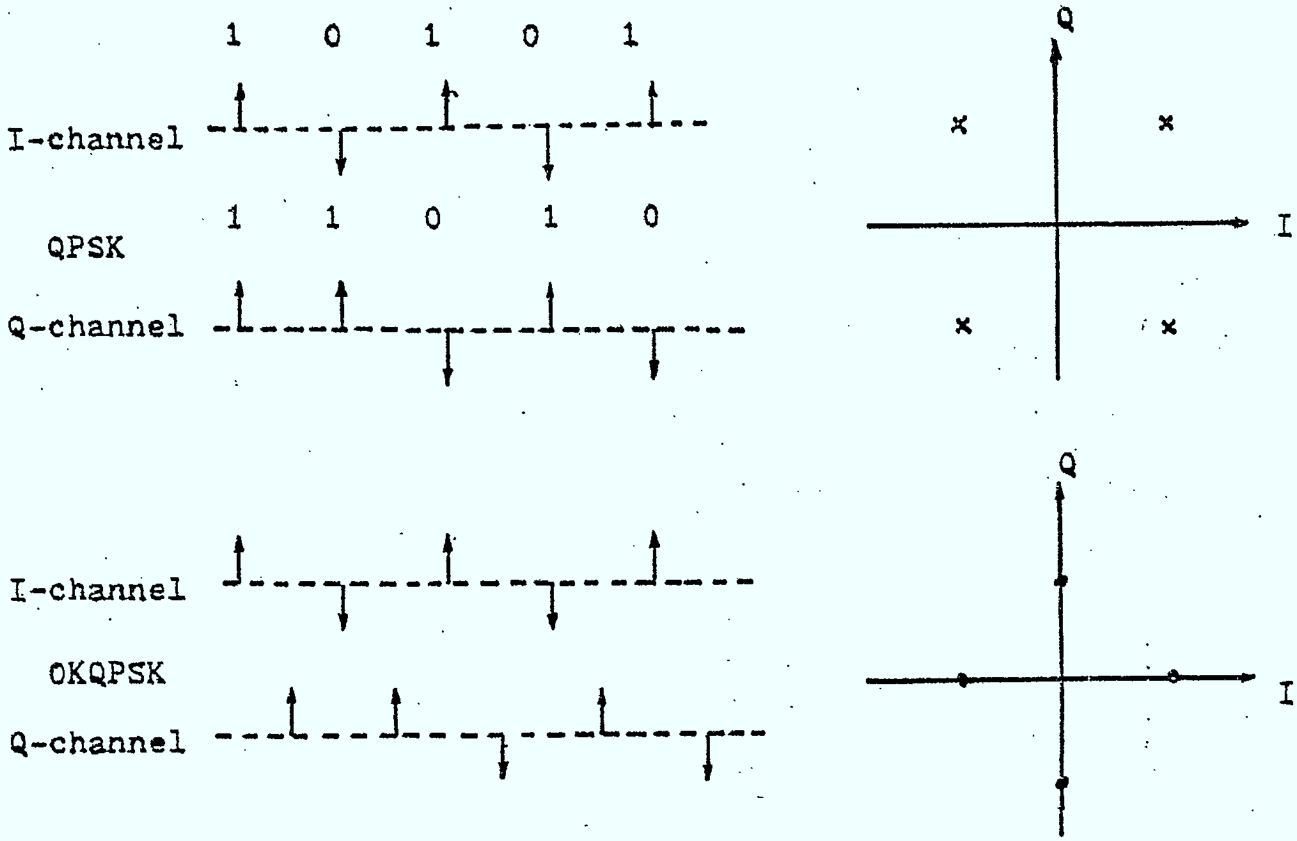


Fig. 3.10 Impulsed QPSK and OKQPSK Signals and Corresponding Signal Space Diagrams

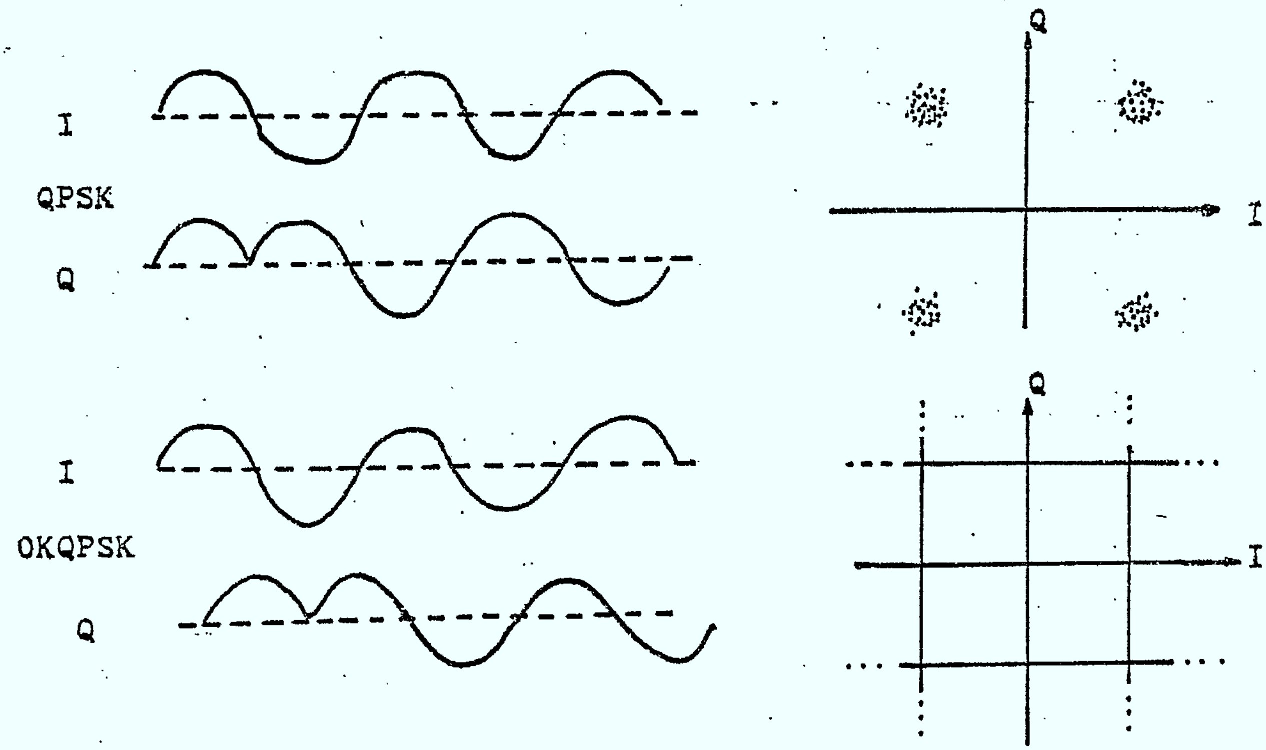
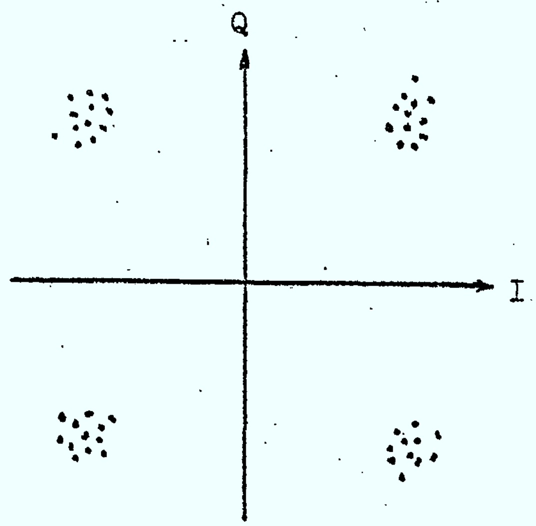
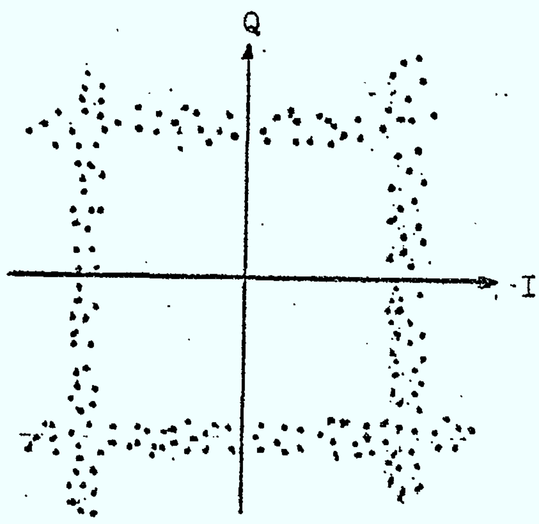


Fig. 3.11 Pulsed QPSK and OKQPSK Signals and Corresponding Scatter Diagrams

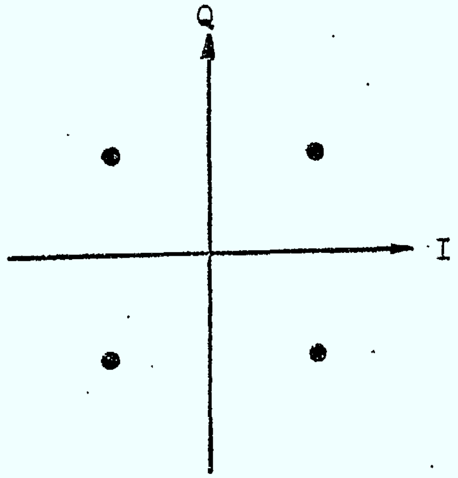


a. QPSK

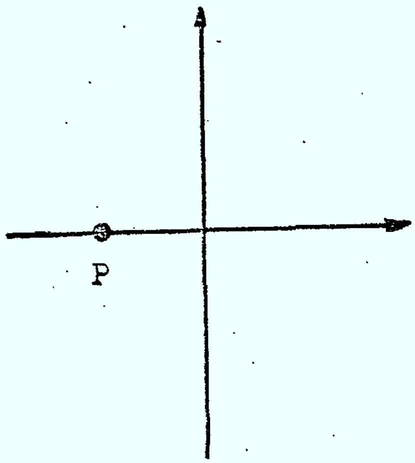


.. OKQPSK

Fig. 3.12 Computer-Simulated Signal Scatter Diagram of QPSK and OKQPSK at the Sampling Instants



QPSK Signal Space Diagram



Fourth-Powered Signal

Fig. 3.13 QPSK and Its Fourth-Powered Signal with $\theta = \pm 45^\circ$, or $\pm 135^\circ$. When fourth powered, this signal becomes,

$$\cos (8\pi f_c t + 4 \theta) \tag{3.22}$$

and $4 \theta = \pm 180^\circ$ or $\pm 540^\circ$, so it converges to one point.

For OKQPSK, the signal states lie on a line (Fig. 3.12), and the fourth powered signal will then lie on the curve as shown in Fig. 3.14. In this figure, it is shown that the vertical line AB of the signal scatter diagram of the filtered OKQPSK (or MSK) signal, when fourth powered, will not only be a point (P in Fig. 3.14) at the four times carrier frequency, but also a continuous curve. The same curve is also obtained when the horizontal line CD is fourth powered. In other words, the fourth-powered signal of OKQPSK at the sampling instants lie on a curve as shown in Fig. 3.14. c.

To demonstrate that this is the case, Fig. 3.14. a is used as an example.

Line AB can be written in polar coordinates as:

$$\overline{AB} = \sqrt{a^2 + y^2} \; \underline{/\tan^{-1} \frac{y}{a}} \tag{3.23}$$

when fourth-powered, it becomes:

$$\overline{AB}^4 = (a^2 + y^2)^2 \; \underline{/4 \tan^{-1} \frac{y}{a}} \tag{3.24}$$

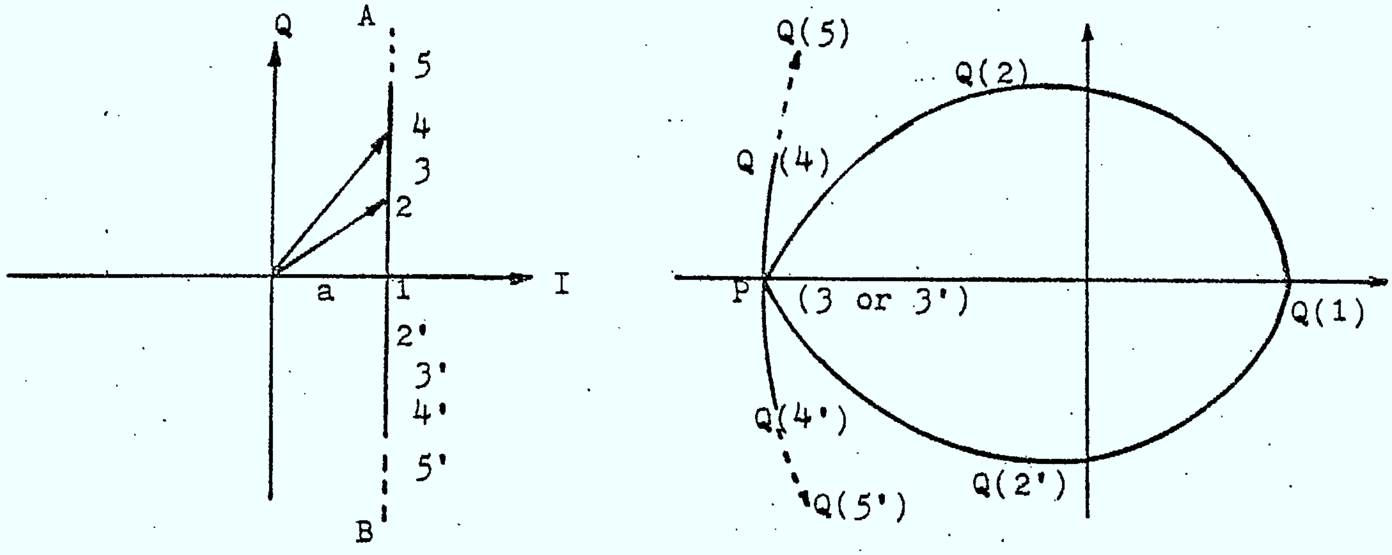
For point 1 on line AB, it can be represented by a $\underline{/0^\circ}$ when fourth-powered, it becomes $a^4 \underline{/0^\circ}$ and is indicated as point Q_1 in Fig. 3.14. a.

For point 3 on line AB, the ordinate y is equal to a. It can be represented by $\sqrt{2}a \angle 45^\circ$. When fourth-powered, it becomes $2 a^4 \angle 180^\circ$ as indicated by P in Fig. 3.14. a. Any point along $\overline{AB}$ between point 1 and point 3 when fourth-powered, lies on the curve $\widehat{P Q_2 Q}$. In a similar way, any point along point 1 and point 3', when fourthed powered, lies on the curve $\widehat{P Q'_2 Q_1}$. Point P is the degenerate point for points 3 and 3'.

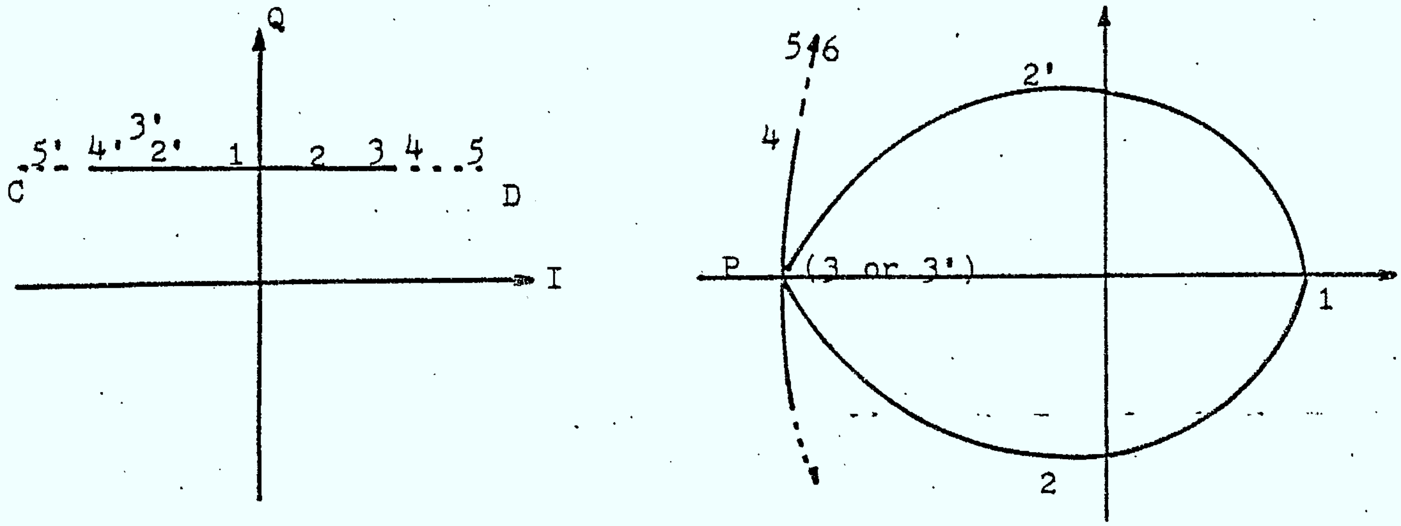
Any point on AB beyond point 4 and point 4', when fourthed powered, can be shown to lie on curve $\widehat{Q_4 P Q_{4'}}$.

This fourth - powered signal can be computed using Eq. (3.19). Due to the variational nature of this signal at the sampling instants, computer simulation was used. Shown in Fig. 3. 14. d, the shape of the computer simulated signal resembled that plotted in Fig. 3.14. c, except that Fig. 3.14. d shows also the distribution pattern of the variation of the fourth-powered signal. Fig. 3. 14. d was obtained by sampling the received signal at the receiver using the simulation programs.

