PERFORMANCE EVALUATION OF NONLINEAR SATELLITE LINK BY
COMPUTER SIMULATION
by
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Thesis submitted to the Faculty of the
Virginia Polytechnic Institute and State University
in partial fulfillment of the requirements for the degree of
Master of Science
in
Electrical Engineering

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December, 1985
Blacksburg, Virginia
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(ABSTRACT)

The performance of a nonlinear satellite channel using QPSK (Quaternary Phase Shift Keying) and MSK(Minimum Shift Keying) has been studied by computer simulation. In the simulation, the pseudo randomly generated input data stream modulates the carrier and this modulated carrier passes through the typical satellite communication link, that consists of a transmit earth station, a satellite transponder, and a receiving earth station. All the signals used in the simulation procedure are real-valued and are transformed back and forth between time and frequency domains depending on the necessity by the Fast Fourier Transform (FFT) or Inverse Discrete Fourier Transform (IDFT). The simulation result represented as average Bit Error Rate (BER) gives the basis for comparison of the performance in various link conditions such as linearity, nonlinearity, band-limited and noisy channels which are expected to be encountered in practical situations.
ACKNOWLEDGEMENTS

I would like to thank Dr. Tri T. Ha for his kind suggestions and guidance. I am indebted to him for introducing me to the theoretical concepts needed for this thesis. I am also grateful to Dr. T. Pratt for his interest and encouragement.

Special thanks are due to my wife and my parents for their tremendous supports.
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The study of a communication system is inherently concerned with the performance of a signal through the system. A system is a combination of devices and networks chosen to perform the desired function. Because of the sophistication of modern digital satellite communication systems, computer simulation of a satellite link is regarded as an economical alternative to hardware simulation and analytical modeling, especially when there exist nonlinear elements like High Power Amplifiers (HPA) and Travelling Wave Tube Amplifiers (TWTA) in the link, which cause difficult problems which cannot be solved by purely mathematical methods. In addition to the primary advantage of computer simulation, a variety of configuration can be analyzed once the simulation software has been developed. This thesis concentrates on the computer simulation of a satellite link that includes the nonlinear TWTA using two types of digital modulation technique: QPSK and MSK.

A typical single satellite communication link is depicted in Fig.1.1. At the transmit earth station, the input data stream, that consists of NRZ (non-return to zero) rectangular pulses, is generated by the pseudo-random sequence generator. This stream modulates a sinusoidal carrier. The modulated
carrier passes through the modem transmit(Tx) filter which shapes the waveform to reduce the Intersymbol Interference(ISI). The bandlimited waveform is then amplified by the earth station HPA. At the satellite transponder, the input multiplexer filter rejects adjacent channel interference to the specific channel. The TWTA amplifies the received signal, which is greatly attenuated by the space and retransmits the signal to the receive earth station. The output multiplexer filter reduces the signal energy falling into adjacent channels. At the receive earth station, the modem receive(Rx) filter provides additional waveform shaping to reduce ISI and also to reduce the input noise. In the link, thermal noise enters both at the satellite input stage and at the earth station's Rx filter.

In simulation, the following assumptions are used: 1) the bandwidth-symbol duration product, BT, is moderately large (1.2—1.6); 2) The channel is downlink limited, that is, thermal noise entering the overall path is represented by downlink Additive White Gaussian Noise(AWGN) with density of No/2 (W/Hz) at the modem Rx filter; 3) earth station HPA is assumed to operate in the linear region. Therefore, the nonlinear element dealt with in this thesis is restricted to the TWTA in the satellite transponder. Chapter 2 provides the review of digital modulation schemes namely, QPSK and MSK, necessary to understand the main stream of the simu-
lation procedure. QPSK is the most popular digital modulation in satellite communications, whereas MSK is a promising digital modulation for Ka band (20—44 GHz) satellite systems with wider transponder bandwidth because of its spectral efficiency. Chapter 3 presents an investigation on the major problem to be dealt in the thesis, namely, the nonlinearity of the transponder TWTA and ISI caused by the nonlinearity. Since the TWTA has both amplitude and phase nonlinear characteristics, performance degradation exists in the transmitted signal. This fact inevitably makes the modem Tx/Rx filters play an important role in minimizing signal degradation. Chapter 4 is devoted to the detailed description of modeling procedures, which include the pseudo random number generator, QPSK/MSK modems, Tx and Rx filters, compensator, and sampler. Then, the satellite transponder incorporating the TWTA, input/output multiplex filters are described. In addition, the method of representing the link performance through average BER is also explained. Chapter 5 presents the description of the various tests conducted to get the simulated data at first, then the results are plotted for comparison and for discussion.
Figure 1.1: A typical single satellite communication link model.
CHAPTER 2 REVIEW ON DIGITAL MODULATION

2.1 INTRODUCTION

Of many digital modulation techniques, only a handful of them are desirable for satellite communications, because a satellite nonlinear TWTA requires a constant envelope, bandwidth efficient modulation technique. QPSK and MSK are the most feasible modulation techniques for the above requirement because they have constant envelopes and they utilize a relatively small amount of bandwidth. This chapter presents a review of QPSK, which is a good choice for nonlinear and severely band-limited channels and MSK which is a good choice for the channels where the up-link adjacent channel interference is a primary concern.

2.2 QPSK (QUATERNARY PHASE SHIFT KEYING)

QPSK is a digital phase modulation technique where the information of the digital signal is transmitted in the phase of the modulated carrier. The phase of the carrier can take on one of 4 possible values during each signalling interval. The modulated QPSK signal is represented by

\[ S_{\text{QPSK}}(t) = A \cos(\omega_c t + \phi_k) \quad 0 < t < T_s \]  

(2.1)
where
\[ A : \text{carrier amplitude} \]
\[ \omega_c : \text{carrier radian frequency} \]
\[ \phi_k = k\pi/4 \text{ for } k = 1, 3, 5, 7 \]
\[ T_s : \text{symbol duration}. \]

If we expand Eq(2.1) using trigonometric identities, we can get the other form of QPSK signal represented by

\[ S_{QPSK}(t) = (A\cos\phi_k) \cos\omega_c t - (A\sin\phi_k) \sin\omega_c t \]
\[ = \pm \frac{A}{\sqrt{2}} \cos\omega_c t \pm \frac{A}{\sqrt{2}} \sin\omega_c t \quad 0 < t < T_s. \]

In Eq(2.2), the first and the second terms are all BPSK signals. The first term is a BPSK signal in phase with the carrier, whereas the second term is a BPSK signal in quadrature with the carrier. Therefore, if we combine two BPSK signals in phase quadrature, QPSK signal can be generated. For notational convenience, we call the first term the I (in-phase) channel and the second the Q (quadrature phase) channel. When the binary symbol 1 (or +1 in NRZ form) represents \( +1/\sqrt{2} \) volt, and the binary symbol 0 (or -1 in NRZ form) represents \( -1/\sqrt{2} \) volt, we can say that the unique dibits (pairs of bits) \((1,0), (0,0), (0,1) \) and \((1,1) \) generate the four possible QPSK phases such as \( \pi/4, 3\pi/4, 5\pi/4, \) and \( 7\pi/4 \) respectively. Fig.2.1 shows the phasor diagram of QPSK signal associated with their unique dibits.
Figure 2.1: QPSK signal phasor diagram with unique dibit.
Figure 2.2: (a) QPSK modulator (b) QPSK demodulator.
QPSK MODULATOR

Fig. 2.2.a shows the block diagram of QPSK modulator, which consists of demultiplexer (also called serial to parallel converter), a pair of carrier multipliers or mixers, local oscillator, 90 degree phase shifter and summer. When the incoming NRZ data stream with bit rate $R_b$ bits per second (bps) is applied to the modulator, the demultiplexer divides the incoming data stream into odd bits stream and even bits stream. Usually the odd bits stream is represented as $S_I(t)$ and the even bits stream is represented as $S_Q(t)$. To achieve the bandwidth efficiency of 2 bps/Hz, the signaling interval (also called symbol duration $T_s$) is made twice as long as the bit duration $T_b$ (where $T_b = 1/R_b$) of the incoming data stream. Then the odd bit stream $S_I(t)$ modulates the in phase carrier generating the I-channel BPSK signal and the even bit stream $S_Q$ modulates the 90 degree phase shifted carrier, thus generate the Q-channel BPSK signal. Finally these two I-channel and Q-channel signals are summed to form the QPSK signal. Because two input data streams $S_I(t)$ and $S_Q(t)$ are synchronously aligned such that their transitions coincide, the phase shift in QPSK can occur once after one symbol duration, $T_s$. As we can see from the QPSK phasor diagram depicted in Fig. 2.1, the phase shift occurs only when one or both bit streams changes bits in one symbol duration. Therefore, if both channel data streams change their data,
±180 degree phase shift occurs, whereas ±90 degree phase shift occurs when only one channel data stream changes its data. When there are no changes in both channel data streams, no phase shift occurs. The sign of phase change depends on which channel has changed its data in one symbol duration compared with the previous state.

**QPSK DEMODULATOR**

In general, QPSK signal can be demodulated using a coherent demodulation technique if a phase reference is available. Because the QPSK signal is the sum of two BPSK signals in phase quadrature, the typical QPSK demodulator contains two BPSK demodulators. Fig.2.2.b shows the QPSK demodulator, which consists of two binary correlators connected in parallel, a pair of samplers, and a multiplexer. The correlator itself consists of a carrier multiplier or mixer and an integrate-and-dump (I & D) circuit. The symbol timing recovery and carrier recovery parts are excluded from the diagram because we have assumed the perfect symbol timing and carrier recovery in simulation for simplicity. The in-phase correlator extracts the cosine component of the carrier phase, while the quadrature correlator extracts the sine component of the carrier phase. Under noise-free condition, the input signal $m(t)$ to the QPSK demodulator is represented by

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\[ m(t) = A \cos (\omega_c t + \phi_k). \] (2.3)

After carrier multiplication, the I-channel component \( m_{I1}(t) \) can be expressed as

\[
m_{I1}(t) = A \cos(\omega_c t + \phi_k) \cos \omega_c t = \frac{A}{2} [\cos \phi_k + \cos(2\omega_c t + \phi_k)]
\] (2.4)

and the Q-channel component \( m_{Q1}(t) \) is written as

\[
m_{Q1}(t) = A \cos(\omega_c t + \phi_k) \sin \omega_c t = \frac{A}{2} [-\sin \phi_k + \sin(2\omega_c t + \phi_k)] \] (2.5)

These two signals \( m_{I1}(t) \) and \( m_{Q1}(t) \) are passed through the I&D filters which eliminate the second harmonics, and then are sampled at the symbol rate \( R_S = 1/T_S \). Immediately after sampling, the integrators are instantaneously discharged. The signs of the time sampled integrator outputs uniquely decide which one of the phase angle was transmitted. The integrators are usually replaced by the low-pass filters (LPF) in the high-speed modems (modulator and demodulator) where the input bit rate is more than 10M bps. The filtered I-channel signal and the filtered Q-channel signal are the baseband signals. The two baseband signals are then passed through threshold detectors. For each channel, if the signal is positive, a binary symbol 1 is assumed to be transmitted, whereas a binary symbol 0 is assumed to be transmitted if the detected signal is negative. Finally these two binary se-

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quences from the I and the Q channels are combined in a multiplexer to restore the originally transmitted binary input data sequence.

2.3 MSK (MINIMUM SHIFT KEYING)

The desire to achieve the maximum spectral efficiency in digital modulation for wideband satellite channel inevitably leads to the modulation scheme which maximizes the bandwidth efficiency and minimizes the spectral components which fall outside the Nyquist bandwidth. Compared to QPSK which has a bandwidth efficiency of 2 bps/Hz and the falloff rate of its spectra is proportional to the inverse square of frequency($f^{-2}$), CPFSK (Continuous Phase Frequency Shift Keying) with proper control of its phase can achieve the same bandwidth efficiency as QPSK and its power spectra also decrease at a rate proportional to at least four times ($f^{-4}$) away from the carrier frequency. [pasupathy.1]

The typical form of CPFSK signal is

$$S(t) = A \cos \left[ \omega_c t + \phi(t) \right],$$  \hspace{1cm} (2.6)

where $\phi(t)$ is no longer constant as in QPSK. Let the radian frequencies $\omega_1$ and $\omega_2$ represent the binary symbol 0 and 1 respectively, and the carrier radian frequency $\omega_c$ is chosen to be $(\omega_1 + \omega_2)/2$. In addition, if the radian frequency devi-
ation \( \Delta \omega \) is chosen to be \( (\omega_2 - \omega_1)/2 \), then the phase function \( \phi(t) \) with respect to carrier frequency in one bit interval can be expressed as

\[
\phi(t) = \pm \Delta \omega t + \phi(0) \quad 0 < t < T_b,
\]

\( \phi(0) \) is the initial phase needed to maintain the phase continuity and depends on the past history of modulation process. To achieve orthogonal frequency shift keying, the frequency separation in the bit interval \( (0, T_b) \) must be chosen to satisfy the condition \( 2\Delta \omega T_b = n\pi \). Therefore the minimum frequency separation requires \( n = 1 \), hence

\[
\Delta \omega = \frac{\pi}{2T_b}.
\]

This particular choice of CPFSK is called MSK (Minimum Shift Keying). Thus MSK is CPFSK with a frequency separation equal to the one half of the bit rate. Therefore the expression of the phase function in Eq(2.7) can be written as

\[
\phi(t) = \pm \frac{\pi}{2T_b} t + \phi(0) \quad 0 < t < T_b.
\]

The phase function \( \phi(t) \) of Eq(2.9) is confined to one bit interval \( T_b \). To see the phase variation with time \( t \) over one \( T_b \), let's assume \( \phi(0) = 0 \). Then it can easily be seen that \( \phi(t) \) takes only two values of \( \pm \pi/2 \) at odd multiples of \( T_b \) and only
two values 0, π at even multiples of $T_b$. Therefore the phase variation over each $T_b$ is ±π/2. By defining $b_k$ as a binary switching function taking the value of ±1 only for $k = 1, 2, \ldots$ and $\phi_k( = \phi[(k-1)T_b])$ as excess phase at the beginning of $k$-th bit interval to maintain continuous waveform, we can write the phase function $\phi(t)$ over more than one $T_b$ as

$$\phi(t) = b_k \cdot \frac{\pi t}{2T_b} + \phi_k. \quad (2.10)$$

In addition, from the fact that $\phi(t)$ varies only ±π/2 in one $T_b$, the recursive phase constraint can be obtained as

$$\phi_k = \phi_{k-1} + b_k \frac{\pi}{2}. \quad (2.11)$$

Therefore, substituting Eq(2.10) into Eq(2.6) yields the MSK signal in CPFSK form over more than one bit interval

$$S_{\text{MSK}}(t) = A \cos \left[ \omega_c t + b_k \frac{\pi}{2T_b} t + \phi_k \right]. \quad (2.12)$$

If we expand Eq(2.12) using trigonometric identities, it can be expressed in terms of in-phase and quadrature components as

$$S_{\text{MSK}}(t) = A[\cos(\frac{\pi t}{2T_b} b_k + \phi_k)\cos\omega_c t - \sin(\frac{\pi t}{2T_b} b_k + \phi_k)\sin\omega_c t]. \quad (2.13)$$
If we assign the even numbered data pair of the input binary data stream to the in-phase components and the odd numbered data pair to the quadrature components, the excess phase $\phi_k$ will always increase by 0 or $\pi$ (modulo $2\pi$) for the successive value of either of the components values, and the phase difference between components will be $\pm\pi/2$ by the phase constraint in Eq (2.11). This phase difference can be taken care of by giving a relative time delay of $T_b$ between the in-phase and quadrature components as is used in Staggered QPSK. [Stremler.7] If Eq (2.13) is expanded further using trigonometric identities and the fact that $\sin\phi_k = 0$ for both the in-phase and the quadrature components, we get

$$S_{MSK}(t) = A(\cos\phi_k \cos\frac{\pi t}{2T_b} \cos\omega_c t - b_k \cos\phi_k \sin\frac{\pi t}{2T_b} \sin\omega_c t).$$

(2.14)

The terms $\cos\phi_k$ and $b_k \cos\phi_k$ in Eq (2.14) can be considered as two quadrature data channels $S_I(t)$ and $S_Q(t)$ of Staggered QPSK. Then we can finally express the MSK signal derived from Staggered QPSK as

$$S_{MSK}(t) = A (S_I(t) \cos\frac{\pi t}{2T_b} \cos\omega_c t - S_Q(t) \sin\frac{\pi t}{2T_b} \sin\omega_c t).$$

(2.15)

Therefore we can conclude that MSK is either a special case of Staggered QPSK with cosine pulse shaping or a special case...
of CPFSK with the frequency separation of half of the incoming bit rate.