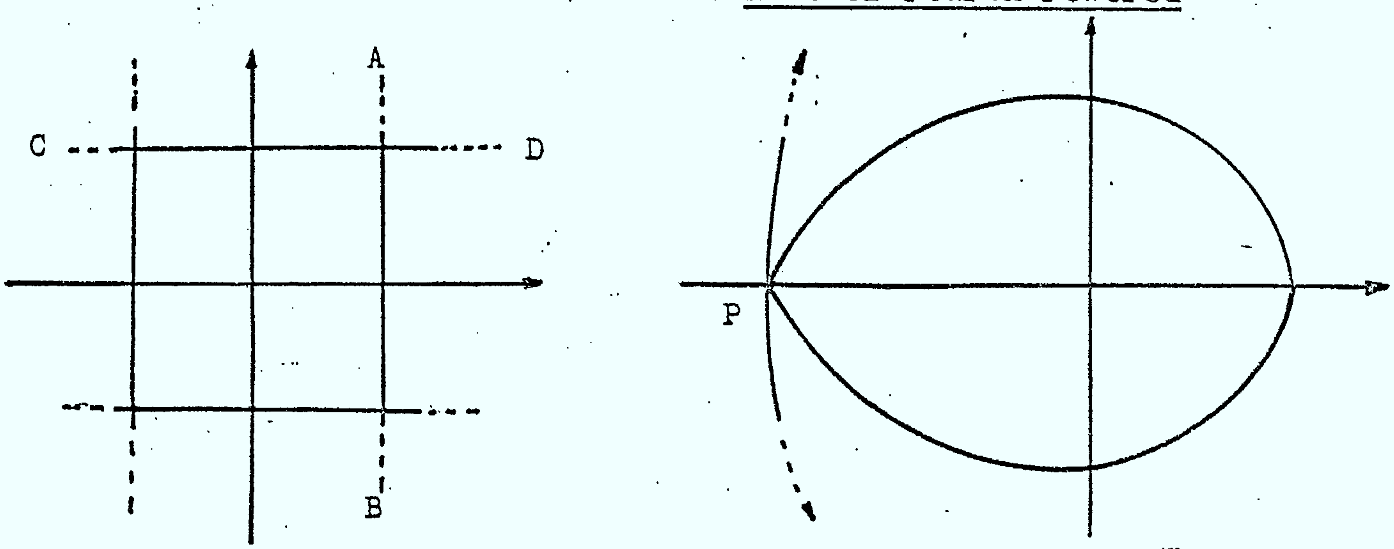
If the shaped pulses are of the (sin x/x) shape, then theoretically this fourth powered signal in OKQPSK can have infinite envelope fluctuations for one channel at the sampling instants of the other channel. This situation occurs when a data pattern like :



a. Line AB Fourth Powered



b. Line CD Fourth Powered



c. Fourth Powered Signal

Fig. 3.14 Fourth Powered OKQPSK

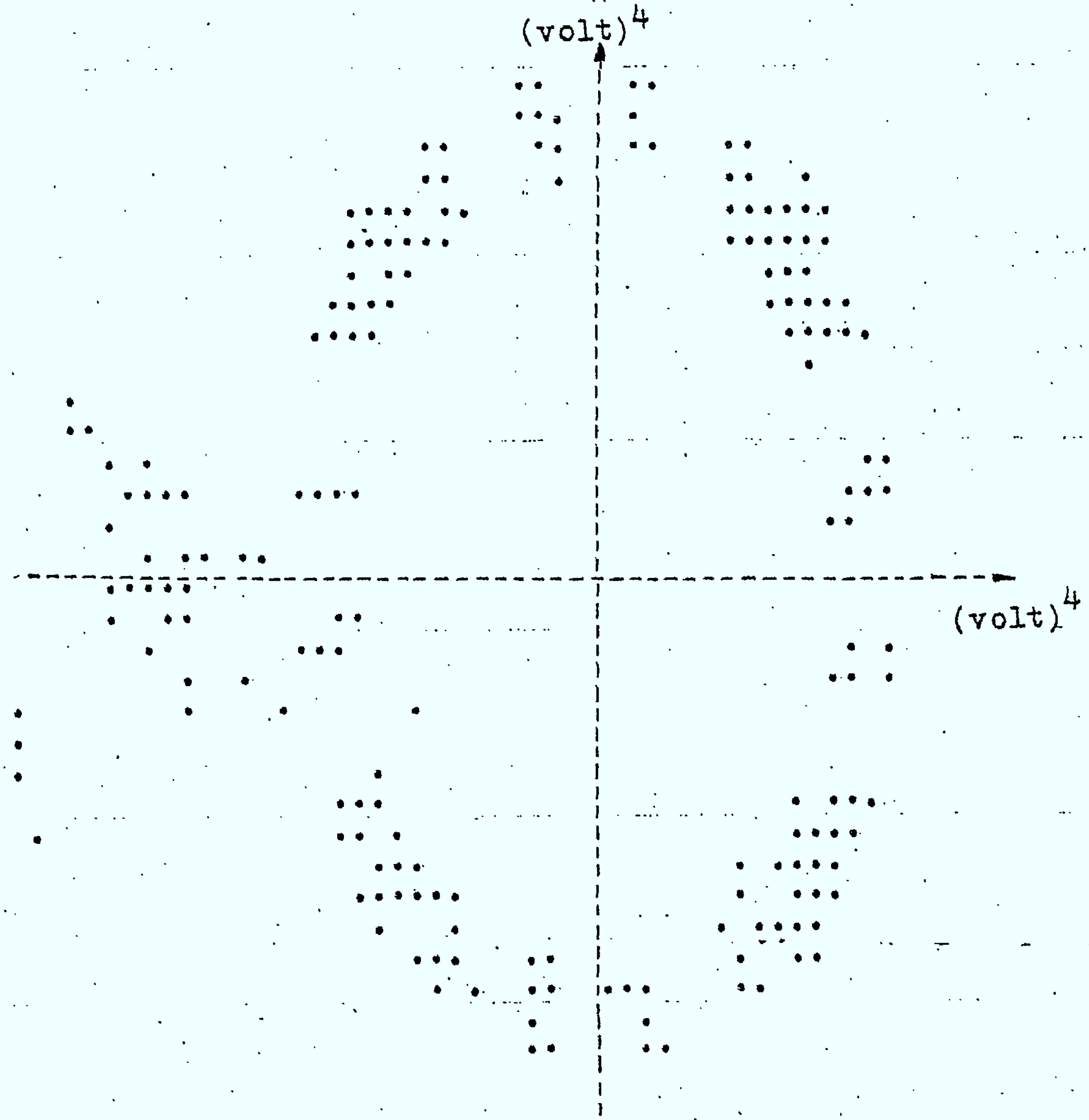


Fig. 3.14.d Fourth-Powered Signal of OKQPSK at the Sampling Instants

$$\ldots\ldots\; \underbrace{0\,1\,0\,1\,0\,1}_{\substack{\text{n pre-}\\ \text{vious}\\ \text{bits}}}\; \underset{\substack{\text{pre-}\\ \text{sent}\\ \text{bit}}}{\overset{\uparrow}{1}}\; 0\,1\,0\,\underbrace{1\,0\,1\,0}_{\substack{\text{n future}\\ \text{bits}}}\ldots\ldots\ldots\ldots \qquad (3.23)$$

arises. For example, for n = 80, the eye pattern has only 17% eye width and the envelope fluctuations are very large (varying along lines AB or CD) for channel at the sampling instants of the other channel (Fig. 3.15).

If n is assumed infinite, then this variation can become the infinite lines (the dotted portion) shown in Fig. 3.14. When fourth powered, it can scatter to infinity as shown by the dotted line of Fig. 3.14.d.

Hence, even in the absence of noise and intersymbol interference, the OKQPSK (or MSK) signal can have inherently very large amplitude fluctuations at the sampling instants. In Ref. [13], the relative amount of phase jitter of the recovered carrier for these three types of signals was obtained using computer simulation. These results show that the phase jitter of the recovered carrier in QPSK is indeed much smaller than that of OKQPSK (and MSK) in a narrow-band channel.

(Lines AB and CD as shown in Fig. 3.14.c)

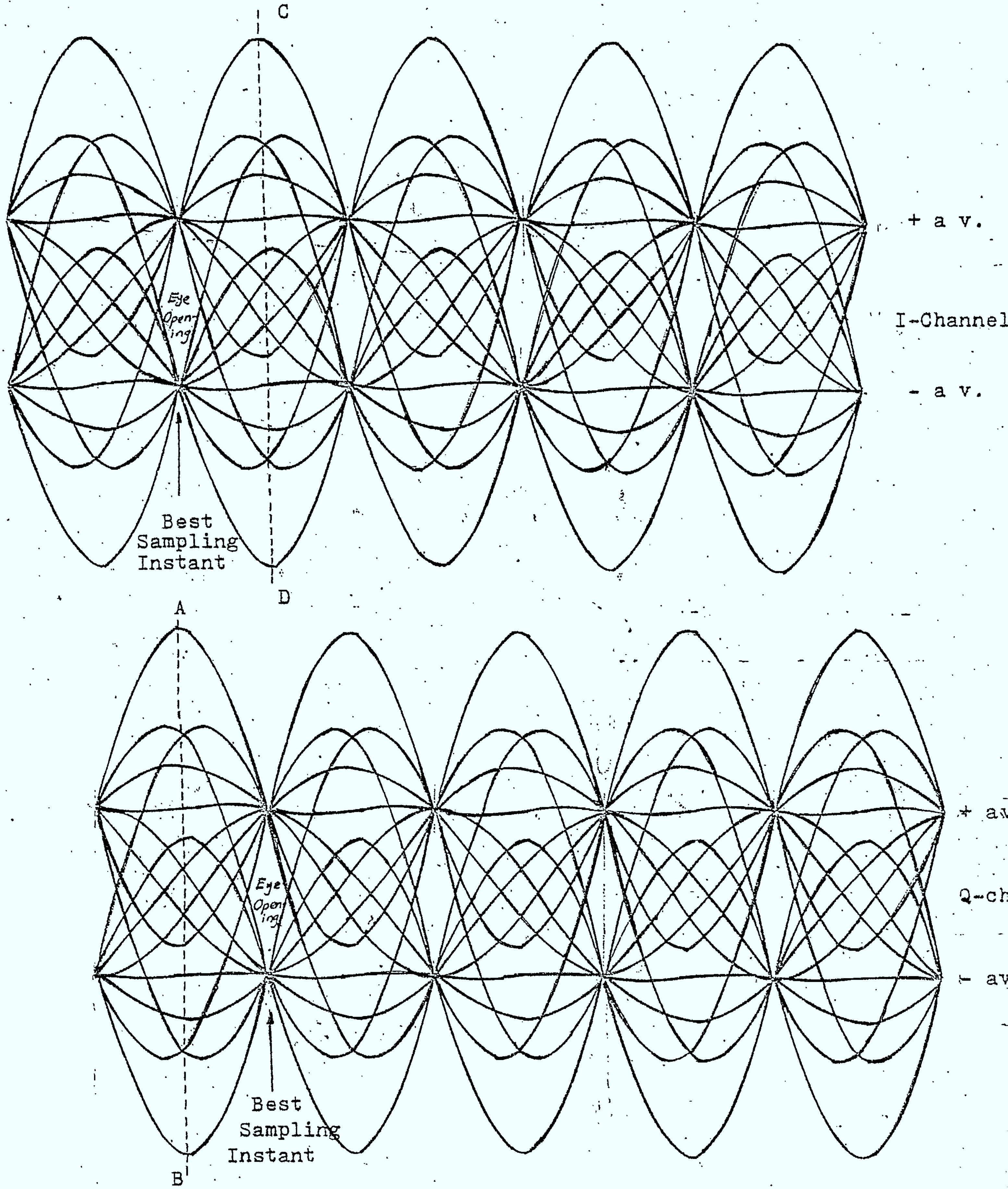


Fig. 3.15 I- and Q-Channel Eye Diagrams when (sin x/x) Pulses are Used as Signaling Elements in OKQPSK

3.5 SPECTRUM SPREADING

Two types of baseband signals can be considered [1]. The first one contains certain amount of data pattern jitter while the second one is jitter free. In this section, we will study spectrum spreading which occurs when these two types of signals are modulated separately by a carrier as in Binary PSK (BPSK) and then hardlimited.

Type 1: Spectrum Restoration

Fig. 3.16 shows a filtered random data being modulated by a carrier as in the Binary PSK. At the keying instants, phase transitions of 180° occur for both the unfiltered and filtered BPSK signal (Figs.3.16.a and 3.16.b) [14]. At these 180° phase transitions, the filtered BPSK signal has zero amplitude as the signal vector is shifted through the origin.

When the signal shown in Fig. 3.16. b is passed through an ideal limiter, the limited signal (Fig. 3.16. a) is exactly identical to that of the unfiltered signal. The filtered jitter free and limited signal therefore, has the same infinite bandwidth spectrum as that of the original

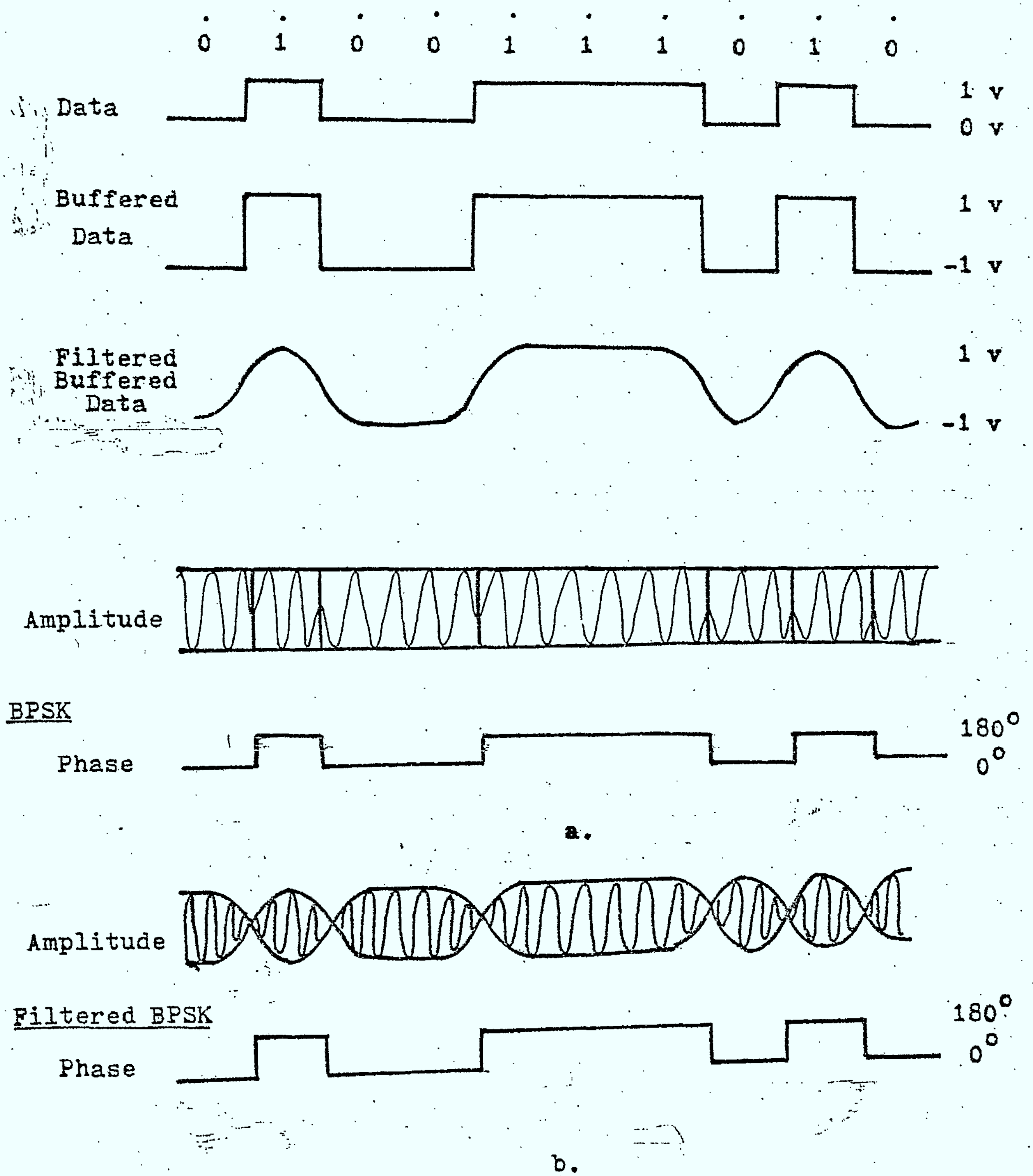


Fig. 3.16 Filtered and Limited Waveforms of BPSK

unfiltered signal. In other words, the spectrum is completely restored.

To demonstrate that this is indeed the case, the following experiment was carried out.

Type 2: Spectrum Regrowth and Modification

If a random data stream which has almost zero ISI but with a certain amount of jitter was also modulated by a carrier and then hardlimited, the spectrum of the limited signal in this case, shows that in addition to the restoration of sidelobes, the nulls disappear from the modified spread spectrum.

Spectrum modification can thus be partly attributed to the timing jitter of the signal, although a large part of spectrum spreading is still caused by hardlimiting.

Since QPSK is a quadrature combination of two BPSK signals, the same reasoning also applies. This means that filtered QPSK signal, after passing through a nonlinear device, will also induce spectrum spreading. Spectrum spreading of filtered QPSK signal as well as that of filtered OKQPSK and MSK signals have been considered in great details in References [11], [15], [16] and [17].

3.6 PERFORMANCE STUDY OF BANDLIMITED QPSK, OKQPSK AND MSK SIGNALS THROUGH CASCADED NONLINEARITIES

As discussed earlier, QPSK, OKQPSK and MSK are the three modulation schemes most frequently considered for a nonlinear satellite communication channel. The main advantages of using these three modulation techniques are that they are bandwidth efficient (theoretically, 2 b/s/Hz, practically about 1.8 b/s/Hz) and that they have low interchannel interference susceptibility [12]. Also the hardware design of QPSK is relatively simple while that of OKQPSK and MSK, even though slightly more involved, is not very complex.

Since their inventions in the 1960's [18, 19, 20], both OKQPSK and MSK have been proposed, as an alternative to QPSK, for use in bandlimited satellite channels. The main reason cited is that either OKQPSK or MSK, when bandlimited and then passed through a hardlimiter, has less spectrum regeneration than QPSK [11], [15], [21], [22].

MSK, as illustrated earlier in section 3.2, can be considered one form of OKQPSK with half cosine pulse shaping. Due to its inherent wider main spectral lobe, it is generally agreed that in narrowband channel environment, the performance of MSK degrades more than that of OKQPSK. For example, in Ref. [5], Gronemeyer and Mcbride showed that for $BT < 1.1$, OKQPSK has an advantage over MSK while for $BT > 1.1$, the performance of MSK is superior to that of OKQPSK. (B is the noise bandwidth of the two cascaded filters used in their simulation). In Ref. [22], Constellano concluded that the best way to utilize the spectral characteristics of MSK is to use wideband filters. A similar conclusion was also presented by Murakami et al. [13] and Lundquist [23].

The comparative performance of QPSK and OKQPSK, on the other hand, has not yet reached a consensus. As mentioned earlier, it is generally agreed that a filtered OKQPSK signal induces less spectrum spreading than a filtered QPSK signal after each passing through a saturated nonlinear power amplifier. Yet, for overall system performance comparison, other factors, beside spectrum regeneration, must also be considered.

Constellano [22] in his simulation study showed that, based on the trade off between in-band degradation and adjacent channel interference, OKQPSK requires less transmitter power than QPSK for the satellite up-link subsystem. His study concluded that OKQPSK is more resistant to carrier phase offset and group delay distortions than QPSK, but less resistant to linear amplitude distortions. In Ref. [24], Gilton and Ho also demonstrated that OKQPSK has an improved phase jitter immunity than QPSK.

On the other hand, Murakami et al ,[13] showed that the error rate performance of QPSK is superior to OKQPSK . They also demonstrated that phase jitter of the recovered carrier in QPSK is smaller than that of OKQPSK. In the study of the European Communication Satellite (ECS) 120 Mb/s system, Harris [25] presented the rationale in choosing QPSK over OKQPSK. In the narrowband channel environment of ECS, OKQPSK appears to degrade more from nonlinear distortion than QPSK. Lundquist

[23], in his study for the Intelsat V's 60 Mb/s satellite system through a 36 MHz bandwidth channel, also concluded that QPSK outperforms OKQPSK in a nonlinear channel. A similar conclusion was also reached by Chakraborty et al [26].

The reason that OKQPSK suffers more phase jitter in the recovered carrier is its large amplitude fluctuations at the sampling instants. This was explained earlier in section 3.4.

QPSK, on the other hand, has less phase jitter in its recovered carrier. Even though QPSK induces more spectrum spreading than OKQPSK when passing through a saturated nonlinear power amplifier, its overall performance is still better than that of OKQPSK. By properly designing the QPSK modem in the linear mode and by operating the HPA and TWT input power a few dB below saturation, a compromising performance for QPSK can be achieved. In Ref. [26], Chakraborty et al. concluded that the optimum HPA and TWT operating points for QPSK are at about 6 and 4 dB input backoff, respectively.

3.6.1 Computer Simulation Model

In the following, the computer simulation used in our study will be described. Afterwards, results obtained from this simulation study will be presented. As few details were discussed in the available literature, comments and comparison with other computer simulation methods are somewhat difficult. The conflicting results obtained by Constellano [22], versus those given by Harris [25] and Lundquist [23], all sponsored by the European Space Agency, typify the complexity involved in the simulation study of a nonlinear satellite channel. Nevertheless, an attempt will be made to clarify the method used in our study.

In the linear channel, it is easier to study the system performance by using a computer analysis program, in which the system is excited by applying a single pulse of duration T in the in-phase baseband channel. As the data stream in the quadrature channel is statistically independent of that in the in-phase channel, there is no need to use both quadrature channels in this method.

By the use of the Fast Fourier Transform (FFT), the single-pulsed signal is modified by the frequency responses of the various filters in the systems. Since the system components are all linear, superposition may be used to compute the peak and rms eye closure for a random data sequence. This is because the power spectrum of the single isolated pulse has the same shape as that of a random data stream.

However, this computer analysis program cannot be adopted in a nonlinear satellite channel since the principle of superposition is no longer valid. Hence, a simulation program in which a pseudo-random data sequence is generated must be used. In a similar way, this random data sequence, by means of the FFT algorithm, is alternated between its time-domain and frequency-domain representation. In the frequency domain, it is modified by the complex transfer function of the simulating filters. In the time-domain, it is modified by the nonlinearities of the two cascaded power amplifiers.

Based on the simulation model of Fig. 3.1, Fig. 3.19 shows the flow chart of the signal processing in our simulation study. The signaling format for QPSK, OKQPSK and MSK was obtained in the equivalent complex baseband form (Fig. 3.2) from a pseudo-random data source. After Fourier transformation, it was modified by the frequency response of filter F_1. It was then inverse transformed back in the time domain and modified by the polynomial coefficients of the quadrature model of the HPA [27]. The description of the quadrature model for a nonlinear power amplifer such as HPA and TWT is contained in chapter 2.

The distorted signal after the HPA was again Fourier transformed and modified by the frequency response of filter F_2. Inverse Fourier transformed back into the time domain, this signal was again modified by the polynomial coefficients of the quadrature model of the TWT. The signal, after modification by the frequency responses of filters F_3 and F_4, was then decoded at the receiver and compared with the original transmitted data sequence.

White Gaussian noise was added to the signal at the receive filter input (F_4 in Fig. 3.19). The total noise power at the output of F_4 was then computed. The up-link noise was not included in our simulation as it was quite involved if noise was to be simulated [8].

Different filtering strategy has been used in the various simulation studies quoted earlier. In the linear channel, the overall system response is usually assumed to be raise cosine. Optimum performance is attained whenwith raised cosine filtering is partitioned equally between the transmitter

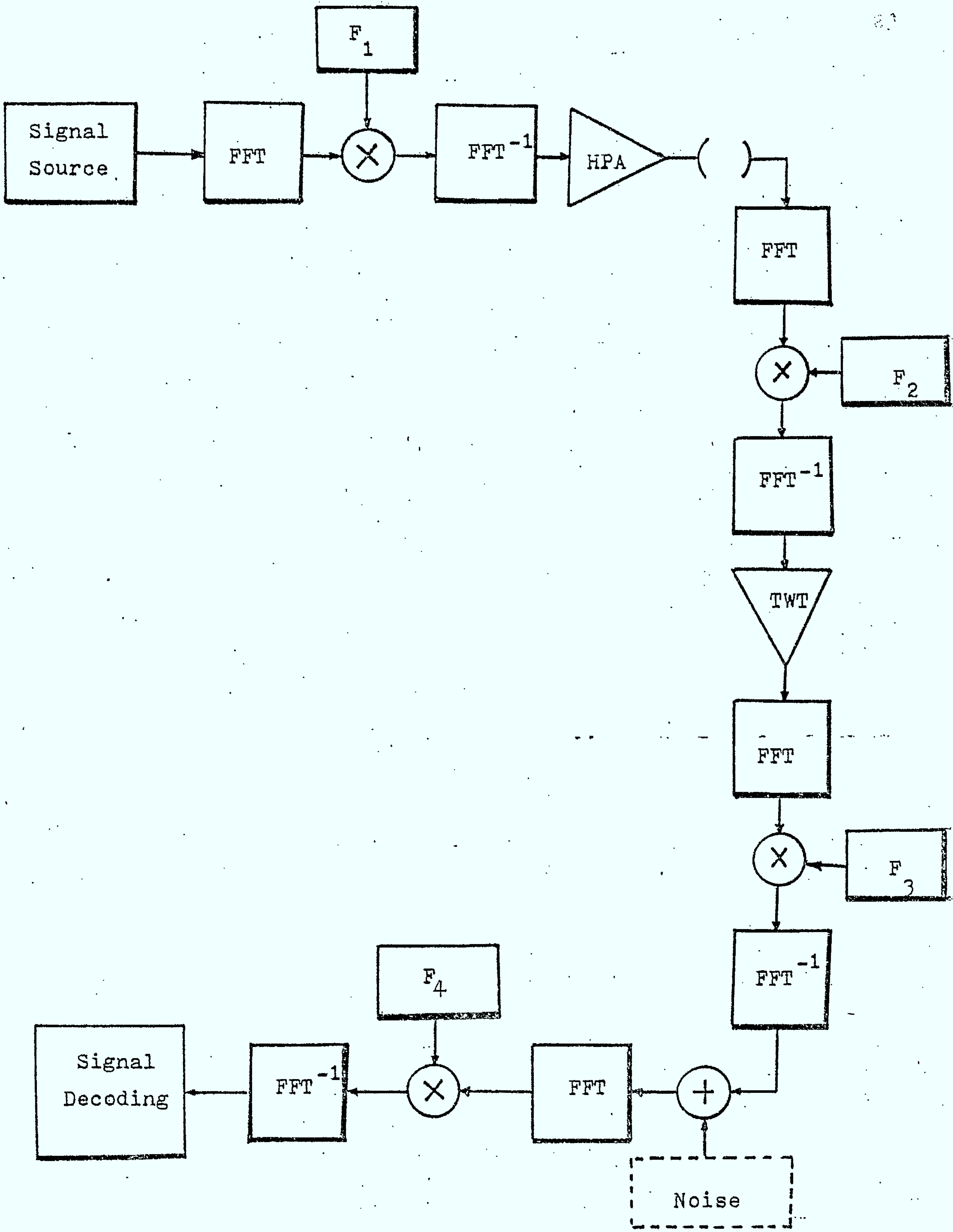


Fig. 3.19 Signal Processing of Computer Simulation

and the receiver. The situation became more complicated in a nonlinear channel. This is because in the actual satellite communication systems, the earth transmitting station and receiving station sometimes might belong to two different countries, with equipment supplied by different manufacturers. In Ref. [13], the specifications for the transmit and receive filter are simply divided with a sharp cut-off filter at the transmitter and a Nyquist shaping filter with an operature equalizer at the receive side. In Ref. [23], a rolloff factor between 40% and 50% with close to half of Nyquist shaping at the transmitter was employed, while in Ref. [26], a Nyquist shaping filter at the transmitter and an elliptic filter at the receiver were assumed.

The computer programs developed in this study, in their present format, only have the capability to evaluate the single-channel performance of QPSK, OKQPSK and MSK signals through a nonlinear satellite channel. A simulation program capable of evaluating performance of these three signals in a multi-channel environment may become quite involved. Not only the carrier information for the multi-channel information must be included in the simulation, but also an optimization routine having iterative capability must be used to find the optimum filtering parameters.

In our study, a simple approach using a typical filter, namely a 4 pole Chebychev with 0.5 dB passband ripple, was adopted. This filter with different 3 dB bandwidth was assumed for F_1, F_2, F_3 and F_4 in our simulation model shown in Fig.3.1. In Ref. [8], Chan et al also adopted this simplified filtering approach by assuming the transmit and receive filters to be of Chebychev type. Similarly in Ref. [28], Devieux in his QPSK satellite link simulation assumed all the four filters to be of the Butterworth type.

3.6.2 Computer Simulation Results

To check the accuracy of the simulation programs, the wideband performance of these three modulation schemes was obtained and compared as shown in Fig. 3.20. In this wideband model, the BT product of F_1 and F_4 (where B is the double sided 3 dB bandwidth of the Chebchev filters and T is the symbol interval) is equal to 2, while the BT product of F_2 and F_3 is about 2.5. In this case, the OKQPSK performs slightly better than the QPSK and MSK signal. By further increasing the BT product of each of these four filters, all these three types of signals require a SNR of approximately 8.8 dB for a BER of 10^{-4}. In comparison, the theoretical SNR requirement with the noise in the bit rate bandwidth for 10^{-4} is 8.4 dB for these three types of signals.

The narrowband performance of these three signals was then evaluated by reducing the BT product of F_1 and F_4 first to 1.3, and then to 1, while the BT product of F_2 and F_3 was kept about 2.5. This corresponds to the narrowband model used by Devieux [28] in his study for the performance of QPSK through two cascaded nonlinearities. In our simulation study, QPSK performs approximately the same as in Devieux's work. Results in Fig. 3.20 also show that the performance of OKQPSK becomes worse than that of QPSK when the BT product of F_1 and F_4 was reduced from 1.3

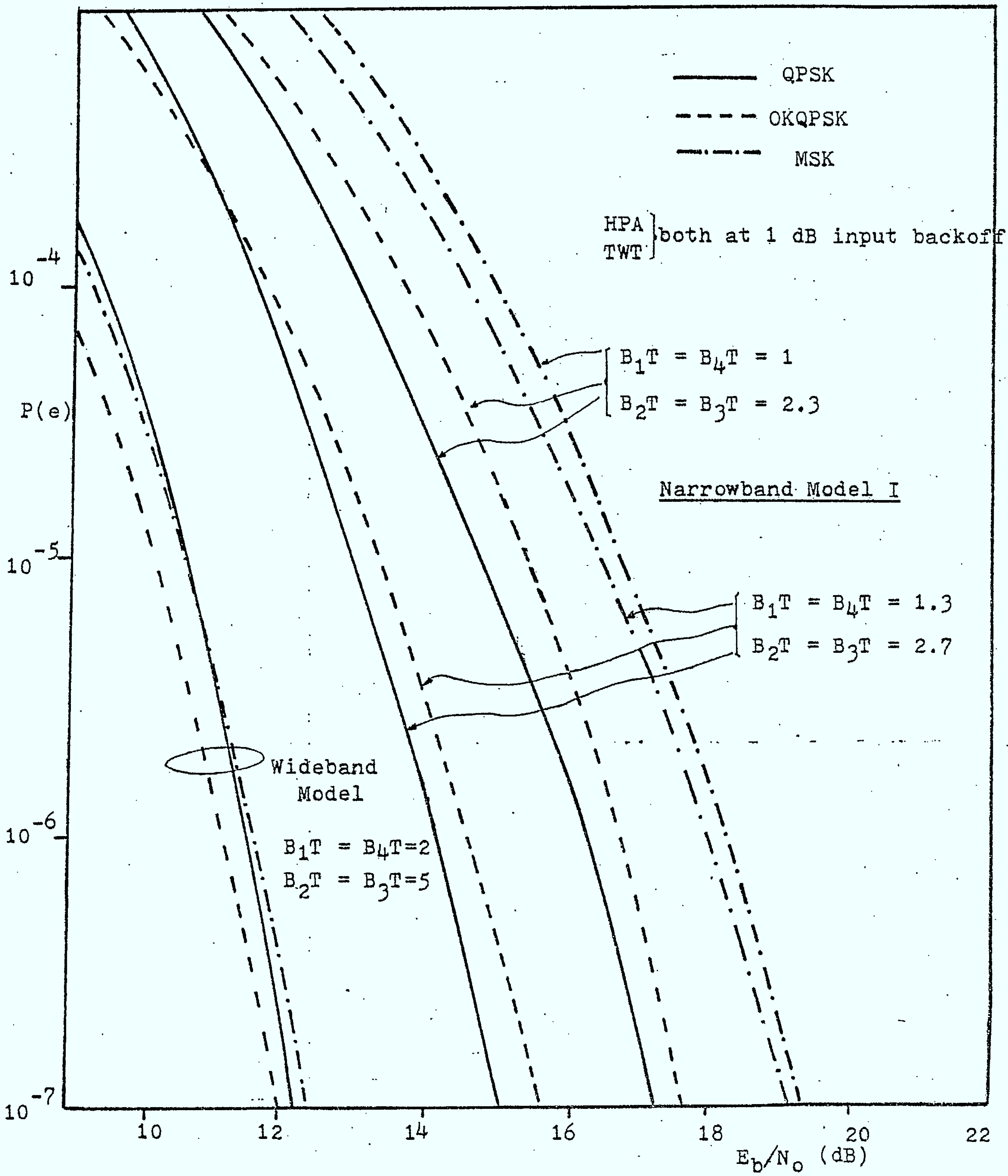


Fig. 3.20 Performance of QPSK, OKQPSK and MSK Signals in the Wideband and Narrowband Model I

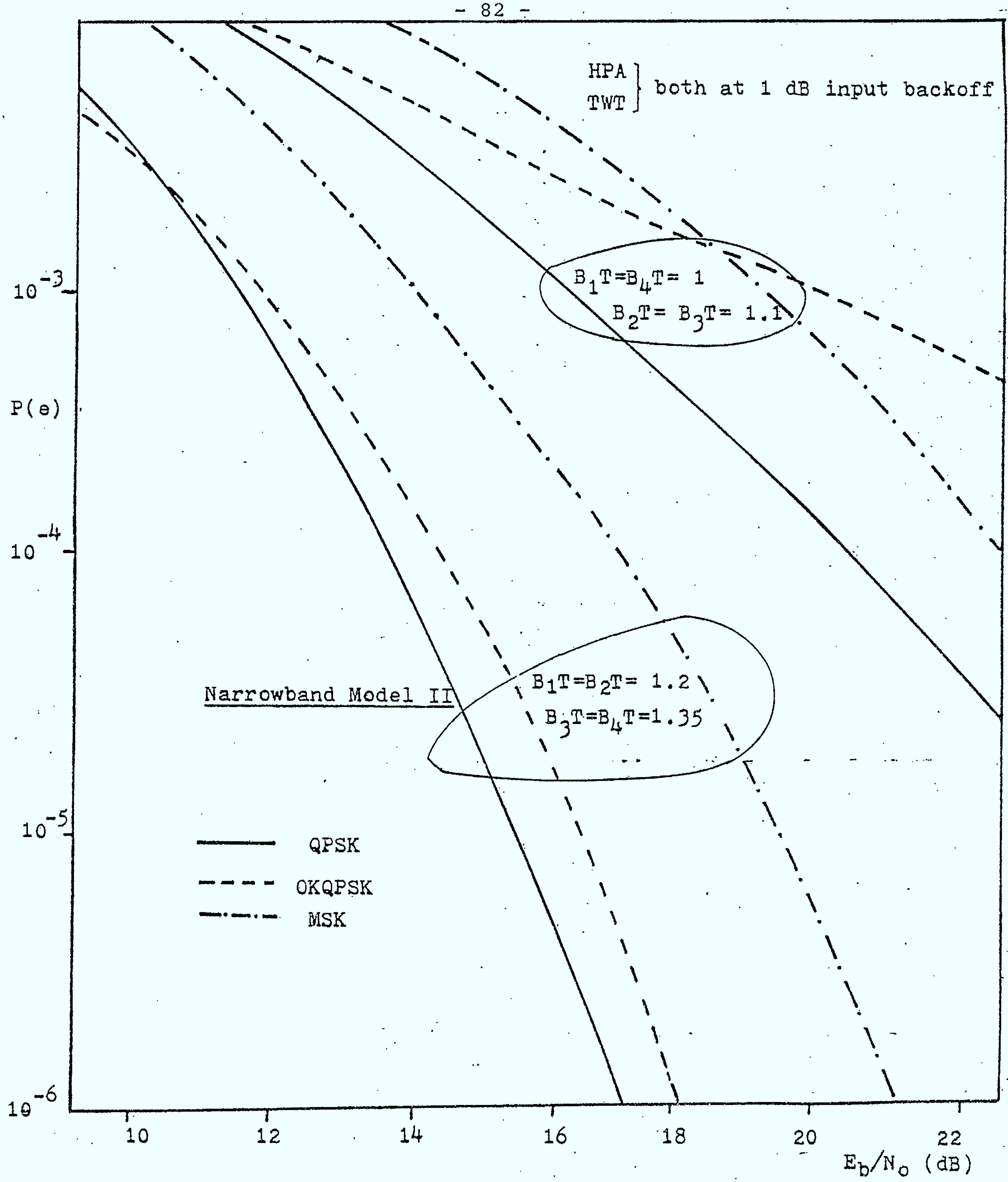


Fig. 3.21 Performance of QPSK, OKQPSK and MSK Signals in Narrowband Model II

to 1. In either situation, the performance of MSK is worse than that of either OKQPSK or QPSK.