**MSK MODULATOR**

The fact that MSK can be regarded either as a special case of Staggered QPSK with a half sinusoidal pulse shape rather than rectangular pulse shape or as a particular case of CPFSK with a frequency spacing equal to the one half of the bit rate provides two different modulation schemes. Fig.2.3.a represents the MSK modulator considered as Staggered QPSK with half sinusoidal pulse shaping. The MSK modulator is analogous to that of QPSK except the offset delay and the cosine pulse shaper. Therefore, by demultiplexing the input data stream, the odd bit stream and the even bit stream are generated. Then, the odd bit stream $S_I(t)$ is multiplied with a half-sinusoidal pulse $\cos(\pi t/2T_b)$, while the even bit stream $S_Q(t)$ is delayed by one bit and multiplied by a half-sinusoidal pulse $\sin(\pi t/2T_b)$. These sinusoidally shaped bit streams are then multiplied by the orthogonal carriers and the resulting signals of the I-channel and the Q-channel are subtracted to yield the MSK signal expressed in Eq(2.15).
MSK DEMODULATOR

Fig. 2.3.b shows the MSK demodulator. The matched receiver for a distortionless channel consists of two correlators followed by an integrator. When assuming no noise and no ISI, the input MSK signal is

\[ Y(t) = S_I(t) \cos \frac{\pi t}{2T_b} \cos \omega_c t - S_Q(t) \sin \frac{\pi t}{2T_b} \sin \omega_c t. \quad (2.16) \]

then after the integration operation, the integrator outputs

\[ Y_I(t) = \int_{(2k-1)T_b}^{(2k+1)T_b} Y(t) \cos \frac{\pi t}{2T_b} \cos \omega_c t \, dt. \quad (2.17) \]
\[ Y_Q(t) = \int_{(2k)T_b}^{(2k+2)T_b} Y(t) \sin \frac{\pi t}{2T_b} \sin \omega_c t \, dt. \]

when these outputs are sampled at even or odd integer of bit duration, the sign of the sampler output decide the binary state of \( S_I(t) \) or \( S_Q(t) \). Then, by passing the sampler output through the decision threshold device and by multiplexing, the desired transmitted data stream can be obtained.
Figure 2.3: (a) MSK modulator (b) MSK demodulator.
CHAPTER 3 FILTERING AND NONLINEARITY

3.1 INTRODUCTION

Among the problems encountered in the digital transmission of data, inter-symbol interference (ISI) and adjacent channel interference are the most critical ones. In addition, the degradation by filtering and the nonlinear effect caused by the TWTA in the satellite transponder are also the object of this study. In this chapter, the first part is devoted to ISI and adjacent channel interference in the filtering process, then the nonlinearity of the TWTA is studied.

3.2 ISI AND ADJACENT CHANNEL INTERFERENCE

ISI is a direct result of the bandwidth limitation of the channel when a sequence of pulses is filtered; because a single pulse spreads into other bit intervals, there will be interference introduced between successive bits, therefore the current bit will be affected by the tails of the past bits as well as those of the next bits. This leads to the fluctuation of amplitude when these pulses are sampled synchronously. Shaping the signal for transmission to minimize ISI is the design objective of filtering. For ISI-free transmission, Nyquist proved that it is possible to have a de-
formed bandlimited wave and have a receiving device that receives and regenerates a perfect signal. i.e, if synchronous pulses, having a transmission rate of $f_s$ symbols per sec, are applied to an ideal linear phase lowpass channel with constant amplitude response and its cutoff frequency $f_N = f_s/2$ Hz, then the response can be observed independently without ISI. [Feher.9]

Unfortunately this kind of filter is not realizable, because to achieve the infinite attenuation slope of the Nyquist Minimum Bandwidth Filter, an infinite number of filter sections are needed and the slow decaying lobe of the time response also causes a relatively large ISI if the sampling instant is not perfect. For the realization of a Nyquist type filter, it is necessary to add a skew symmetric real-valued function to the ideal lowpass filter. The Raised Cosine Filter satisfies this condition. Details on this will be dealt with later.

In the linear channel, the ISI can be considered independent of thermal noise. But in a nonlinear channel, the ISI is a composite function of transmit signal, noise and nonlinearity. In a transmission systems using the constant envelope digital modulation technique, the amplitude nonlinearity in the nonlinear amplifier causes the spectrum spreading. Because the filtered sideband of the carrier will resume their original level after passing through the nonlinear amplifier, they may cause adjacent channel interference. The satellite
output multiplexer and the earth station receive filter are selected to reduce this interference.

3.3 FILTERING IN THE LINEAR/NONLINEAR CHANNELS

Based on the Nyquist Minimum Bandwidth Theorem, the Raised Cosine Filter approximated with the raised cosine function is the closest one to the ideal filter for ISI-free transmission. The frequency response of the Raised Cosine Filter is

\[
H(f) = \begin{cases} 
1 & 0 < f < f_N - f_X \\
1/2(1 - \sin \frac{\pi}{2r}(f/f_N - (1 - r)) & f_N - f_X < f < f_N + f_X \\
0 & f_N + f_X < f 
\end{cases}
\]  

(3.1)

where \( r = f_X/f_N \): roll-off factor
\( f_X \): arbitrary frequency
\( f_N \): Nyquist frequency.

The above frequency response consists of a flat amplitude portion and a sinusoidal roll-off portion. The frequency response can be specified with the roll-off factor \( r \) which can be expressed as the rate of the used bandwidth to the Nyquist bandwidth. Therefore \( r = 0 \) means that minimum bandwidth is the Nyquist frequency itself, while \( r = 1 \) indicates the total bandwidth used is the twice the Nyquist bandwidth. i.e. the increase of \( r \) means an increase in bandwidth used. Decreasing...
r makes the symbol timing and channel equalization more difficult.

The Nyquist filter approximated above assumes the transmission of impulses, not finite width pulses. This restriction inevitably leads to the transformation of the input signal other than impulse for ISI-free transmission. The Nyquist Generalized Theorem [Feher.9] gives the basis for the necessity of an amplitude equalizer on the assumption of zero phase response. For an impulse transmission, the Fourier transform of an impulse is constant over all frequencies, but for an arbitrary pulse, it is no longer constant. For the rectangular pulse such as that in an NRZ signal, the Fourier transformation has \( \text{SIN}(X)/X \) form. Therefore, to get the same response from the Nyquist type filter as in the impulse excitation, it is necessary to transform the rectangular pulse to have a constant spectrum. Therefore an amplitude equalizer with \( X/\text{SIN}(X) \) form is needed for the rectangular pulse transmission such as NRZ data in QPSK. In the MSK case, a different form of equalizer is needed because of sinusoidal pulse shaping rather than rectangular pulse shaping in QPSK.

For a linear channel, the raised-cosine filter is divided evenly between the transmitter and the receiver, that is, the transmit filter is a square root raised-cosine filter and the receive filter is also a square root raised-cosine filter. [Haykin.11] But in a nonlinear channel, it may not be a good choice. This is the case because it is possible to
eliminate ISI by equalizing the signal at the receiver input and ISI can be dealt with independently with thermal noise in a linear channel, while ISI is the composite function of thermal noise and filter characteristics in a nonlinear channel.

For the optimum filtering in a nonlinear channel, there have been many results and recommendations. Jones[12] reported that instead of the sharp roll-off Nyquist type filter, selection of a little wider bandwidth non-Nyquist type or smoother roll-off Nyquist type filter gives better performance, and the amplitude equalizer does not significantly affect the final performance, unless the system is for frequency reuse purposes. On the contrary to the conventional results from many papers, G.Satoh [10] stated that square root Nyquist type filter with milder roll-off (r = 0.47—0.5) gives the best performance even for the nonlinear channel.

Some of the result from other papers and the results above reveal the fact that regardless of the model of nonlinearity, smoother rather than sharp roll-off Nyquist type filters give better performance and the required bandwidth must be increased comparing with that of linear case. In addition, the combination of square root raised-cosine filter in the transmitter with non-Nyquist type filter such as Butterworth or Chebyshev filter in the receiver was reported as a good alternative for the nonlinear channel.
3.4 NONLINEARITY

Typical satellite communication channels incorporate nonlinear elements like an HPA in the earth station and TWTA in the satellite transponder. For simplicity of discussion, the nonlinearity in this section is confined to a TWTA of the Helix type which contains no memory (while Klystron and Coupled Cavity TWTA have memory), mainly because there is no general treatment of the nonlinearity for devices having memory. Since the nonlinear device has a bandwidth much higher than the possible rate of the envelope change, these nonlinear devices can be safely assumed to be memoryless. When nonlinear devices exist in the channel, the transmission analysis becomes complex and simulation is often the only practical approach to estimate the channel performance. A nonlinear TWTA exhibits two types of nonlinearities: amplitude nonlinearity or AM/AM conversion and phase nonlinearity or AM/PM conversion.