In the above model, the bandwidths of F_2 and F_3 were made sufficiently wide so that the cascaded nonlinear characteristics of HPA and TWT can be considered as one single nonlinearity. In some satellite channels, the bandwidths of these two filters are also narrowband so as to bandlimit the up-link thermal noise and to reduce the adjacent channel interference.

For this second narrowband model, the BT products of F_1 and F_4 was again kept at either 1.3 or 1, while the BT products of F_2 and F_3 was reduced from 2.5 to 1.1. Results in Fig. 3.2 again show that QPSK performs better than either OKQPSK or MSK in this narrowband model.

Results shown in Fig. 3.20 and Fig. 3.21 are for the HPA and TWTA both operating at 1 dB input backoff. When they both were operated at 12dB input backoff, results in Fig. 3.22 again show that QPSK outperforms OKQPSK in the narrowband model 11. In the narrowband model 1, results plotted in Fig.3.22 show that in one situation ($B_1T = B_4T = 1.3$, where B_1 and B_4 denote the f_{3dB} bandwidth of filters F_1 and F_4), OKQPSK performs slightly better than QPSK. In the second situation, ($B_1T = B_4T = 1$), QPSK performs better than OKQPSK. In the wideband model, OKQPSK performs slightly better than QPSK.

In our earlier study, it was shown that OKQPSK and MSK have much larger amplitude variations at the sampling instants than QPSK. After passing through the nonlinear amplifiers, in QPSK, the sampled signal amplitude is still concentrated about a fixed point. This results in a smaller

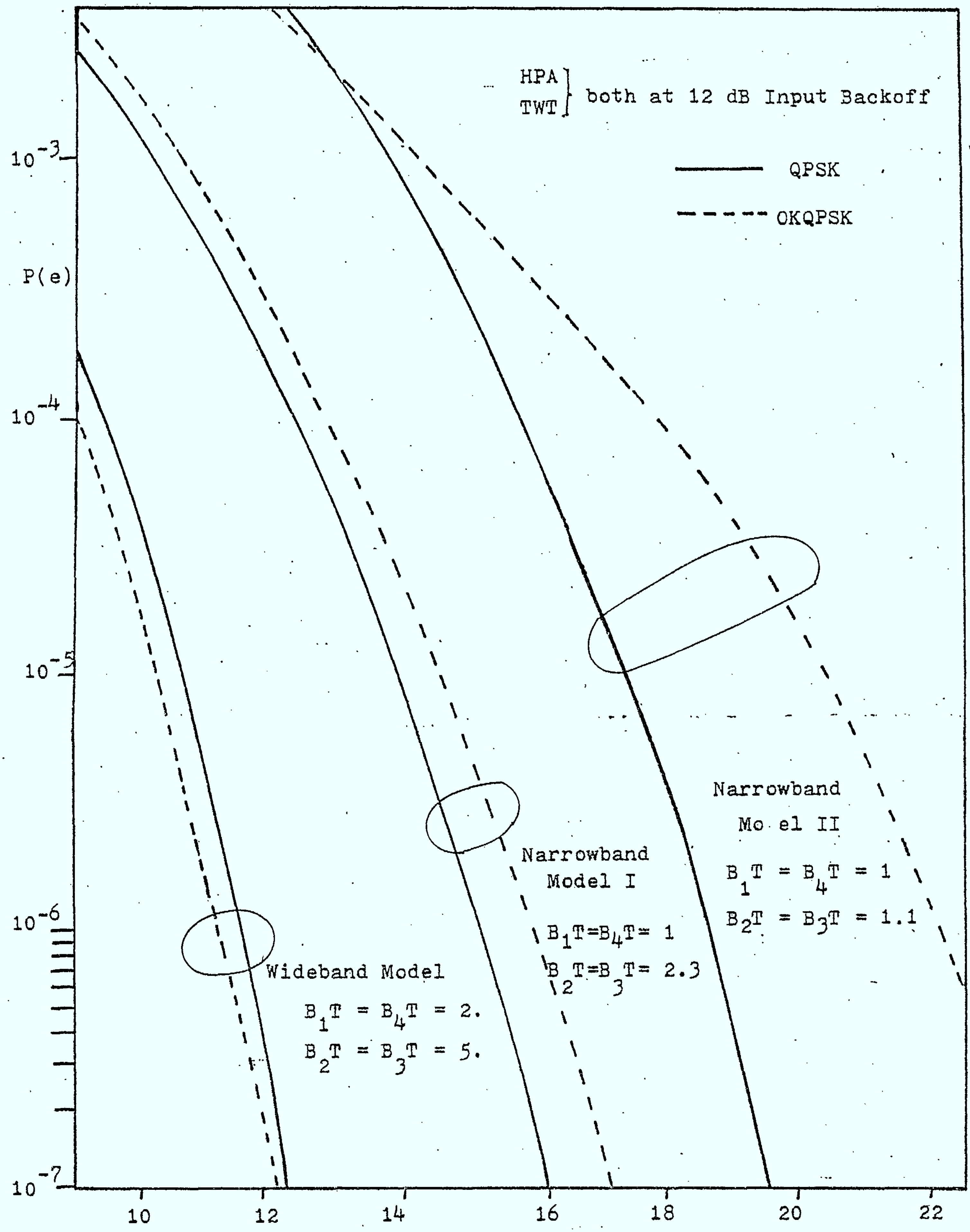


Fig. 3.22 Performance of QPSK, OKQPSK and MSK in Three Models

degree of phase jitter. The coincidence of the sampling points between the in-phase and quadrature channels also induce less cross-modulation interference. On the other hand, the much larger amplitude variation of these two signals at the sampling instants and the misalignment of the sampling instants between the two quadrature channels cause a larger degree of signal scattering at the sampling instants. Thus, in a severely band-limited channel, OKQPSK or MSK has a much larger phase jitter than QPSK, resulting in larger overall system performance degradation.

As there are many constraints and parameters involved in the simulation model of a nonlinear satellite channel, the argument in favor of or against the use of OKQPSK versus QPSK still needs further study. Only through practical field testing and measurement, can this problem be solved. Nevertheless, the simulation method and results obtained in this study, after further extension, will help lead to a more favorable answer in the near future.

3.7 SUMMARY

An analysis to study the amplitude fluctuations, both the overall average and at the sampling instants, of bandlimited QPSK, OKQPSK and MSK signals, was undertaken. Although a filtered OKQPSK (or MSK) signal has less overall amplitude fluctuations than QPSK, it has much larger amplitude fluctuations at the sampling instants. Thus the overall performance of a filtered OKQPSK signal degrades more from nonlinearities than that of QPSK. Both the signal space and scatter diagrams of QPSK, OKQPSK and MSK were verified by computer simulation.

Two types of spectrum spreading that occur when a filtered PSK signal with or without jitter passes through a hardlimited were experimentally demonstrated. The first type of spectrum restoration occurred when a jitter free signal was hardlimited. As there is no timing jitter in the PSK signal, complete spectrum restoration can thus be reasoned as caused entirely by hardlimiting alone. The second type of spectrum restoration and modification arose when a jittery signal was hardlimited. Spectrum modification can thus be partly attributed to the timing jitter that exists in the filtered signal.

By the use of computer simulation, the performance of QPSK, OKQPSK and MSK through cascaded nonlinearities and bandlimiting elements as in a typical satellite channel in a wide-band model and two narrowband models was evaluated. Results from the wide-band model show that OKQPSK performs slightly better than conventional QPSK and MSK. On the contrary, results from the two narrowband models show that conventional QPSK outperforms OKQPSK and MSK.

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CHAPTER 4

SIMULATION COMPARISON OF REGENERATIVE AND NON-REGENERATIVE SATELLITES

4.1 SIMULATION COMPARISON OF SYSTEMS

In this section of the report a set of simulations have been performed in order to have a comparison between regenerative and non-regenerative satellites. Due to the limited amount of time available for the development of simulation programs, only QPSK modulation has been considered. The bit rates used in the simulation are non standard but serve for illustrative purposes in this first phase of the project. As an initial step towards a more extensive study, the consideration first of QPSK is the logical direction to take. There are a few reasons for this: first, the simulation of QPSK is relatively simple, when compared to OKQPSK, MSK or differentially encoded schemes [1,2]; the relative spectral efficiency of QPSK in a linear channel is quite good, i.e., up to 2 bit/s/Hz [3]; the performance of QPSK appears best among the different modulation schemes [4]. With these points in favor of QPSK, it was decided to place the emphasis on the study of QPSK.

To evaluate the performance of both a conventional repeater satellite link and a regenerative repeater satellite link, these two systems have been simulated.

The simulation of a conventional link is based on the work by J.C. Huang [2] and was documented earlier in this report.

The simulation of a regenerative repeater link is based on the hypothesis that the uplink and downlink can be considered as two independent links. The analysis of a regenerative repeater satellite system thus essentially becomes the analysis of a two-hop system. The probability of error is given by

$$P(e) \simeq P_u + P_d$$

In other words, the probability of error is essentially the sum of the individual error probabilities, where P_u and P_d respectively stand for the uplink and downlink error probability. The simulation of a regenerative repeater link then consists of simulating two single-hop systems and adding their respective error probabilities. The model being simulated is illustrated in Figure 4.1. Because of the linearity of the uplink and downlink channel, the transmit and receive filters F_t and F_r were lumped as single filters in the actual numerical simulation.

The probability of error curves obtained for the two systems illustrate their relative performance. The curves shown in Fig. 4.2 illustrate the performance of QPSK in a conventional repeater link at bit rates of 84, 92, and 100 Mbits/s with Chebychev filters having 0.1 dB ripple and a 3 dB bandwidth of 54 MHz at both ground transmit and receive stations and with similar Chebychev filters having a 3 dB bandwidth of 60 MHz for the two transponder filters. This corresponds to an effective channel spacing of 60 MHz. Thus, the bit rates used represent, respectively, spectral efficiencies of 1.4, 1.53, and 1.67 bits/s/Hz. From the curves of Figure 4.2, we observe that to maintain a $P(e) = 10^{-4}$,

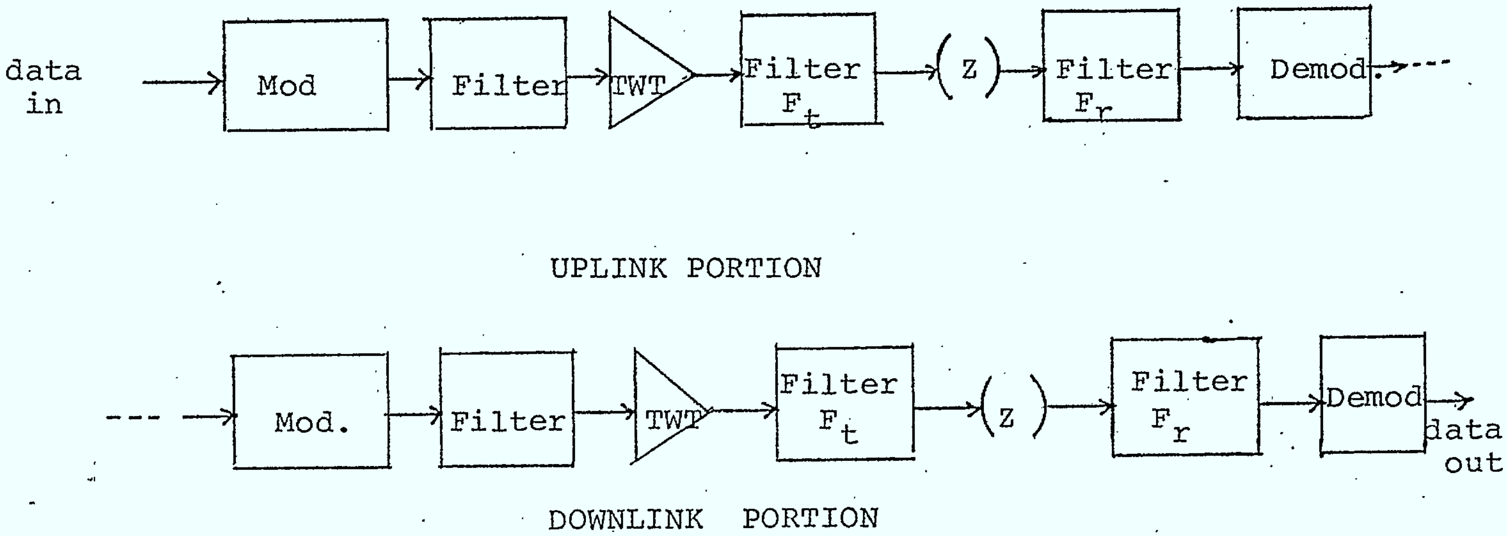

Fig. 4. 1 Regenerative Repeater Link Model

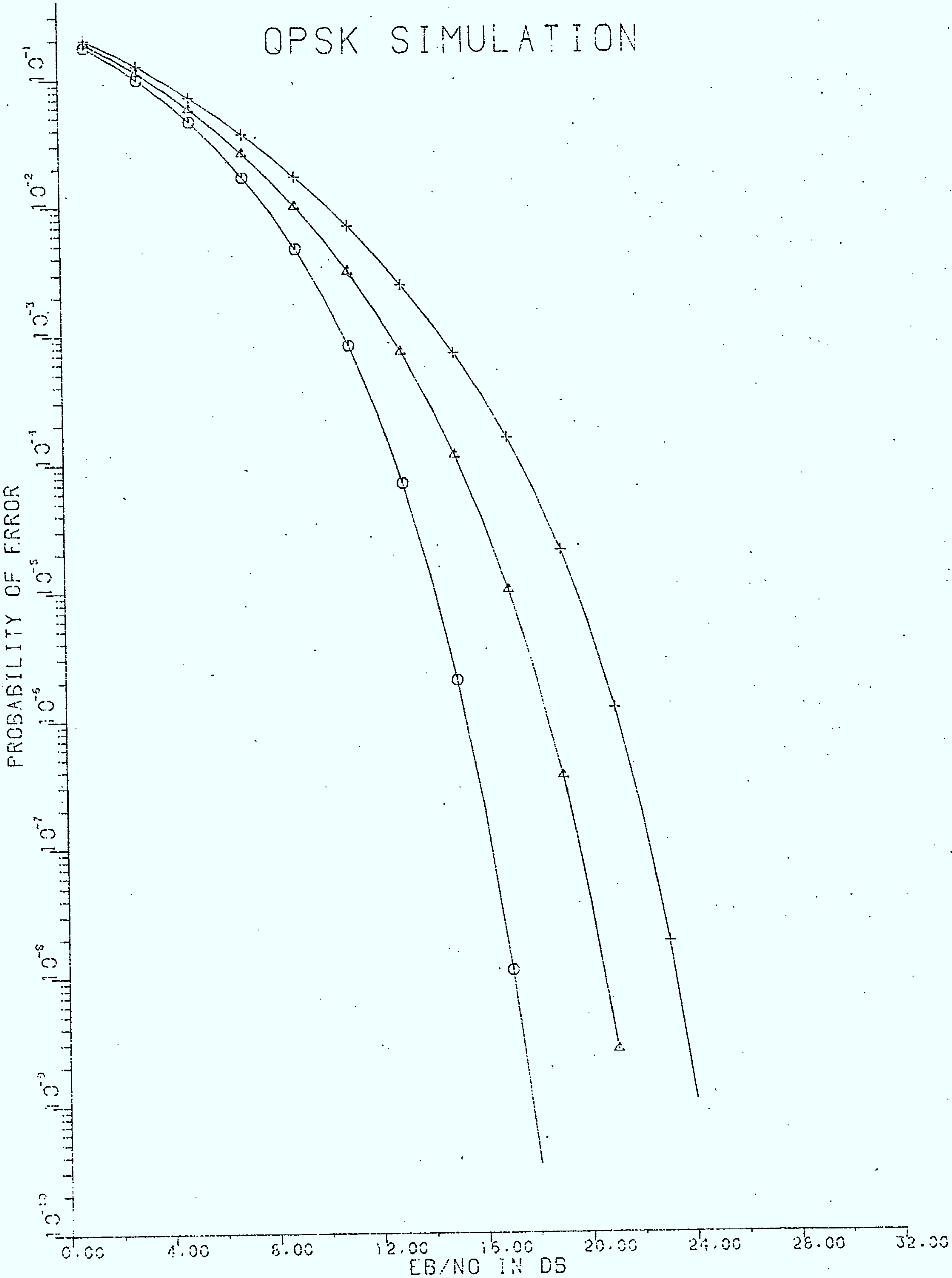
OPSK SIMULATION
PROBABILITY OF ERROR
EB/NO IN DB
0.00
4.00
8.00
12.00
16.00
20.00
24.00
28.00
32.00

Fig. 4.2

approximately 4.5 dB more is required to transmit 100 Mbits/s instead of 84 Mbits. This increase in the required E_b/N_o is due partly to the increased ISI because of the more severe bandlimiting and also partly due to the larger envelope fluctuations, resulting from the bandlimited signal, which are processed by the nonlinear amplifiers.