AM/AM conversion effect and AM/PM conversion effect occur when a time varying input bandpass signal enters the TWTA. The envelope fluctuation in the input signal caused by filtering is converted to an output signal amplitude variation and phase shift when this fluctuating input signal is amplified by the nonlinear TWTA. In the power limited satellite, for operation near or in saturation, the AM/AM conversion effect is dominant. These effects can be expressed simply
as following. [Palmer.2] Let the input signal in complex form be

\[ E_{\text{in}}(k\Delta t) = \sqrt{S_{\text{in}}^*(k\Delta t)S_{\text{in}}(k\Delta t)}, \]  \hspace{1cm} (3.2)

where \( S_{\text{in}}(k\Delta t) \) is the input signal complex envelope. If we express the AM/AM conversion function as \( A[E_{\text{in}}(k\Delta t)] \), and the AM/PM conversion function as \( P[E_{\text{in}}(k\Delta t)] \), then the output envelope and phase shift for each input envelope value are expressed by

\[ E_{\text{out}}(k\Delta t) = A[E_{\text{in}}(k\Delta t)] \cdot \exp[jP[E_{\text{in}}(k\Delta t)] + j\phi_{\text{in}}(k\Delta t)]. \] \hspace{1cm} (3.3)

But the expression of the nonlinear function above in analytic form needs complex mathematical steps and approximations. All the reported methods for modeling these functions can be classified into the amplitude model and the quadrature model. The former needs amplitude and phase functions, whereas the latter needs in-phase and quadrature functions as its model. The assumption of a memoryless TWTA permits an approximation with a power series expansion, but the series expansion needs high accuracy up to saturation level to get the coefficients and parameters of the series. Eric[Eric.5] also used a power series expansion for his quadrature model, but his model has been reported to have
problems in expressing the effects at very large inputs. [Saleh.3] Hettrakul and Taylor [Het.4] use two parameter formulas with a quadrature model involving modified Bessel function of the 1st kind. Among those reported models of nonlinearity, Saleh's model is the simplest in notation and gives good agreement with the measured data for a TWTA.

Saleh [Saleh.3] proposed a simple two parameter formula for the Amplitude-Phase model or the Quadrature model. In his formula for the amplitude and phase model, if the input signal to the TWTA is written as

$$x(t) = r(t) \cos(\omega_c t + \varphi(t)), \quad (3.4)$$

where \(r(t)\) is the input signal envelope, \(\varphi(t)\) is the phase offset, and \(\omega_c\) is the carrier radian frequency, then the output is written as

$$y(t) = A[r(t)] \cos (\omega_c t + \varphi(t) + \psi[r(t)]), \quad (3.5)$$

where \(A(r)\) represents the AM/AM conversion function, and \(\psi(r)\) represents the AM/PM conversion function. The proposed two parameter formula for the amplitude nonlinear function, \(A[r(t)]\) and the phase nonlinear function \(\psi[r(t)]\) are written as
As can be seen from Eq(3.6) and Eq(3.7), \( A(r) \) approaches \( 1/r \) and \( \phi(r) \) approaches a constant for very large values of \( r \).

To use this formula, it is necessary to find the four parameter values for a specific TWTA model from the measured data or from the characteristic curves. Saleh also gives these parameter values for some frequently used models for simulation. [Saleh.3]

3.5 FFT(FAST FOURIER TRANSFORM) AND IDFT(INVERSE DFT)

The filtering process is basically the time convolution of the input signal and the filter impulse response. But in simulation, this approach is not preferable because of the computation time. Therefore the method of multiplying the discrete Fourier transform of the time sampled signal by the filter frequency response in the frequency domain, and returning this result to the time domain, is frequently used.
The Discrete Fourier Transform can be thought of as a discrete version of the continuous Fourier transform under some restrictions since:

- the continuous signal must be truncated to a finite length.

- within this interval, the signal must be available as the sequence of $N$ equally spaced values, where $N$ is the number of time samples.

- this truncated interval can be extended periodically to yield the discrete harmonics.

By the periodicity of the result, the accuracy of the DFT is greatly affected by aliasing, but this effect can be eliminated with the selection of a high sampling rate. Because the DFT needs $N^2$ computation steps, it is almost impossible to calculate if $N$ is a large number. The FFT is based on the Cooly-Tukey algorithm, in which, instead of $N^2$ computation, only $(1/2)N\log_2 N$ steps are necessary with the $N$ discrete frequency components made from $N$ discrete time samples using the circular symmetry of the exponential function. The restriction that $N$ be a power of 2 can be avoided by substituting additional zeros if the sample number doesn't fit this condition.
IDFT is the reverse procedure of FFT as in the case of inverse Fourier transform of continuous signal. It is the way to return frequency transformed values to the time domain under the condition that the FFT values and the IDFT values are complex conjugates and a 1/N scaling factor is applied to all the IDFT values. Detailed explanation can be found in the Brigham's in Chapter 9 and 10. [Brigham.8] Some basic properties of the FFT are listed below for the validity of the FFT results. [Stremler.7]

- if there were N time samples, there also exist N discrete frequency components.

- by the property of symmetry, the sample point Oth and Nth are identical in both domains.

- the frequency components from Oth to N/2th are positive frequencies, whereas N/2th to Nth are negatives.

- for the real valued function, the positive and the negative components are complex conjugate, but the Oth and the N/2th are common to both domains and are real values.

- the highest frequency component is 1/2T Hz and frequency spacing is 1/NT, where T is the sampling time interval.
4.1 SIMULATION OVERVIEW

Most of the channel modeling programs developed for the performance evaluation of communications satellites allow the channel to be modeled from a block diagram bases. To model and simulate the channel, it is fundamental to perform four major steps as signal generation, filtering, linear or non-linear operation, and demodulation regardless of the modulation scheme used.

The simulation program used in this thesis consists of four major parts: the pseudo random signal generator, the modulator and demodulator, the filter and the nonlinear TWTA. These functional blocks can be located as transmit(Tx) earth station, satellite transponder and receive(Rx) earth station. In addition, some programs like BER calculator and General Plotter are also made for the analysis of results. Fig.4.1 shows functional blocks of the single link used in simulation. Even though two kinds of modulation schemes, QPSK and MSK, are used, most of the elements can be used in common.

The first step in simulation is the signal generation. The pseudo randomly generated NRZ pulse sequence is applied to the QPSK/MSK modulator, where even and odd bit streams re-
Figure 4.1: Simulation block diagram.
spectively modulates the two orthogonal carriers and sinusoidal pulse weighting is done for the MSK case. These modulated carriers are then added and passed through the Tx filter. Before filtering, amplitude equalizing with a specific function depending on QPSK or MSK is performed (Pre-filtering). In filtering, the fast Fourier transform of time sampled data of the modulated signal is multiplied by the specific filter frequency response. All the filters' phase responses are assumed to be zero for simplicity. The filtered signal is obtained by returning the above multiplied frequency domain signal into the time domain by using the inverse discrete Fourier transform. In this step, to get the phase and amplitude information of the filtered signal easily, filters are modelled with their low-pass equivalents, and all the inputs and the outputs of the filters are expressed as complex envelopes. By using the complex envelope notation, it is possible to do the simulation in baseband regardless of the carrier frequency used. From the complex envelope of the filtered output, the amplitude and the phase are easily obtained. This procedure is needed especially when the amplitude and phase model is used for the TWTA model.

Because we already assumed that the earth station HPA is operated only in the linear region, the filtered signal is linearly amplified and transmitted to the satellite transponder. In the satellite transponder, the uplink noise is
assumed to be smaller than the downlink noise. Therefore, only the downlink noise is considered. Because the bandwidth of the input/output multiplexer filters is normally larger than that of the signal, they are excluded in the single link simulation at first. But the addition of these filters for future usage can be done easily by inserting one command line in the program. The signal is amplified by the TWTA using the amplitude and phase model described in Sec 3.4. The retransmitted signal from the satellite passes through the Rx filter in the earth station. This Rx filter shapes the spectral components of the signal which is demodulated by carrier multiplication and sinusoidally pulse shaped for MSK. The baseband signal is then low-pass filtered and time sampled by the sampler.

For the BER calculation, the signal power is calculated at the input of the Rx filter at first, then for the desired value of carrier to noise ratio (CNR), the noise power must be calculated. To do this, instead of direct addition of noise (downlink thermal noise) at the input of Rx filter, analytically calculated noise power is added in the simulation and an average BER as a function of $E_b/N_0$ (energy per bit/noise power spectral density) is found. The results of the simulation are plotted by the general plotter, finally.

Throughout the simulation procedure, the signal that is generated and processed is real-valued, even if it takes a com-
plex form in some cases for transformations. At the signal generation, the real-valued signal is time sampled with the rate of 16 samples for one bit duration. Therefore the discrete power spectra at the modulator would span 8 times the bit rate in one (+ or - ) direction of frequency, for example, for the bit rate of 70 Mbps, the frequency spectra will span from -560MHz to +560MHz. The time sampled signal data are converted to frequency sampled data in a filtering process by the FFT. After multiplying with the filter frequency response, the filtered result is converted to the time domain by the IDFT. This procedure of filtering is the same for every filter used in simulation. The signal used in the TWTA is real-valued in the time domain. Using the same transformation procedure as the transmit case, the demodulated and time sampled value is the real-valued signal. In sampling at the demodulator, samples should be taken at the sampling point within each symbol where decision metrics are derived. Sometimes this procedure require linear polynomial interpolation between samples.

4.2 MODELING PROCEDURE DESCRIPTION

This section is devoted to the detailed description of simulation elements based on the block diagram in Fig.4.1. For convenience, the description is divided into 3 sections as
4.2.1 TX EARTH STATION

Fig. 4.2 depicts the block diagram of the typical Tx earth station. In the diagram, the physical devices like the up-converter and the earth station HPA are excluded on the assumption that the earth station HPA is operating in the linear region and all the group delay characteristics are compensated by the equalizer. Therefore the earth station consists of a symbol generator, QPSK/MSK modulator, and a Tx band-pass filter with amplitude equalizer. Because we have assumed zero phase response for the filter, the equalizing is needed only for the amplitude.

SYMBOL GENERATOR

To generate the input binary bit stream, the algorithm of a basic uniform pseudo random number generator routine 'G8UBS' in IMSL (International Math and Statistics Library) was modified for this purpose. With the provided value of output range, an integer value of DSEED, and a random number denoted by NR1(integer), this routine outputs a sequence of random binary number 0 or 1 with uniform distribution. By assigning all the 0s as -1, we can get the pseudo random sequence of
an NRZ signal with its amplitude +1 or -1. By demultiplexing this sequence, the even and the odd ordered bits are grouped into separate bit streams by assigning the 1st bit in the original sequence as 0th bit. For example, if the generated sequence of 8 bits was -1, -1, +1, -1, +1, -1, -1, +1, then, the even bit stream consists of -1, +1, +1, -1 and odd is -1, -1, -1 +1. To achieve the sampling effect of 16 samples per data bit duration, the even and the odd data stream bit rates are increased by a factor of 16. This means that the one bit value of each stream is duplicated 16 times, therefore -1 is changed to -1, -1,........., -1(16 ea) and the same for the +1 case. For notational convenience, the original even bit stream is called the I channel data and the odd bit stream is called the Q channel data. To deal with MSK, it is necessary to skew the above two bit stream by one bit duration in time. For QPSK, the same alignment of time is used, so no modification is needed. Therefore a one bit delay effect for the second bit stream is achieved by the above skewing. In simulation the I-channel data are made to advance by 1 bit duration, rather than delaying the Q-channel data. As a result, the I channel data of MSK has a vacant entry in its 16th position, and this value is filled with a dummy value for convenience.
Figure 4.2: Tx earth station block diagram.
QPSK/ MSK MODULATOR

Basically the modulators for the two schemes are the same except for two facts. MSK needs additional sinusoidal pulse shaping and the bit streams used in MSK are skewed by 1 bit duration of each other. Because the latter is already done in the symbol generation procedure, only the former is needed in MSK signal generation.

The even and the odd bit streams applied to the modulator with half of the original incoming bit rate modulate the in-phase and the quadrature phase components of the carrier. After carrier multiplication, these two orthogonal components are added. When the input data stream is applied to the modulator, it is necessary to make the carrier and shaping function (MSK only) into the time sampled form to get the time sampled data of modulated carrier. Therefore the modulated signal can be expressed with I-channel data stream $S_I(t)$ and Q-channel data stream $S_Q(t)$ as

$$x(t)_{QPSK} = S_I(t)\cos \omega_c t + S_Q(t)\sin \omega_c t. \quad (4.1)$$

$$x(t)_{MSK} = S_I(t)\cos \frac{\pi t}{T_s} \cos \omega_c t + S_Q(t)\sin \frac{\pi t}{T_s} \sin \omega_c t. \quad (4.2)$$

The two terms in Eq (4.1) represent the two BPSK signals which can be detected independently due to the orthogonality...
of the carrier. In the program, the above QPSK output is expressed as

\[ X(t) = \sqrt{2} \cos \left( \omega_c t - \theta(t) \right), \]  

(4.3)

where \( \theta(t) = \tan^{-1}(S_Q(t)/S_I(t)) \) and \( \theta(t) \) are 45, 135, 225, 315 degrees depending on the combination of I channel input \( S_I(t) \) and Q channel input \( S_Q(t) \) as \((1, 1), (-1, 1), (-1, -1), (1, -1)\). In MSK, the above Eq(4.2) can be simplified by using the trigonometric function and the fact that I and Q channel data are always \( \pm 1 \), then

\[ x(t) = \cos \left( \omega_c t + b_k(t) \frac{\pi t}{T_s} + \phi_k \right). \]  

(4.4)

where \( b_k \) is 1 when I-channel and Q-channel data have opposite sign and -1 otherwise, and \( \phi_k \) is 0 or \( \pi \) when the I-channel data is +1 or -1. As stated before, to do the simulation in the baseband signal, it is necessary to express these modulator outputs with their complex envelopes at first. By doing so, we can easily use the baseband signal component for further process like filtering and nonlinear amplification.

**MODEM TX FILTER**

The design and selection of channel filters plays an important role in the system performance. All the filters simu-
lated here are assumed to have zero phase response for simplicity and are low-pass equivalents of their band-pass responses. To simulate the filtering process, we need to have time-sampled values of input signals and the filter function expressed in time and frequency domains. Therefore if \( x(t) \) is the modulator output signal and the filter impulse response is \( h(t) \), the filtered output in the time domain will be expressed by

\[
y(t) = x(t) * h(t).
\]

The Fourier transform of \( y(t) \) is

\[
Y(f) = X(f) H(f).
\]

Because the modulator output can be expressed as a complex envelope

\[
z(t) = x_r(t) + jx_i(t),
\]

the corresponding Fourier transform of the complex envelope of filter output will be

\[
\hat{Y}(f) = X_r(f)\hat{H}(f) + jX_i(f)\hat{H}(f),
\]

where \( \hat{Y}(f) \) and \( \hat{H}(f) \) are Fourier transform of the complex envelope of the filter output and the filter impulse response. The filtering starts from the transformation of the time-sampled value of modulator output signal into a frequency domain value by the FFT. Because the FFT and IDFT routine is programmed to accept and output the complex values, if the real-valued signal is to be processed, it must be expressed in complex form first by substituting zero in its imaginary part. Multiplying the two frequency responses and using the
IDFT yield the desired filtered output in the time domain expression. Because of the property described in Sec. 3.5, the filtered output from the IDFT has the same number of time samples before filtering. The simulated Tx filter is a low-pass equivalent of the band-pass filter. Because of the zero phase response assumption of filter, the low-pass equivalent filter frequency response is symmetric. By using the complex envelope of the baseband signal as an input to the desired filter and multiplying this baseband complex envelope without carrier component with the frequency response of the low-pass equivalent filter corresponding to the desired band-pass filter, the filtering process is made much simpler and easier to manipulate. The inclusion of the carrier contribution can be easily done at any desired point in the simulation. This is the baseband simulation and is preferred because of its simplicity.

Five kinds of analog band-pass filters are simulated in the program with their low-pass equivalent frequency response. Those are the Raised-Cosine Nyquist Filter, Square Root Raised-Cosine Nyquist Filter, Butterworth Filter, Chebyshev Filter and conventional Low-pass (brick-wall) Filter. The first two Nyquist filters are already described in detail in Chapter 3 and the characteristics of the other filters can be found easily in filter design handbooks, so the description will be omitted here. Because of the difference in the pulse shaping between QPSK and MSK, the amplitude
equalizer must be different for each case. Therefore, to simulate the filtering process, the parameter indicating the modulation type must be provided at first. Then the desired filter type, and the mode which indicates whether the filter is used in Tx or in Rx part, and the option that amplitude equalizer is included or not are also provided as well as the basic parameters like roll-off value, Nyquist frequency, number of poles, ripple value and bandwidth product.