To see the improvement obtained by regeneration, we refer to Figure 4.3 where, with the same filters QPSK was simulated at a bit rate of 100 Mbits/s. The curve obtained certainly indicates a considerable gain to be obtained from regeneration. Both simulations were performed assuming 12 dB of input back-off at both the ground station and the satellite transponder.

Based on the data obtained the results have only limited value because for one, the results for regeneration appear somewhat optimistic and in general the ground station amplifier and the satellite transponder amplifier are not necessarily operated at the same back-off. However, for illustration purposes, if we assume that both the uplink and the downlink have the same probability of error, we obtain a regeneration gain versus probability of error as shown in Figure 4.4. As we can see from this graph, the gain can be as high as 8 dB which appears too good. Thus more research in this area is required to arrive at conclusive results. This will be included in Phase II.

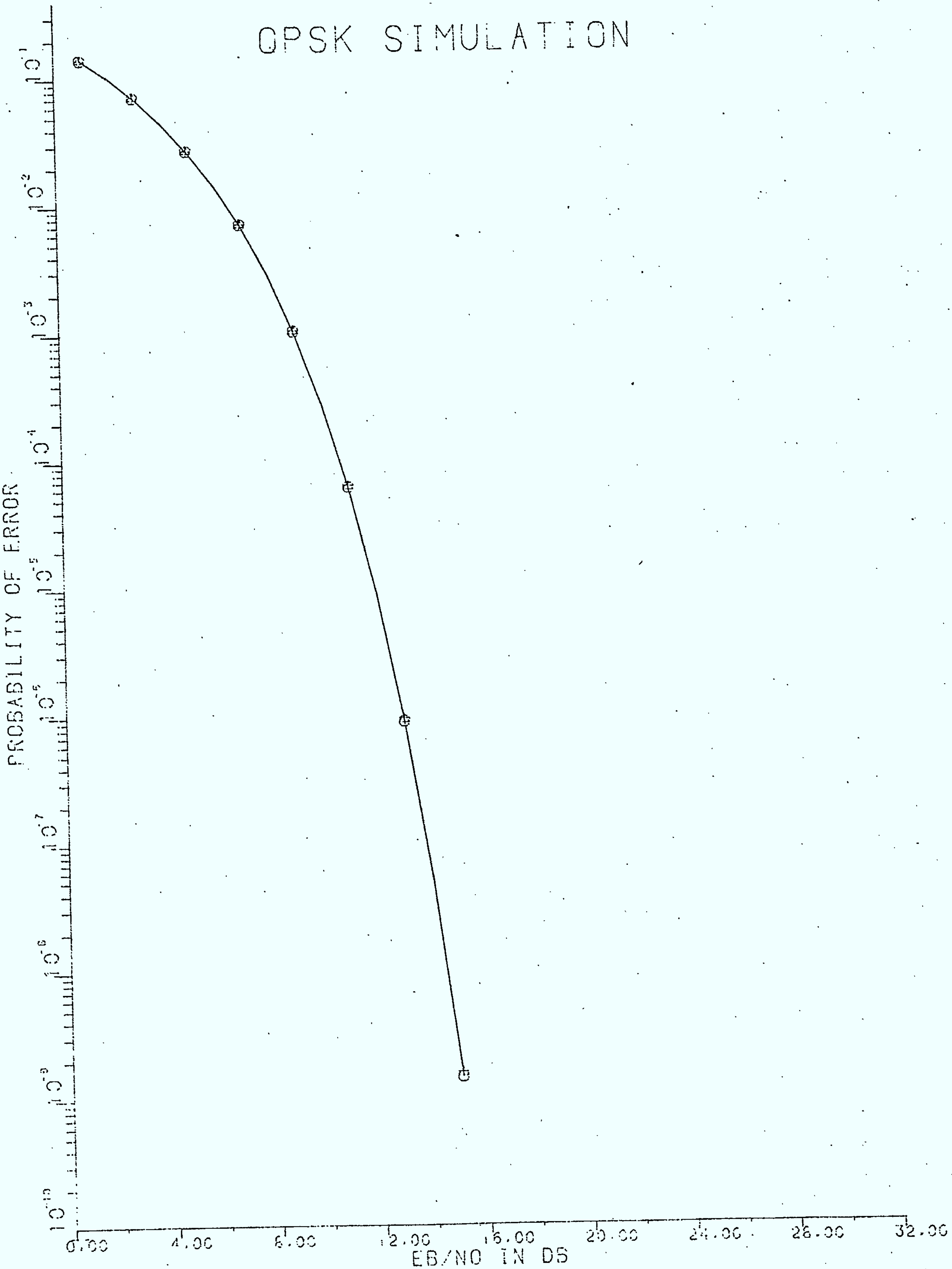
QPSK SIMULATION
PROBABILITY OF ERROR
EB/NO IN DB
0.00
4.00
8.00
12.00
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Fig. 4.3

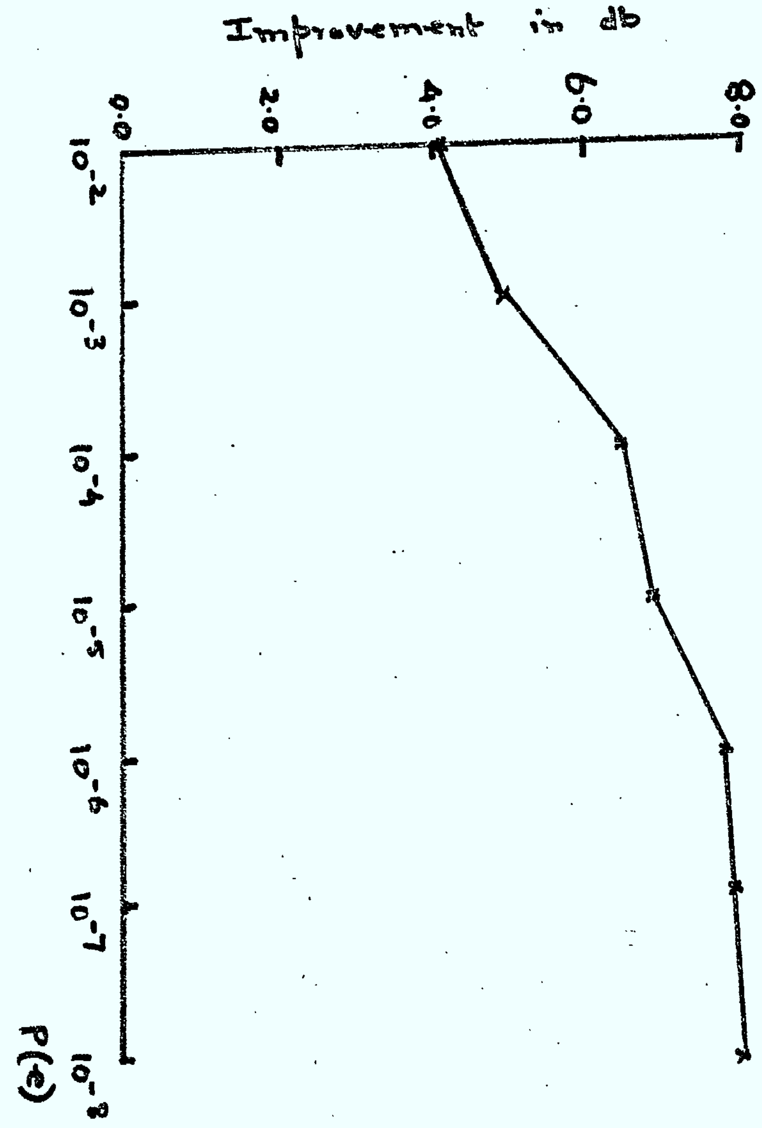

Fig. 4.4.

4.2 Summary

A regenerative satellite link was simulated and its performance compared to a non-regenerative satellite link. From the limited results obtained, the indication is that regenerative satellites show a substantial improvement in performance for the same E_b/N_0, or that they require smaller values of E_b/N_0 to achieve the same performance as conventional ones.

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CHAPTER 5

PARTIAL RESPONSE MODULATION METHOD CONSIDERED IN REGENERATIVE AND NON-REGENERATIVE SYSTEMS

5.1 INTRODUCTION

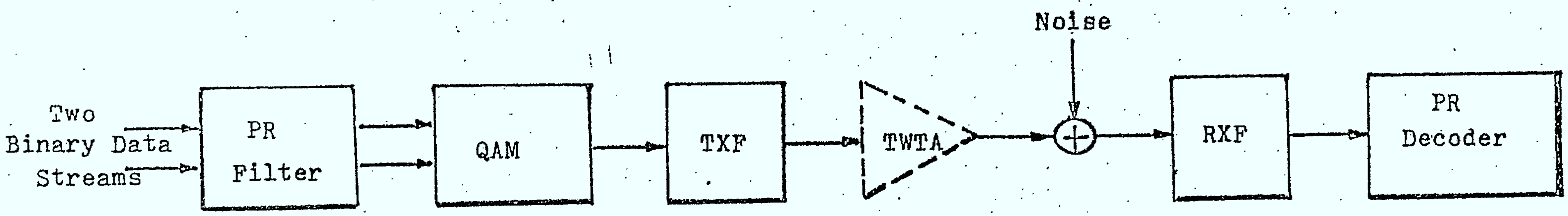
Partial-response signalling (PRS) is a bandwidth efficient baseband method of transmitting data and is hence suitable for carrier modulation. This efficiency is achieved using relatively simple, practically realisable and perturbation tolerant filters. The relatively simple hardware realisation of PRS, compared to other efficient methods, and its inherent capability of detecting errors without using redundant bits, are added advantages. In modulation with PRS, QAM enjoys more popularity and in this context it is referred to as Quadrature Partial-Response Signalling (QPRS). Examples include a Canadian product manufactured by Northern Telecommunications Limited, an 8 GHz long-haul high-capacity digital radio developed by Bell-Northern Research. This radio uses QPRS and has a spectral efficiency of 2.25 b/s/Hz [1]. In Japan, Fujitsu has developed a 2 GHz digital radio, also using QPRS, with a spectral efficiency of 2 b/s/Hz [2]. In the U.S.A., Avantek has been manufacturing 2 GHz QPRS digital radios which have a spectral efficiency of 1.8 b/s/Hz [3]. GTE Lenkurt has developed duobinary 3-level FM and 7-level correlative FM radio systems.

Although Partial Response is a spectrum-efficient modulation technique frequently employed in terrestrial microwave radio relay systems, no serious considerations have been given to its use in digital satellite applications. This is because its multi-level signalling nature could lead to severe AM/AM and AM/PM conversion degradations when a nonlinear power amplifier such as a TWTA is used. However, no readily available quantitative evaluation of these degradations is currently obtainable. Hence, it is of great interest to study the performance of a QPRS signal passing through a nonlinear device such as a TWTA. This study will not only establish a quantitative evaluation of the nonlinearity effects on QPRS but can also be used in the study of the feasibility of adopting QPRS modulation scheme in a regenerative satellite configuration.

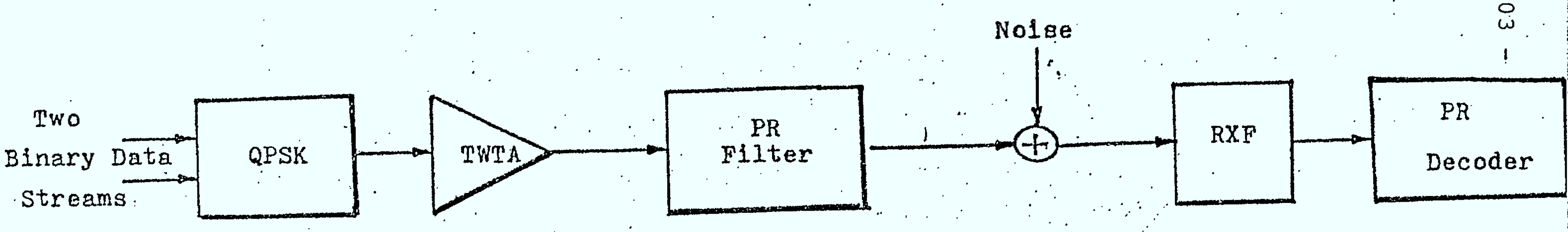
5.2 SIMPLIFIED QPRS MODELLING

A simplified model such as used in our computer simulation programme is shown in Figure 5.1. In this model, the channel nonlinearity arises from the use of a travelling wave tube amplifier (TWA) at the transmitter output. This device exhibits both AM/AM and AM/PM conversions as described in Chapter 2.

In the computer simulation [4], the two pseudo-random binary sequences of the quadrature channels are first precoded (using a modulo-2 summation) separately before being passed through a 3-level Class I encoder. Precoding is normally employed so as to avoid the propagation of errors due to adjacent bit dependency of the PR signal [5]. The resulting three level signals are then double-sideband amplitude modulated on two quadrature carriers and passed through a transmit bandpass (or its equivalent baseband) filter for spectral shaping.



a). Configuration of Computer Simulation for the Study of Filtering Effect.
If TWTA were Included, Severe Degradation Will Result.



b). Configuration of Computer Simulation for the Study of TWTA
Nonlinearity Effect

Fig. 5.1 Block Diagrams of a Simplified QPRS Radio System

In the present study, the characteristics of a typical TWTA are used. The power transfer characteristics and the output phase shift of the TWTA are as depicted in Figure 5.2. Starting from these characteristics, the nonlinear coefficients of the polynomial representation of the TWTA used in the simulation are obtained.

The effect of the position of the TWTA in relation to the bandpass filter is an interesting point to consider. First the case is considered where the bandpass filter preceeded the TWTA. In this situation it is found that the QPRS signal suffers severe degradations even when the TWTA was operated in the quasi-linear mode (8-12 dB input backoff).

Then the case is considered where the TWTA is placed before the partial-response filter. In this mode of operation, the signal at the TWTA input is essentially the undistorted QPSK signal which has a constant envelope. The degradation in this case is much less severe than the previous case. This is due to the nearby constant envelope of the unfiltered QPSK signal. For Class I QPRS, filtering should therefore be done after the nonlinear amplifier. The adoption of this strategy is mostly typified in the design of the 8 GHz digital radio by Bell-Northern Research [1].

In order to check the accuracy of the computer simulation programme, a brickwall Nyquist filter was first used both at the transmitter (for spectral shaping) and at the receiver

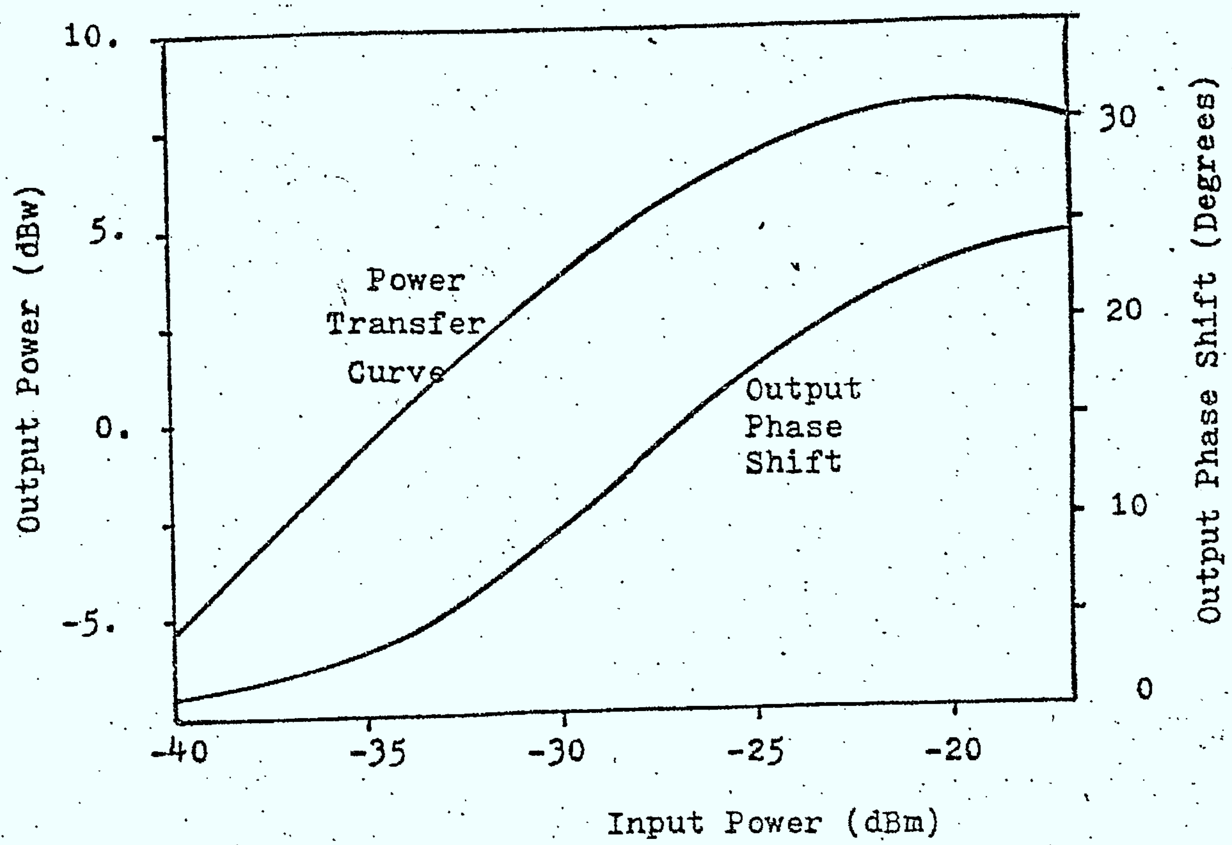


Fig. 5.2 The Single-Carrier Characteristics of Hughes 261-H TWTA

(for bandlimiting the noise and adjacent channel intereference). The result for this ideal case shows that a signal-to-noise ratio of SNR = 14.6 dB is required for a probability of error P(e) = 10^{-4}. This compares very favourably with the theoretical value of 14.4 dB for this model.