AMPLITUDE EQUALIZER

Prior to passing the modulated signal to the filter, amplitude equalizing is necessary to satisfy the Nyquist Theorem. Because this theorem is applicable only for the impulse transmission, a more generalized theorem by Nyquist states that for the transmission of arbitrary pulses other than impulses, pre-filtering in which the spectrum of the input signal is made constants, is necessary to keep the same frequency response of the channel as for impulse excitation. [Feher.9] Therefore in QPSK, where the excitation is a rectangular pulse, the \( \frac{X}{\sin X} \) with \( X = \frac{2\pi f T_s}{2} \) type equalizer is used. Because \( T_s = \frac{1}{R_s} \), \( \frac{X}{\sin X} \) can also be expressed with radian frequency \( \omega \) and bit rate \( R_b \) as

\[
\frac{x}{\sin x} = \frac{\omega R_b}{\sin (\omega / R_b)}. \tag{4.5}
\]
But in MSK, the excitation is neither impulse nor rectangular pulse. Because the pulse shaping function in MSK is half wave sinusoidal as $\cos(\pi/T_s)t$, the Fourier transform of this function is given as

$$F[\cos\frac{\pi}{T_s}t] = \frac{2T_s}{\pi} \frac{\cos(\pi f T_s)}{1 - (2f T_s)} \text{ for } \frac{T_s}{2} < f < \frac{T_s}{2}. \tag{4.6}$$

To make this spectrum into constant, the required pre-filtering must be the inverse form of the transformed one above as

$$\frac{1 - (2f T_s)^2}{\cos(\pi f T_s)}. \tag{4.7}$$

4.2.2 SATELLITE TRANSPONDER

The typical transponder contains an input multiplexer filter, a TWTA, and an output multiplexer filter. But for a channel in single carrier operation, the everything except the TWTA can be neglected because the key point to this simulation is the inclusion of the nonlinear effects of the TWTA, like AM/AM and AM/PM conversions. Therefore we can confine the modeling of the satellite transponder to the nonlinearity of TWTA only. As stated earlier in Sec.3.4, the Amplitude and Phase model of Saleh [Saleh.3] is used. The modeling proce-
The input signal applied to the TWTA is always a band-pass signal. This signal has the new phase and envelope different from the modulated signal because it has passed through the band-pass filter. If a channel filter exists, the result is only the cascading of filter functions. In baseband simulation, the modulator output expressed in its complex envelope passes through the low-pass equivalent of the desired band-pass filter, and the filtered output also can be expressed by its complex envelope. The usual expression for the input signal to the TWTA is

\[ x(t) = r(t)\cos [\omega_c t + \psi(t)], \quad (4.8) \]

because only the complex envelope of \( x(t) \) is dealt with in baseband simulation, if we use the fact that \( \cos \theta = \text{Re}(e^{j\theta}) \), then the cosine term of \( x(t) \) will be

\[ \cos(\omega_c t + \psi(t)) = \text{Re}(e^{j\psi(t)} \cdot e^{j\omega_c(t)}), \quad (4.9) \]

therefore \( x(t) \) can be written as

\[ x(t) = \text{Re}(r(t) e^{j\psi(t)} \cdot e^{j\omega_c(t)}), \quad (4.10) \]
then the complex envelope of $x(t)$ denoted by $z_1(t)$ can be written

$$z_1(t) = r(t) e^{j\psi(t)}.$$ (4.11)

When the input band-pass signal expressed in its complex envelope $z_1(t)$ passes through the nonlinear TWTA, the output will be expressed as Eq(3.5). By using the similar manner, the output of the TWTA can be expressed in its complex envelope from

$$z_2(t) = A[r(t)] e^{j\psi(t)} e^{j\theta[r(t)]},$$ (4.12)

therefore the amplified amplitude is expressed as $A[r(t)]$ and the phase will be denoted as $\theta(t) = \psi(t) + \theta[r(t)]$. The AM/AM nonlinear function $A[r(t)]$ and the AM/PM nonlinear function $\theta[r(t)]$ are the same as Eq(3.6) and Eq(3.7) in Sec.3.4. But to get the practical output for the specific TWTA, the four parameters that represent the relations of time varying input and the nonlinear amplitude function and nonlinear phase function, $\alpha_a$, $\alpha_\phi$, $\beta_a$, $\beta_\phi$ (refer to the Eq(3.6) and Eq(3.7) in Chapter 3) must be decided by curve fitting with the measured data. In the simulation done in Sec.5.1 and Sec.5.2, the following parameter values which are
derived from the curves of Kaye, George, and Eric [5] by using the proposed two nonlinear formula of Saleh[3] were used.

\[
\begin{align*}
\alpha_a &= 2.1587 \\
\alpha_\phi &= 4.0033 \\
\beta_a &= 1.1517 \\
\beta_\phi &= 9.1040,
\end{align*}
\] (4.13)

Whereas for the test done in Sec.5.3, the above four parameter values are found from the characteristic curves by using minimum mean-square error curve fitting procedure. Details will be explained in Sec 5.4. Once the nonlinear modeling is finished, it is necessary to normalize the input envelope voltage (or power) to the saturation level. Many paper use different scales for the axis of nonlinear curve such as dBm, or Volt(rms or peak). If we use volts as the scale, the unit of the input and output will be the volt. Saturation level is set as 0 dB or 1 volt assuming 0 dB back-off. Depending upon the system requirement, it is frequently required to move the operating point from saturation to the desired point to improve the bit error rate. This is called back-off. The usual way to quote the input back-off is dB. For example, an IBOF( input back-off ) of 3 dB means the input level is reduced to the one half of the operating point for saturation of the TWTA output. Similarly the OBOF( output back-off ) of 3 dB means the output power is reduced to the half of the saturated level, i.e. the voltage is reduced to the 0.707 times of the saturated output voltage level.
Therefore for the general case, the $X$ dB IBOF leads to the calculation of the new input operating point given by

$$r(t)_{IBOF} = 10^{-X/20},$$  \hspace{1cm} (4.14)$$

therefore the input back-off $A[r(t)]$ and $\phi[r(t)]$ are easily found by substituting the argument $r(t)$ with $r(t)_{IBOF}$. For $X$ dB OBOF, the corresponding input level (Voltage) $r(t)$ is derived from the amplitude and phase model's nonlinear function as

$$A[r]_{OBOF} = 10^{-X/20}. \hspace{1cm} (4.15)$$

When we substitute $r$ in Eq(3.6) with $r_{OBOF}$, we get the relation as

$$A[r]_{OBOF} (1 + \beta_a r^2_{OBOF}) = a r_{OBOF}. \hspace{1cm} (4.16)$$

By solving this 2nd order equation for $r$ and taking the left side root to yield the desired output back-off

$$r_{OBOF} = \frac{a - \sqrt{a^2 - 4A[r]^2 \beta_a}}{2A[r] \beta_a}. \hspace{1cm} (4.17)$$
4.2.3 RX EARTH STATION

The receiving earth station mainly performs down-converting and demodulation. The Rx band-pass filter, the demodulator, the sampler are the major elements of the receiving earth station model. After the receiving procedure, BER calculation is done finally to evaluate the performance. The non-linearly amplified signal in the transponder is band-limited by the output mux filter to reduce the out of band spectral components. This signal is also corrupted by downlink noise and ISI. Assuming noise-free conditions, the received signal in time sampled form is converted to the frequency domain by the FFT and multiplied with the discrete frequency components of the receiving band-pass filter transfer function. Then converted to time domain by the IDFT as in the Tx filter case. In the demodulator, this filtered signal is multiplied by the locally generated carrier and the sinusoidal pulse shaping function in the MSK case. The locally generated carrier has the same frequency as that of the modulator. The demodulated signal passes through the integrator in which integration is carried out for one symbol duration. This carrier multiplication and integration in the linear noise-free condition gives the optimum correlation receiver for the infinite bandwidth signal. But for the bandlimited signal like the practical ones, the integrator is replaced with a conventional low-pass filter. The demodulated signals in the
orthogonal channels are time sampled with specific rate for which the sampled values are in the most stable condition. Because there is no fixed sampling time, it is necessary to find the optimal sampling time empirically until the used sampling time ensures minimum bit error.

The correlation receiver stated above is optimum in linear noise-free conditions, and the square root raised-cosine Nyquist filter is the optimum filter in the linear channel. But in the channel where nonlinearity exists, this receiver no longer ensures the best performance. Instead, the selection of a simple Non-Nyquist filter or a Nyquist filter with softer roll-off is preferred. In simulation, to get the relative performance for different combinations of Tx/Rx filters, the combination of square root raised-cosine Nyquist filter pairs, or square root raised-cosine Nyquist filter in Tx part and Butterworth filter in Rx part are used. As mentioned earlier in this section, theoretically the thermal noise should be added at the input of the receive band-pass filter, and this signal is demodulated and sampled, then compared with the known Tx symbol, the error is counted to get error rate. This is called a direct simulation method and usually used when there is a nonlinear element in the detection or modulation process itself. This method is computationally time consuming and inefficient because error counting is needed for all the symbols transmitted. The hybrid simulation [Palmer.2], mentioned briefly earlier, is
used in this paper, where the noise (thermal noise) is neglected at first, and we proceed with the same steps of demodulation and sampling. After getting the detected value, the analytically calculated noise power is used to get the BER for the fixed carrier to noise ratio (CNR).

To get the desired amount of noise power expressed as power spectral density \( \frac{N_0}{2} \), the equivalent noise bandwidth of the receive filter is usually used. But in M-ary PSK, all the performance measures are based on the bit or symbol. Therefore the required bit energy \( E_b \) to noise density ratio \( N_o \), \( \frac{E_b}{N_o} \) is more convenient. Feher[Feher.9] described the relationship of \( C/N \) to \( \frac{E_b}{N_o} \) as

\[
\frac{E_b}{N_o} = \frac{C}{N} \frac{B_w}{R_b}
\]  

(4.18)

i.e. the \( \frac{E_b}{N_o} \) ratio is proportional to the ratio of receiver noise bandwidth, \( B_w \), to the bit rate \( R_b \) for the fixed \( C/N \) ratio on the conditions that Tx and Rx filters are matched filters and noise is assumed to be the additive white gaussian. Depending on the modulation scheme and the type of detector used, \( \frac{E_b}{N_o} \) required for the given probability of error may vary from 7 to 15 dB in general. The performance can be expressed in terms of the probability of bit error or the Bit Error Rate (BER). For the QPSK and the MSK cases, on the condition that \( C/N \) is high enough to avoid multiple bit
errors per symbol error, the the probability of bit error is expressed as follows.

\[ P_e = P_{e_{\text{BPSK}}} = P_{e_{\text{QPSK}}} = P_{e_{\text{MSK}}} = \frac{1}{2} \text{erfc} \sqrt{\frac{E_b}{N_0}} \quad (4.19) \]

where

\[ \text{erfc}(x) = \frac{2}{\sqrt{\pi}} \int_x^\infty e^{-y^2} \, dy \quad \text{for } x > 0, \quad (4.20) \]

\[ E_b = C \cdot T_b \]

\[ T_b = \text{bit duration}, \quad C = \text{received signal power}. \]

To use the above formula for the calculation of the average BER, the received signal power \( C \) and the noise density \( N_0 \) are found at first. To do this, The normalization of the received signal (in noise-free condition) to the average signal level \( C \) of 1 watt at the input to the Rx filter gives the convenient reference level. To calculate the noise spectral density \( N_0 \) at the same point for arbitrary ratio of \( E_b/N_0 \), Eq(4.18) and the relationship of \( N = N_0 \cdot B_w \), where \( N \) is the noise power in C/N notation, are used. If we set \( E_b/N_0 = x \) dB, \( N_0 \) can be calculated on the conditions that \( C \) is set to 1 watt and \( B_w \) is designed to be the same as the symbol rate. Therefore \( N_0 \) can be calculated easily from Eq(4.18). Once \( N_0 \) for a specific C/N ratio was found, the average BER can be calculated from Eq(4.19) by finding the set of \( E_b \) values (volts) from the sampler outputs, and from the relationship of \( E_b = C \cdot T_b \) [Imrich.6]. In simulation, only one channel, ei-
ther I or Q channel, can be used for BER calculation based on the fact that both in-phase and quadrature channels exhibit the same performance because each channel carries a BPSK signal and the probability of BPSK and QPSK are identical.
CHAPTER 5 RESULTS OF SIMULATION

5.1 SIMULATION BACKGROUND

Based on the specific algorithm described in Chapter 4, the simulation program which performs signal generation, filtering, modulation and demodulation, nonlinear amplification and BER calculation was developed. Because the program is made up of module units based on the block diagram, modification of block elements of the link, like insertion or exclusion of specific block, is directly related to the inclusion or the exclusion of subprograms. This flexibility permits various tests without changing the main stream of the simulation procedure. In simulation, some keystone parameters and certain conditions are kept for uniformness of simulation and for simplicity:

- The simulation is performed with a baseband signal, so it is baseband simulation. The effect of carrier frequency can be inserted at any desired point in the simulation.

- The simulation is performed only for the single channel with two kinds of modulation schemes: QPSK and MSK.
• There exist 32 time samples for the one symbol duration, $T_s$.

• 128 symbols (total 256 bits) are transmitted and received in the simulation, therefore a total of 4096 time samples are processed. This also means that same number of frequency samples are processed in the frequency domain.

• The nonlinear device is restricted to the memoryless TWTA in the satellite transponder.

• The overall group delay is neglected on the assumption that complete equalization is done at the receive earth station.

• The BER used in the simulation is the average BER.

• The optimum sampling point for the decision is simply set to the center of the symbol duration.

With the above conditions and system parameters, the modem back to back test was first performed for QPSK and MSK cases respectively to get the references for further tests. Then the nonlinear tests which incorporate the TWTA were done for QPSK and for MSK by changing the input/output back-off values to study the improvement in performance. All these result are

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plotted for comparison. As a final test to ensure the validity of the simulation program developed, the program was modified with practical data to meet the practical condition. Then the results are also plotted and compared with the results of previous ones.

5.2 QPSK/MSK MODEM BACK TO BACK TEST

The simulation program was first used to establish a back to back reference performance where the output of the modem Tx filter is directly applied to the modem Rx filter input, as depicted in Fig. 5.1. For the incoming bit rate of 70M bps, therefore the symbol rate is the half of the bit rate, 35M sps, the modem filters used are the so-called matched filters implemented with square root raised-cosine Nyquist filters with 50% roll-off. This is the ideal combination to get the modem back to back result. In these modem filters, the Nyquist frequency is one half of the incoming symbol rate, 17.5 MHz. To get the same filter response with impulse excitation, an x/sinx type amplitude equalizer is used for the QPSK and a sinusoidal amplitude equalizer is used for the MSK case in the Tx filters only. To compare the result of the test with theoretical values expressed in BER, the complementary error function is used. Because the theoretical probability of error, $P_e$, of coherent BPSK is same for QPSK

Results of simulation
and MSK, the results of the modem back to back tests are compared with the above mentioned theoretical values.

Fig.5.2 shows the result of the test and the theoretical value expressed in BER versus $E_b/N_0$ for QPSK and MSK cases. As can be seen from the curves, the QPSK modem back to back test result deviates about 0.2 dB from the theory at a BER of $1 \times 10^{-4}$, whereas the result of MSK is a little more deviated from the theory, about 0.3 dB at the same BER. In addition to that, MSK is a little inferior to QPSK, about 0.1 to 0.3 dB in general.

To see how much the amplitude equalizer affects the final result, same modem back to back were performed without the amplitude equalizer in the Tx filter. Fig.5.3 shows the effect. When compared with the equalized cases, about 1.0 to 1.5 dB degradation occurs for QPSK and MSK cases.

In this modem back to back test, the deviation of 0.2 dB in QPSK case and 0.3 dB in MSK case from the theoretical value can be due to the effect of quantitization noise inherent in computer simulation. But the degradation of MSK relative to QPSK of about 0.2 dB cannot be explained properly at this time. The fact that the main lobe of the MSK power spectrum is about 1.5 times larger than that of QPSK may be the reason for the degradation in comparison to QPSK, because in the tight time-bandwidth product used in this test, there exists the possibilities that some portion of the signal power can be lost in the MSK case.

Results of simulation
Figure 5.1: Block diagram of the modem back to back test used to get the results of Fig. 5.2 and Fig. 5.3 for QPSK and MSK.

Results of simulation
Figure 5.2: BER vs $E_b/No$ comparison between the theoretical value and the QPSK/MSK modem back to back test results in single channel based on Fig. 5.1.