5.3 AM/AM AND AM/PM CONVERSION EFFECTS ON QPRS

It is of interest to study the separate effects of AM/AM and AM/PM conversion on the performance of the QPRS signal. Our results indicate that AM/PM conversion appears to have more drastic effects on QPRS than the AM/AM conversion by itself. This result is important because it points to a *possibility* of compensating the AM/PM conversion of a TWTA, if achievable, to *adopt QPRS for use in satellite regeneration systems.*

The TWTA was first assumed to exert only the AM/AM conversion. Based on [4], a 4-pole Chebyshev filter with a 0.01 dB passband ripple were used at both the transmitter and the receiver. This filter is specified in Figure 5.3. The results for this situation are plotted in the graphs of Figure 5.3.

Following this, both the AM/AM and AM/PM conversions of the TWTA were assumed in the simulation. Again a 4-pole Chebyshev filter with a 0.01 dB passband ripple were used at both the transmitter and the receiver. The results for this case are shown in Figure 5.4 for the TWTA operating at different input backoff points.

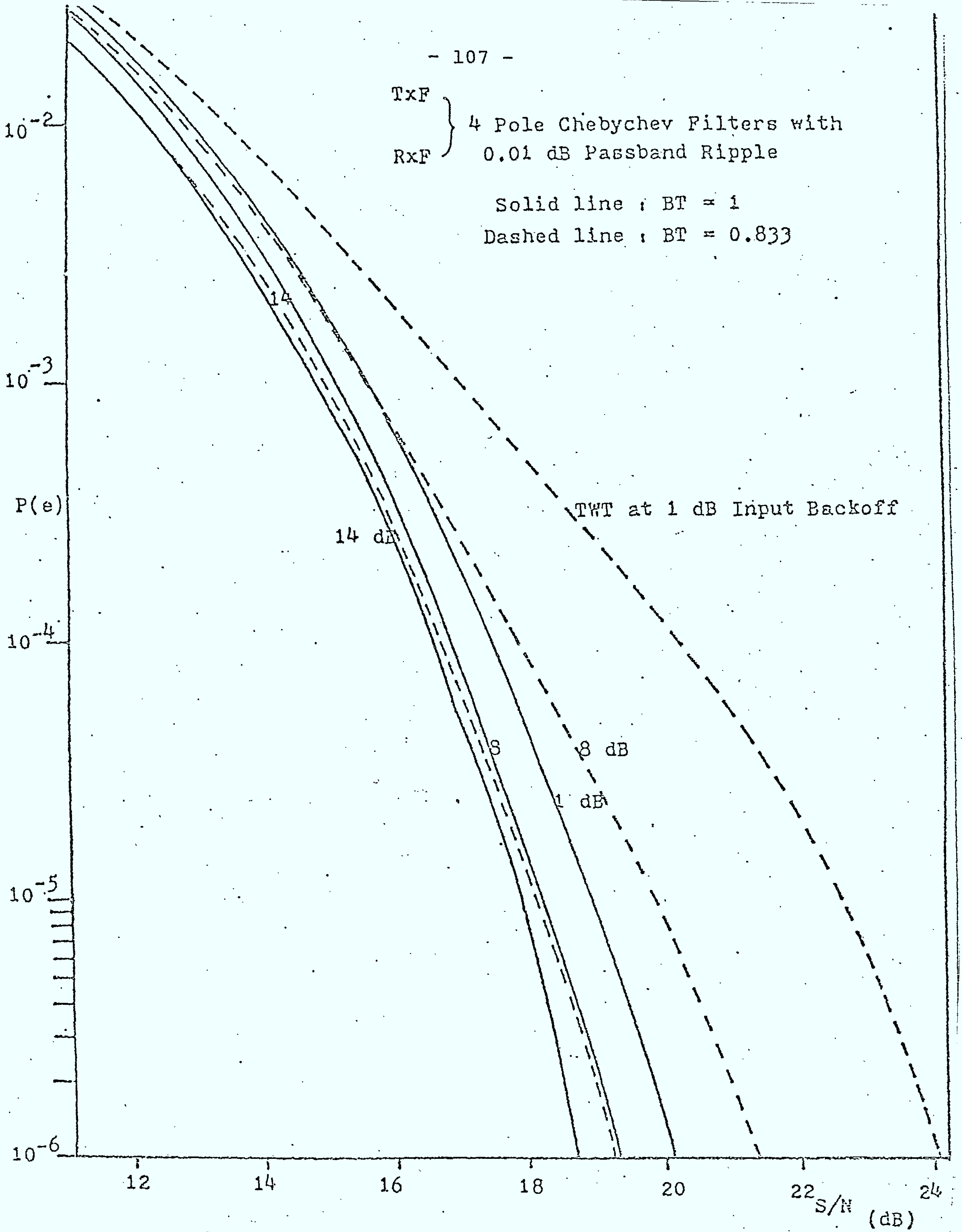


Fig. 5.3 Effect of TWT AM/AM Conversion on QPRS

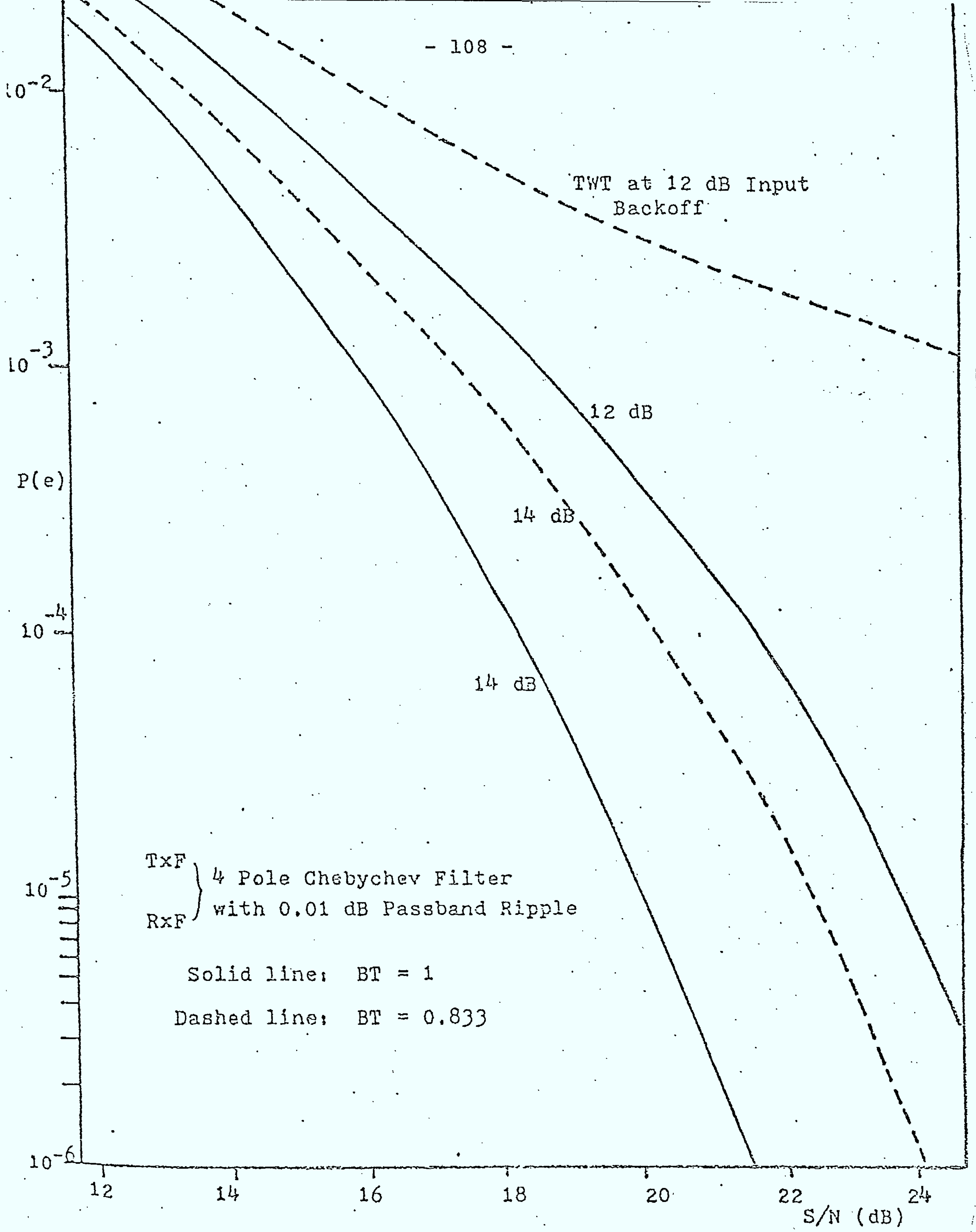


Fig.5.4 Effect of TWT AM/AM and AM/PM Conversion on QPRS

From the computer simulation results, it is clear that the major contribution to the overall degradation is the AM/PM conversion. On the other hand, the AM/AM conversion effect alone only slightly degrades the QPRS system performance. For example for a $P(e) = 10^{-4}$, in moving from an input backoff of 14 dB to an input backoff of 8 dB, the AM/AM degradation is less than 1 dB, whereas in moving from 14 dB backoff to 12 dB backoff, the overall (AM/AM plus AM/PM) degradation is close to 4 dB.

5.4 POSSIBLE REGENERATIVE SYSTEM CONFIGURATION

From the computer simulation results, it appears that if the AM/AM conversion of a TWTA can be somehow compensated for, it might be possible to *adopt QPRS on a satellite link* using regeneration on-board. DQPSK might be used on the uplink as it does away with the necessity of carrier recovery whereas QPRS might be used on the downlink. This configuration could be of interest because from the uplink, the on-board process can multiplex numerous signals from different earth stations into a single stream and then spectrally efficiently retransmit down, using QPRS, to a single receiver earth station. To reduce degradations, the QPRS shaping filter has to placed after the TWTA as mentioned earlier (see Figure 5.5). The TWTA can then be operated with less backoff.

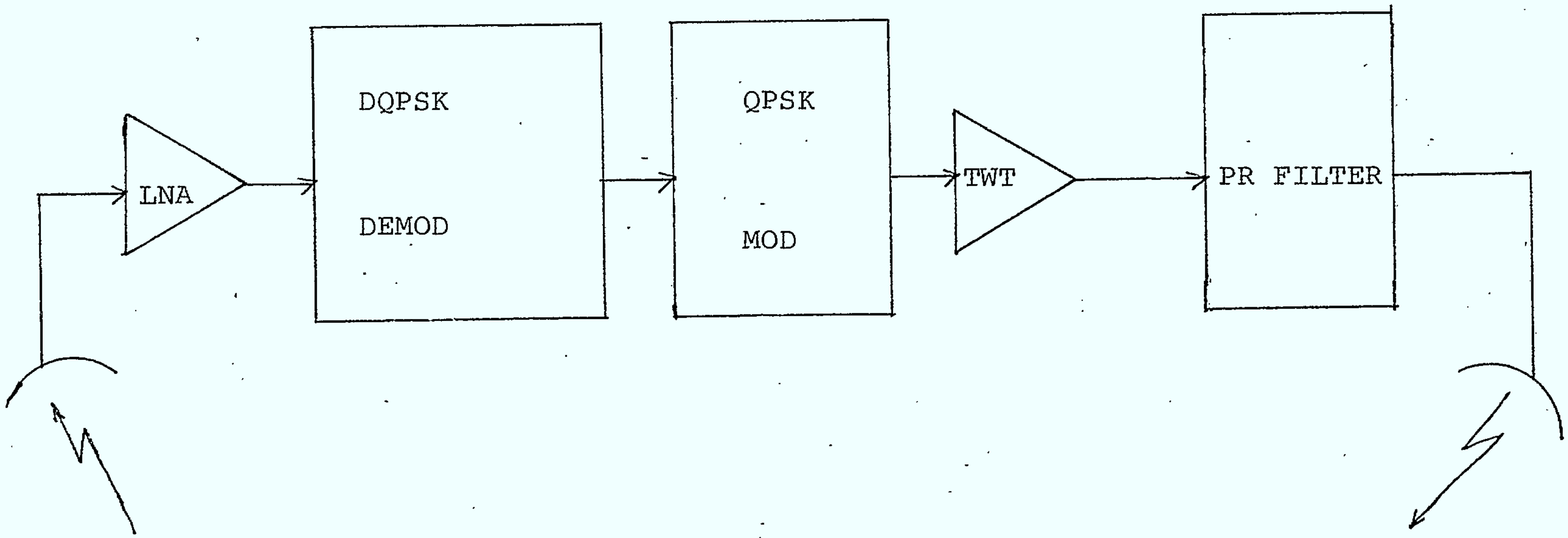


Fig. 5.5. Possible Configuration Using QPRS.

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CHAPTER 6

COMPARISON OF REGENERATIVE AND NONREGENERATIVE SATELLITE SYSTEMS

6.1 INTRODUCTION

On-board satellite regeneration is a good way to enhance the performance of a digital satellite link. It allows capabilities beyond those achievable with conventional translating satellites. These include considerable interference protection and interconnectivity between different types of terminals [1]. In the last decade regenerative satellite systems elicited considerable interest which resulted in some studies which are reported in the literature. In this chapter, based mainly on a litterature survey, the advantages of regenerative satellite systems over conventional non-regenerative satellite systems are explored. A performance comparison of regenerative and non-regenerative satellite systems is presented and the gain in signal-to-noise ratio for regenerative repeaters is given. The shortcomings of some of the results reported in the literature are pointed out.

6.2 ADVANTAGES OF REGENERATIVE SATELLITE REPEATERS

The models for the conventional and regenerative satellite repeaters are given in Figure 6.1. The conventional (translating) repeater essentially amplifies the received uplink signal, in addition to translating the carrier frequency, and redirects it towards the earth as a downlink signal. In other words, the incoming signal is retained in modulated carrier form. The regenerative repeater on the other hand, demodulates the incoming uplink signal into baseband, processes (or switches) the baseband stream and remodulates the RF carrier(s) before transmission towards the earth as downlink signal(s). In Figure 6.1, Nu(t) and Nd(t) are the uplink and downlink noises respectively and $I(t)$ is the uplink radio frequency interference, while $I'(t)$ is the radio frequency interference on the downlink. The amplifier used to amplify the signals in a conventional repeater is usually a TWTA and it is most often operated in a nonlinear mode to maximise the power output (little or no back off). This results in AM/AM and AM/PM degradations for non constant-envelope (bandlimited) signals.

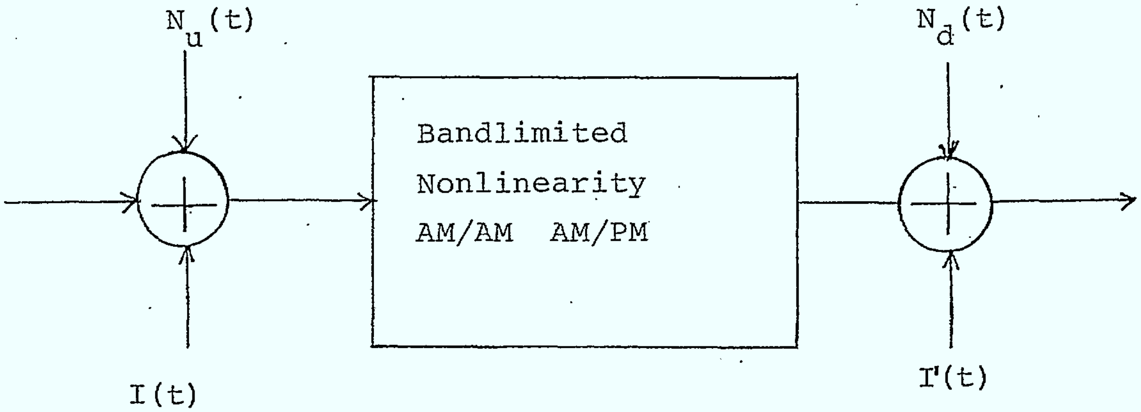
The advantages of regenerative repeaters are due to:

1) Isolation of the uplink from the downlink

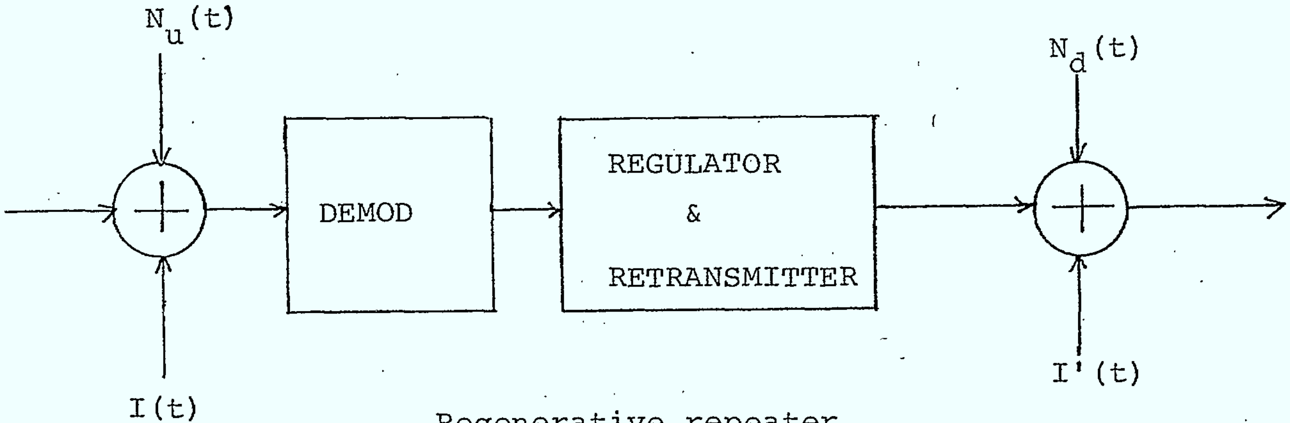
2) Availability of the detected baseband bit stream in the satellite

3) Reduction in size and weight

Regeneration prevents the accumulation of noise and co-channel interference. It passes only the bit errors made in the hard decision detection of the uplink signal at the satellite. As was seen in Chapter 2, this prevention of noise accumulation results in about 2.6 dB saving in the energy per bit-to-noise density ratio (Eb/No) in the linear channel, if the uplink and downlink Eb/No 's are identical.



Conventional (translating) repeater



Regenerative repeater

Figure 6.1 - Repeater Models for Conventional and Regenerative Satellites

The uplink and downlink are isolated due to the hard decision detection at the satellite and this leads to an efficient use of the available EIRP by allowing the optimisation of each link separately. This may lead to a cost-effective satellite system. For instance, each link may be separately equalised to reduce intersymbol interference. Also, different modulation methods may be used for the two links. In the uplink a method may be chosen that will allow simple demodulation in the transponder such as DPSK which requires no carrier recovery, whereas the downlink modulation may be chosen to give the greatest power/bandwidth efficiency. If the uplink bandwidth is limited, for example, 8-phase PSK may be used on the uplink, whereas QPSK may be used on the downlink (see Figure 6.2). It might be desirable sometimes to protect the uplink from radio frequency interference by using spread spectrum techniques on the uplink whereas thre is no such need in the downlink. This is also illustrated in Figure 6.2. If the downlink is power limited (as it usually is), error correcting codes may be used, on the downlink only, to trade power for bandwidth. Regeneration prevents signal anomalies which might occur on the uplink from affecting the downlink. For example it eliminates degradations due to the uplink filter-induced amplitude fluctuations, and the resulting phase modulation (PM) of the PSK carrier caused by the direct amplification of the uplink signal (with AM) by the nonlinear TWTA [2]. This can lead to improved demodulator performance and less link degradation [2].

Another advantage of regeneration is to prevent "power robbing" that occurs when there is a power imbalance among FDMA users - the weak signals are suppressed. With the demodulation/

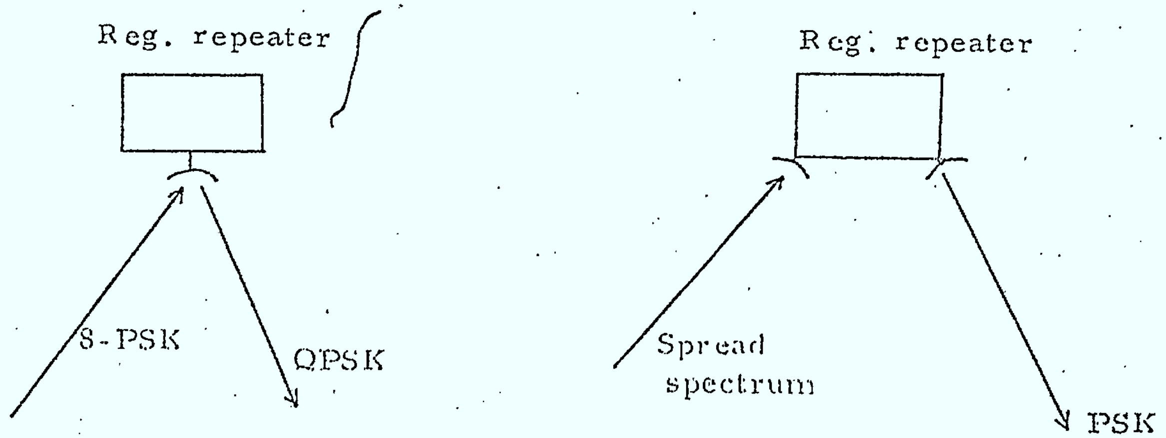


Fig 6.2 – Modulation conversion at the repeater.

remodulation scheme, sufficient downlink power is provided even for the low power uplink signals.

Regeneration provides versatility in the system. It is for example compatible with the store-and-forward concept for use with a single antenna beam. This allows bit-rate conversion to take place at the satellite. This might be called in for if there are different size antennas each using a different modulation method. Because of the insufficient gain of the small-size antennas, BPSK may be used for example while the large-size antenna would use QPSK. Figure 6.3 shows a possible configuration having one transponder for smaller size antennas and one for large-size antennas [3].

The availability, at the satellite repeater, of the detected baseband data stream leads to a high degree of interconnectivity thus additional advantages. It is possible for terminals of different types to communicate. Uplinks and downlinks with different levels or types of error correction coding may be interconnected. A careful combination of modulation technique and error correcting coding will result in better performances. For example, when a link is bandlimited for BPSK and power limited for QPSK, the combination of QPSK and error correction will be the best solution. Since errors on a satellite link usually occur isolated, the satellite is well suited for error correction techniques. In case of double errors (burst errors of length two) caused by differential decoding, simple bit interleaving will effectively convert the situation into a random (single) error one. Figure 6.4 show two types of forward error correcting coding configuration particularly useful when extra protection is required on the downlink. In diagram (a) only the downlink is protected by error correction and only the encoder, which is usually simpler than the decoder, is located at the satellite. In diagram (b) both uplink and downlink have error correction and again only an encoder is located at the satellite; the downlink is doubly coded. This is easier than if the uplink and downlink were encoded and decoded separately.

In case of voice transmission, since the baseband data stream is available, use may be made of Digital Speech Interpolation (DSI) to increase the capacity of the satellite link.

Beam hopping is also possible with regenerative satellites. A transponder is time shared by a multiplicity of spot beams. A phased-array beam antenna and some memory will be required. With a lot of users, the channel bit stream from each uplink may be connected to any downlink constituting a "switchboard in the sky". New conveniences to users, such as insert and drop, which are not now possible will become available. This baseband switching calls for electronic switching which, with recent advances in LSI circuitry, can be sufficiently miniature and very light. Switching is also possible with non-regenerative satellites, but in this case micro-wave switches using PIN diodes would be used. These are much larger and heavier than baseband switches.

In translating repeaters, degradation due to non linearities are very important in band limited situations(almost always) as mentioned earlier. The use of regeneration reduces the impact of these transponder nonlinearities. By replacing the TWTA with high-power direct modulators, signal degradation due to non-linearities will be avoided [3,4,5]. There will be no multiplicity

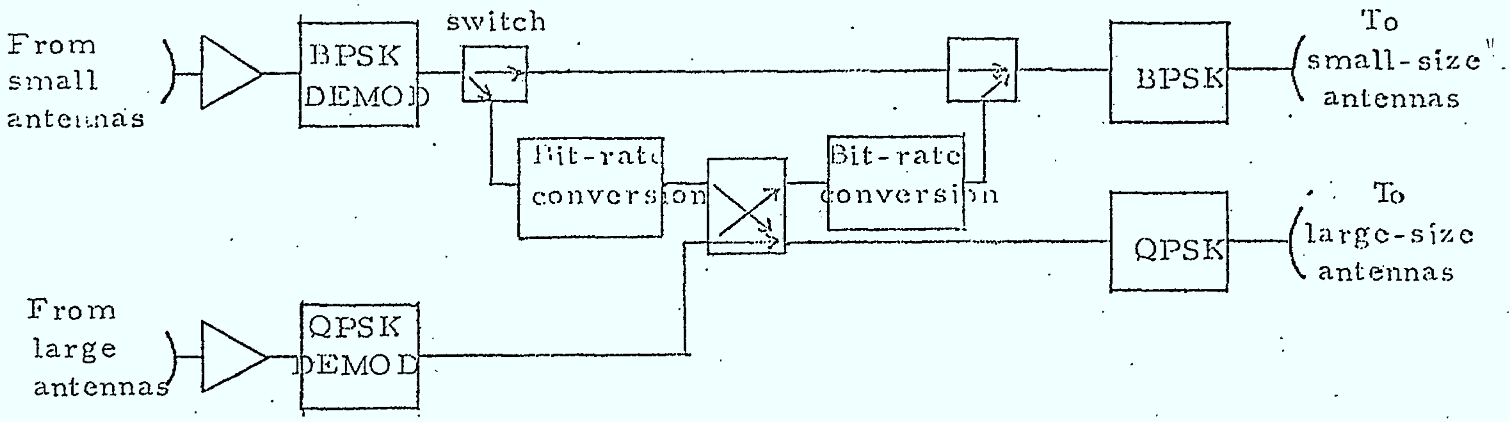


Fig 6.3 - Connexion of Transponders Using Different Modulations

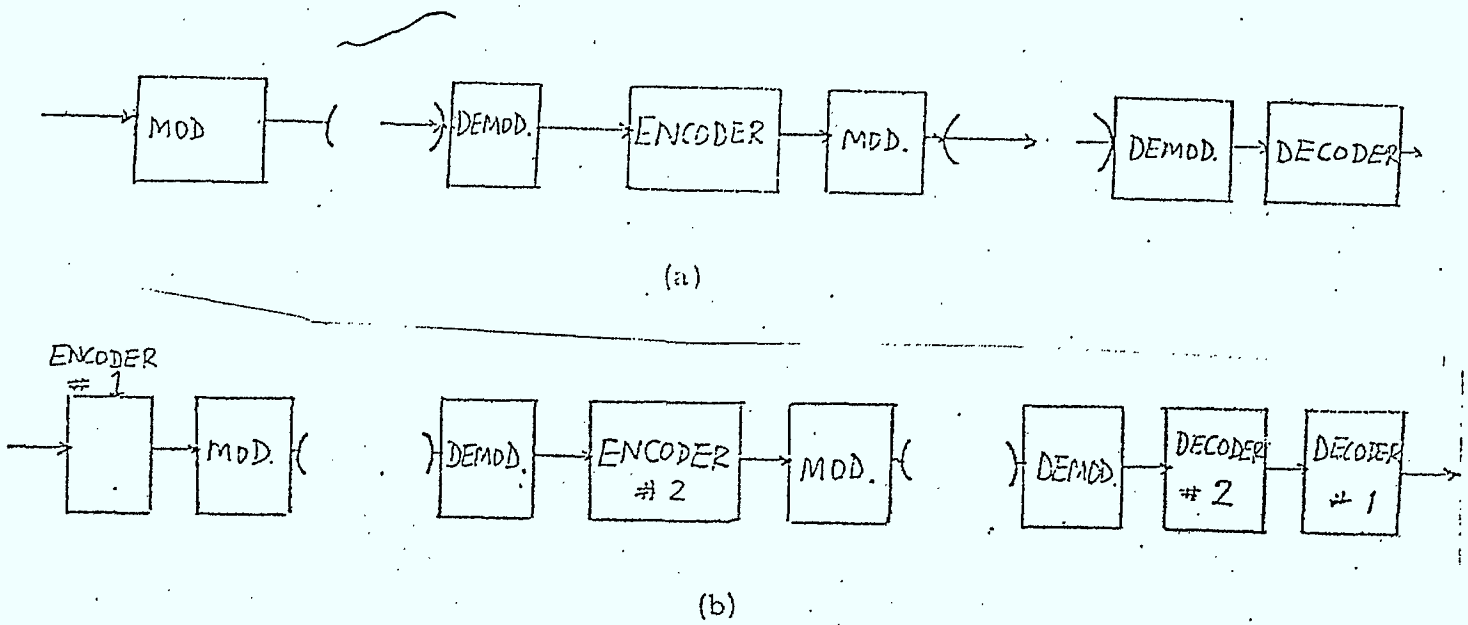


Fig 6.4 - Possible Satellite System Configuration with Error Correction

of (heavy) TWTA's. These high-power modulators would introduce a loss of not more than 1.5 to 2.0 dB.

If coherent detection is used, much of the channel filtering and equalisation can be done at baseband instead of at RF. This facilitates reduced complexity and weight. The remodulated carriers will be phase coherent for all accesses since they are derived from a common source. This eliminates station to station and Doppler differences encountered on the uplink and can lead to simpler ground terminal receiver design, improved demodulator performance, and less link degradation [2]. However, coherent detection calls for carrier recovery which can be complex. The choice between coherent systems or differentially coherent ones on the uplink continues to be a subject of interest [6].

A further advantage with regenerative repeaters is that concerned with a reduced intermodulation products problem among downlinks at different carriers. If the uplink consists of a number of FDM users, after demodulation, these can be multiplexed into one or a few TDM downlink bit streams, to be retransmitted on the downlink via remodulation. The reduced number of downlink carriers reduces somewhat the intermodulation problem. This calls for time-division multiplexers at the repeater and terminals equiped with time-division demultiplexers.

If a packet switching system is used, on-board processing can dramatically increase the throughput efficiency of the system [1].

The disadvantages of regenerative satellite systems include the power and weight that a satellite must carry for its own accomplishment (to carry out the processing), and the high degree of complexity (reduced reliability) that must accompany signal processing on board. However, these costs are wholly or partially recovered due to the more efficient use of satellite power, and the improved signal performance. The demodulated signal at the satellite may be subject to slight degradations due to phase jitter associated with on-board carrier recovery. Although the regenerative satellite was acclaimed flexible, a certain degree of inflexibility of a different kind is introduced. In conventional systems, any type of signal can be amplified at the satellite but with regeneration a transmitting station must use a modulation type the demodulator to which exists on board the satellite. Likewise a receiving station must match one of the downlink signal formats.

6.3 COMPARISON OF TRANSLATING AND REGENERATIVE REPEATERS

Several studies have been done comparing regenerative and nonregenerative satellite repeaters all of which indicate clear advantages of regeneration. Figure 6.5 (taken from Reference [3]) shows the effective total carrier-to-noise plus interference power ratio C/(N+I) as a function of the downlink C/(N+I) taking the uplink power as a parameter, for PSK systems. By effective C/(N+I) we mean that ratio that would attain the same bit error rate if only one link is used. It is seen that for a total C/(N+I) of 15 dB, the regenerative system can use an uplink C/(N+I) of 18 dB and a downlink value of 17.5 dB. The translating repeater system will need 24 dB uplink C/(N+I) and 20.5 dB downlink C/(N+I). This results in an uplink saving *of 6 dB* and downlink saving *of 3 dB*. This figure assumes miscellaneous losses of 0.5 dB on the uplink and 1.5 dB on the downlink for regenerative systems. The large loss on the downlink is due to possible degradation by the high-power direct PSK modulator in the repeater. For the nonregenerative system, total miscellaneous losses of 1.0 dB and TWTA nonlinear loss of 2.5 dB are assumed.

To illustrate further the advantage, the downlink C/N is plotted versus uplink C/N in Figure 6.6 for 4-PSK systems including conventional coherent QPSK, regenerative coherent QPSK and regenerative DQPSK up/CQPSK down (for the same error rate). Ideal Nyquist channels are assumed, and for the conventional transponder only a 1.6 dB degradation in C/N, due to TWTA saturation

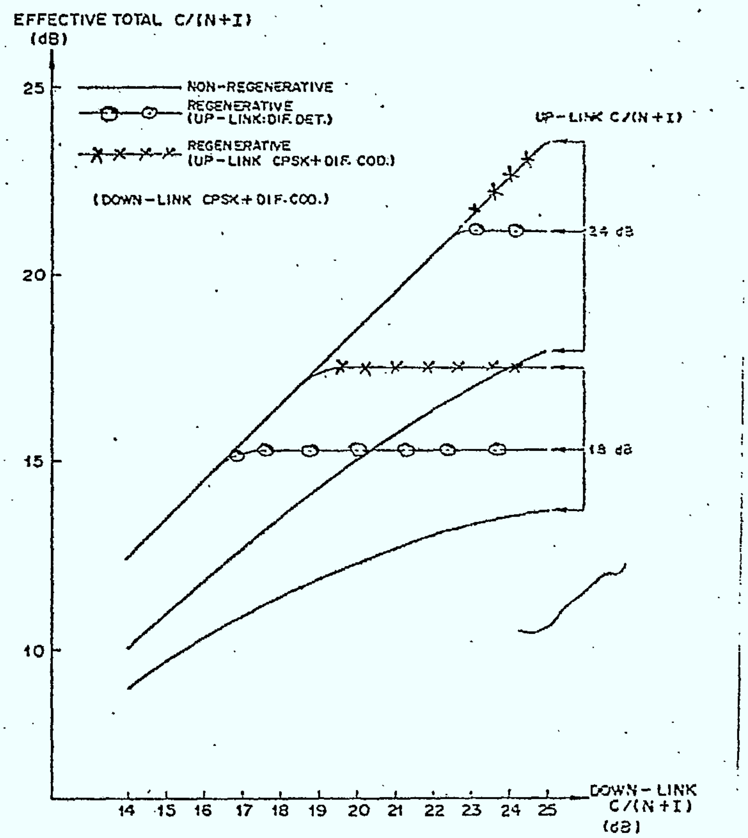


Fig.6.5 Comparison of regenerative and nonregenerative system.

is considered. It is to be noted that, if the downlink is to be maintained at 16 dB say, the nonregenerative repeater requires about 17 dB uplink C/N while the regenerative CQPSK system requires about 12 dB, a *saving of about 5 dB*. It is clear from the figure that the savings in C/N increase dramatically as the downlink C/N decreases. But regenerative DQPSK might actually result in a penalty for high downlink C/N values. A saving in C/N translates into either a *greater fade margin* or a *reduced power requirement* on both the uplink and downlink. It has been shown [2] that the improvement potential of regenerative satellite is greatest in cases where the degradation encountered for the conventional case is greatest, i.e. for the case with 0 dB backoff in both the ground station HPA and TWTA. Depending on the backoff, gains in C/N from 4.5 to 5 dB on the uplink and from 4.5 to 6.8 dB on the downlink can be achieved with regeneration using CQPSK.

Huang et al. [7] compared conventional (linear and hardlimited channels) and regenerative satellite systems for BPSK, QPSK and 8-PSK. Their results are reproduced in Figure 6.7. A given uplink signal-to-noise ratio, SNR, is assumed and then probability of error is plotted as a function of downlink SNR. The figures indicate that regeneration yields a higher gain as the number of phases of the modulation increase. Note also that for BPSK, when the uplink SNR is strong, the hardlimited channel performs almost as well as the regenerative satellite system. From Figure 6.7, it appears that a saturation effect

UPLINK
C/N (dB)

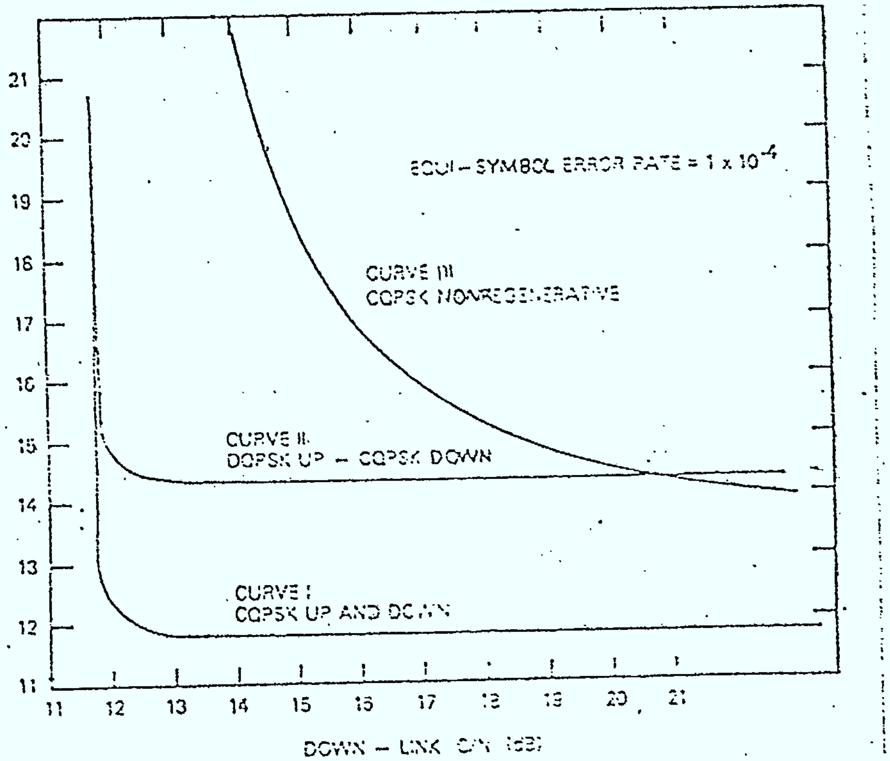
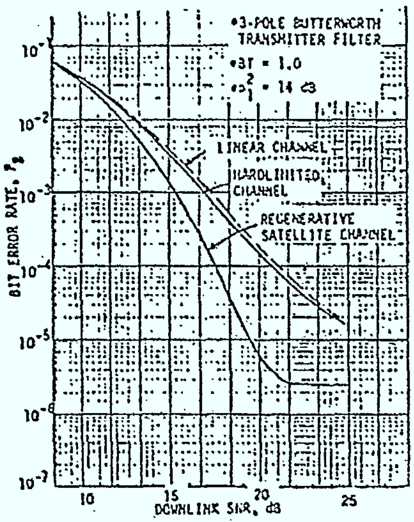
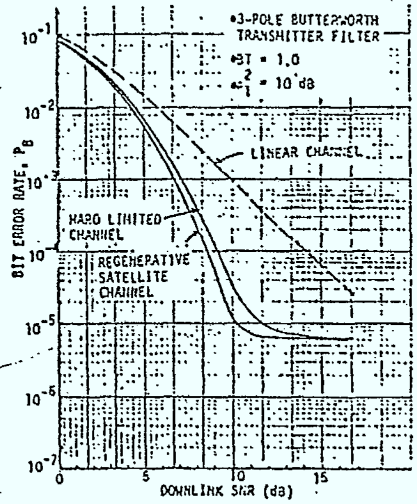


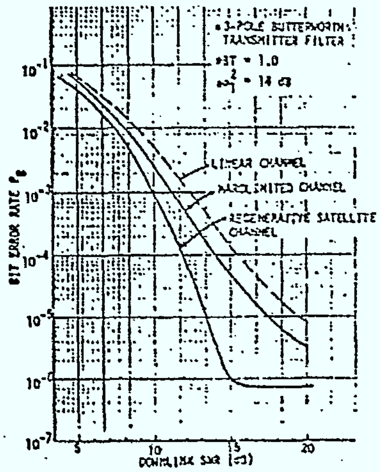
Fig 6.6 – 4- phase PSK systems using a regenerative repeater vs. a convensional transponder



Comparison of Bit Error Rate of 8PSK Signaling.



Comparison of Bit Error Rate of BPSK Signaling.

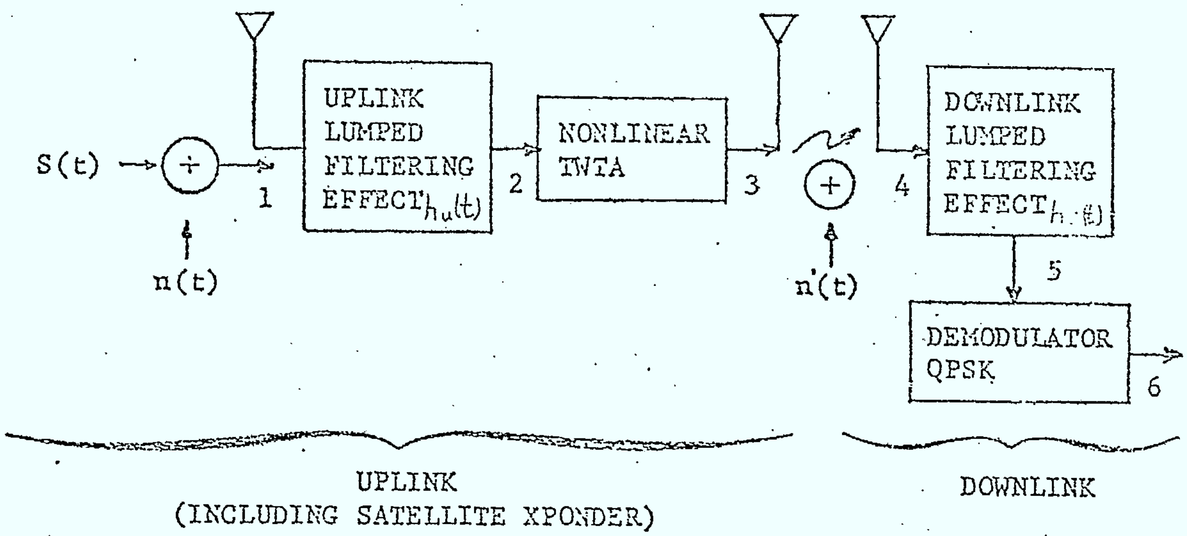


Comparison of Bit Error Rate of QPSK Signaling.

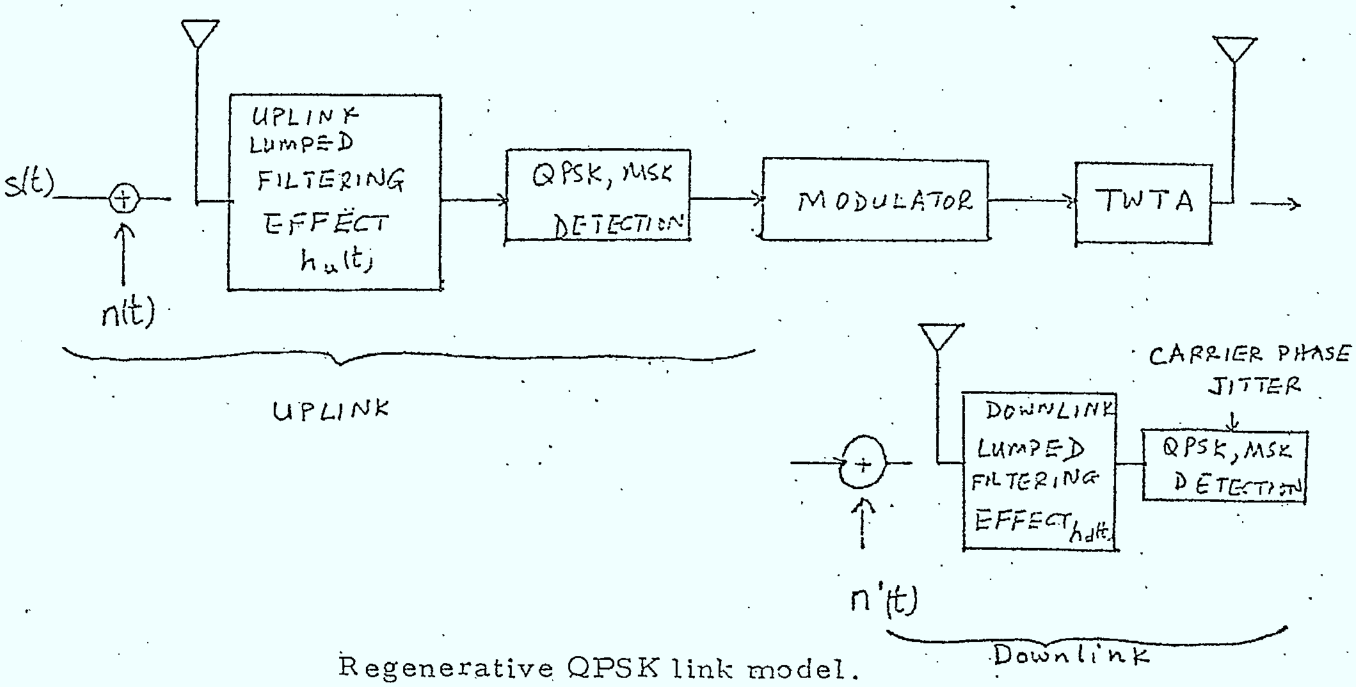
Fig. 6.7

occurs in the probability of error curves for a given uplink SNR. Beyond a certain value of downlink SNR, the probability of error does not change with increasing SNR. For BPSK, this value of downlink SNR seems to be approximately the same as the uplink SNR. Huang et al. compared performance between a mixture of FSK and PSK. It was found [7] that for a given uplink SNR and given that the uplink is FSK, the best performance occurs if the downlink is PSK. The results of Huang et al. are somewhat pessimistic as they deal only with the cases of a linear channel and a channel with a hardlimiter and ignores the practical case of AM/PM which is a major source of degradation in a satellite system.

Chiao and Chethik [5] carried out extensive calculations of performance bounds on the probability of error for both QPSK and MSK using translating repeaters and also regenerative repeaters. The effect of AM/PM and also the effect of reference-carrier reconstruction errors due to noise and intersymbol interference was considered. Their analysis is based on the model shown in Figure 6.8. The effect of intersymbol interference introduced by the filtering effects along the uplink path and along the downlink path are represented by a lumped uplink filter with an impulse response $h_u(t)$, and a lumped downlink filter with an impulse response $h_d(t)$, respectively. The uplink noise, $n(t)$, and downlink noise, $n'(t)$, are assumed to be both white and gaussian.



Translating QPSK Link Model



Regenerative QPSK link model.

Figure 6.8- Models of Translating and Regenerative QPSK Links

Their results indicate that for the conventional non-regenerative system, performance is better when intersymbol interference exists in the uplink only and the downlink is ISI-free. For the regenerative system, performance is better when ISI is on downlink only and the uplink is ISI-free. Just as found by Huang, et al., a saturation effect occurs if either downlink or uplink is held constant. This indicates that for a given value of downlink E_b/N_0 say, there exists an optimum uplink E_b/N_0. A comparison of conventional translating and regenerative performance for QPSK is shown in Figure 6.9 for a situation where no intersymbol interference exists. To check briefly the validity of these curves we note that if both uplink and downlink E_b/N_0 are 13.6 dB, the probability of error for the translating system is 10^{-4}. But this is equivalent to an overall E_b/N_0 of 10.6 dB. Theoretically QPSK is supposed to have an E_b/N_0 of 8.4 dB for $P(e) = 10^{-4}$ [8]. It is therefore seen that these results may be about 2 dB away from theoretical expectation.

From Figure 6.9 it can be seen that for low uplink E_b/N_0 (in the range of 10 dB) the conventional repeater has P(e) performance worse than 10^{-3}. With the same uplink E_b/N_0, regeneration restores the P(e) to the 10^{-4} to 10^{-6} range even with a downlink E_b/N_0 of 10 dB. For higher uplink E_b/N_0 say 15 dB, it

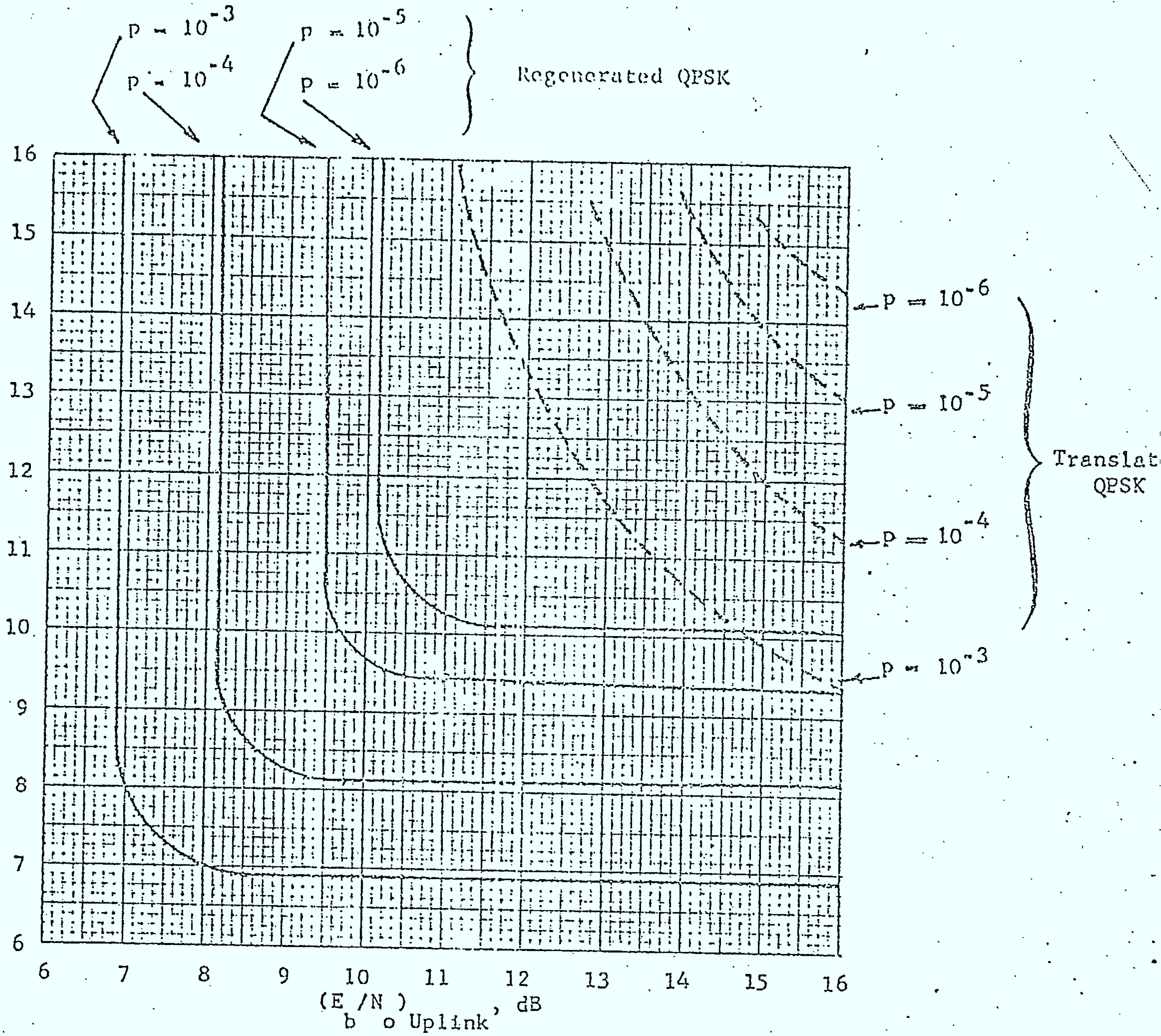


Figure 6.9- BER Comparison Between Regenerative and Convention Repeaters for QPSK

is the downlink E_b/N_0 which limits the performance. For uplink and downlink E_b/N_0 which are not severely P(e) limiting, the regenerative repeater offers 2-5 dB improvement in performance. These results, although they include AM/PM conversions, are rather conservative in as much as they do not consider optimisation of the TWTA backoff. Also they do not include a HPA in their model which appears in actual systems. A more thorough study would have to include a cascaded nonlinearities incorporating a ground station HPA and a satellite TWTA. The optimisation of the backoffs of both HPA and TWTA would be required.

6.4 SUMMARY

Regenerative satellite repeaters were seen to have advantages over conventional translating repeaters especially if the uplink power is limited. Regeneration leads to more efficient satellite EIRP utilisation and may offer maximum interference protection. Depending on the *backoff conditions* regeneration may achieve gains in uplink *E_b/N_0 of up to 5 dB and in the downlink E_b/N_0 of up 6.8 dB*. This means *greater fade margin or a reduced power requirement* in the satellite system and may therefore lead to a *cost-effective* satellite system.

6.5 REFERENCES ON CHAPTER 6

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CHAPTER 7

CONCLUSIONS AND FURTHER CONSIDERATIONS

7.0 RECOMMENDATIONS, FURTHER RESEARCH AND CONSIDERATIONS

Our studies on regenerative satellite systems lead up to conclude that the Department of Communications should seriously study this most interesting and important topic for future digital satellites. On-board satellite regeneration seems to be a promising method of achieving more cost-competitive satellite systems. The reduction in the uplink power requirement to meet a given bit error rate performance criterion means that smaller earth station antennas may be used resulting in a reduced overall cost of the satellite system link. Current satellite systems are bedevilled by degradations which result when a bandlimited signal is amplified by nonlinear devices. On-board regeneration reduces these degradations and hence might be included in future satellites.

The above recommendation will stimulate Canadian Industry to carry out research in this area and hence fulfill Canada's commitment to leadership in Telecommunications.

In Phase II of the expected research contract, a more indepth study of QPSK will be carried out to verify the results obtained so far. The frequently considered modulation methods for satellite communications shall be studied for application to regenerative satellite systems. These will include offset-keyed QPSK, MSK, and DQPSK. Our initial study indicates the possibility of employing partial-response signalling in regenerative satellite system configuration. The feasibility of this technique shall be further explored. Other new modulation methods which might be applicable to regenerative satellites shall be looked into. These could include offset-keyed-8 PSK (OK-8 PSK) and nonlinearly switched filter techniques.

As mentioned earlier in this report, a more complete model of a satellite system has to include adjacent- and co-channel interferences; this shall also be considered. The effects of synchronisation errors should be taken into account. An interesting concept to consider also would be direct RF regeneration [1].

A more thorough consideration of the cost-effectiveness of regenerative satellite systems would result from a study of practical antenna sizes achievable by the use on-board regeneration in a satellite link.

7.1 REFERENCE

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