Results of simulation
Figure 5.3: BER vs $E_b/No$ comparison of QPSK/MSK modem back to back test results between the amplitude equalized cases and the unequalized cases.

Results of simulation
5.3 THE NONLINEAR CHANNEL TEST

The result of the modem back to back test in the previous section is used as the reference for the nonlinear channel test. Fig. 5.4 shows the block diagram of the nonlinear channel test. In this diagram, the input and output multiplexer of the TWTA are excluded, because the bandwidth of these two filters is usually larger than the signal bandwidth. But later in the application to practical data dealt with in the next section, these filters are included to give more realistic conditions. Therefore the Tx band-passed signal is directly applied to the nonlinear TWTA in the transponder. In the TWTA, the AM/AM and AM/PM nonlinear effects are modelled with the Amplitude/Phase functions as described in Sec. 3.4. To see the nonlinear effects caused by the TWTA, various input or output back-off values of 0.5, 1, 2, 3, 6, 9, 12, 14 dB were used for QPSK and MSK respectively. The filters used in this test are the same as those used in the modem back to back test for convenience of comparison. Even though the square root raised-cosine Nyquist filter pair is rarely used in the nonlinear channel, this filter pair is selected to see the effect of the TWTA in the same structure as a linear channel except for the existence of TWTA.

Fig. 5.5 shows the results of the QPSK input back-off cases, and Fig. 5.6 shows those of the MSK input back-off cases. The
results of the output back-off cases of QPSK and MSK are depicted in Fig.5.7 and in Fig.5.8.

From Fig.5.5 of QPSK input back-off, we can see that QPSK requires an extra 4.2 dB of $E_b$ compared with the modem back to back test result to get the BER of $1 \times 10^{-4}$ in the case of 3 dB TWTA input back-off. But in the MSK input back-off case depicted in Fig.5.6, it needs more than extra 6 dB of $E_b$ to keep the same BER with same input back-off value. For the output back-off cases in Fig.5.7 and in Fig.5.8, QPSK deviates 3.2 dB and MSK deviates about 5 dB from the ideal value of modem back to back test result to keep the BER of $1 \times 10^{-4}$ when 1 dB output back-off is given. Fig.5.9.a and Fig.5.9.b show the relation of the required $E_b/N_0$ for the various input back-off values and output back-off values to keep the desired BER based on the curves of QPSK in Fig.5.5 and Fig.5.7 respectively.

5.4 PRACTICAL APPLICATION

The nonlinear single channel test performed in Sec.5.3 has the one additional element, TWTA, when it is compared with the modem back to back test case, i.e., they use the square root raised-cosine Nyquist filters with softer roll-off of 50% in Tx part and Rx part. In reality, this kind of filter combination is seldom used. Instead of these filters, Non-Nyquist type filter such as Butterworth or Chebyshev filters...
Figure 5.4: Block diagram of QPSK/MSK nonlinear single channel used to get the results of Fig.5.5 to Fig.5.8.

Results of simulation
Figure 5.5: BER vs $E_b/No$ comparison for QPSK nonlinear single channel under various TWTA input back-off.

Results of simulation
Figure 5.6: BER vs Eb/No comparison for MSK nonlinear single channel under various TWTA input back-off.
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Figure 5.7: BER vs $E_b/No$ comparison for QPSK nonlinear single channel under various TWTA output back-off.
Figure 5.8: BER vs $E_b/No$ comparison for MSK nonlinear single channel under various TWTA output back-off.

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Figure 5.9: Required $E_b/N_0$ for fixed BER of $10^{-4}$, $10^{-6}$ and $10^{-8}$ in cases of various TWTA back-off for QPSK. (a) input back-off cases. (b) output back-off cases.
are used either as a Tx or a Rx filter. In addition to that, nonlinear tests done in Sec.5.3 did not include the input and the output multiplexers. Therefore, those test results we got in the previous section through the modem back to back test and nonlinear test can not be the practical values in a strict sense. They are only the ideal values we can get from the simulation program developed. To make practical use of the simulation program developed, it is thus necessary to run the program with the practical data which are derived from the practical devices used in the current satellite communication systems. For this purpose, some practical data such as filter frequency response curves, TWTA AM/AM and AM/PM characteristics were collected to be used as the model for the simulation. Following are the detailed procedure of modeling to do the modem back to back test and the nonlinear test with practical data. Fig.5.10 shows the block diagram of the overall single channel to be simulated.

1. The incoming bit rate $R_b$ is 90M bps, total transmitted symbols are 128, therefore 256 bits are transmitted.

2. The modem Tx filter is the amplitude equalized square root raised-cosine Nyquist filter with 40% roll-off. Fig.5.11.a is the practical modem Tx filter to be modeled. In this figure, the Nyquist frequency $f_N$ is one
Figure 5.10: QPSK/MSK simulation block diagram for practical application.
half of the symbol rate, 22.5 MHz. In simulation, -40dB is assumed to be the bottom level of the filter.

3. The modem Rx filter to be used as a model is the 4 pole Butterworth filter having the BT =1.0, i.e, the 3 dB cutoff frequency is the same with the Nyquist frequency, and the maximum amplitude level is set to 0 dB. Fig.5.11.b shows the filter characteristic of this model. The simulated filter has exactly the same attenuation of 3 dB at the Nyquist frequency and -14.4 dB at the 1.5 fN.

4. Fig.5.11.c and Fig.5.11.d show the input and output multiplexer filters used as the models, but in simulation the phase delays are excluded because we have assumed complete equalization in Rx earth station. As we can see, these multiplexer filters have wider bandwidth than those of the modem filters if these are simulated with the Butterworth filters. In addition, the attenuation is faster than the Rx filter. Even though there exists a little attenuation in the passband in both cases, it is neglected. The effect of peaking in the middle of the stopband is also neglected because it is not easy to make this effect with the single filter currently built in the simulation program. Therefore the input multiplexer filter is simulated with 9 poles Butterworth filter with BT=

Results of simulation
1.4 and the output multiplexer filter is also simulated with the same type of filter with 10 poles and $BT = 1.3$.

5. Fig.5.12.a and Fig.5.12.b are the characteristics curves of the AM/AM and AM/PM nonlinear functions respectively which are used in the simulation as the model of a memoryless nonlinear TWTA. Because the model used in the previous section for the nonlinear test was based on Saleh [Saleh.8]'s formula for the amplitude and phase model described in Sec.3.4 and thus the simulation program uses this model rather than the quadrature model, it is required to get the amplitude and the phase nonlinear functions first. For this purpose, it is necessary to get the measured data sets from the AM/AM and AM/PM curves depicted in Fig.5.12.a and in Fig.5.12.b. The table in Fig.5.12 shows the measured data taken using the original scales. Then scales are changed from dB to normalized voltage ratio up to the saturation point and degrees are also changed to the radians for convenience. Then these data sets are used to get the coefficients of the nonlinear amplitude function and phase function by using the standard minimum mean-square error curve fitting procedure as stated in the Saleh's paper.[Saleh.8] The solid lines in Fig.5.12.c and Fig.5.12.d show the results of the curve fitting with 14 and 15 measured points data from the Table, while the dotted lines in the

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same figures express the measured data in modified scales. In Fig.5.12, the optimized 4 coefficients of the amplitude and phase model obtained from the curve fitting with 14 points and 15 points of data are also listed in tabular form.

With those above mentioned simulation models and based on the block diagram depicted in Fig.5.10, modem back to back tests and nonlinear tests were performed for QPSK and MSK respectively. Fig.5.13 shows the results of modem back to back test where the output of the modem Tx filter, implemented by the square root raised-cosine Nyquist filter with 40% roll-off and Nyquist frequency 22.5 MHz, is directly applied to the modem Rx filter which is the 4 pole Butterworth filter with BT = 1.0. The result depicted in Fig.5.13 reveals that at the BER of $1\times10^{-4}$, it has the deviation of less than 0.1 dB from the theory in QPSK case and less than 0.2 dB in MSK case. When we compare this result with that of the modem back to back test we got from the previous section, i.e., using the square root raised-cosine Nyquist filter pair as the Tx and the Rx filters, they are almost identical. Therefore it is evident that the combination of square root raised-cosine Nyquist filter with non-Nyquist type filter can replace the square root raised-cosine Nyquist filter pair without causing degradation in the linear channel, at least. As we shall see in the result of nonlinear test later, the performance of the
Figure 5.11: Typical filters' frequency responses used as models for the simulation performed in Sec. 5.4. (a) Modem Tx filter (b) Modem Rx filter (c) Input multiplexer filter (d) Output multiplexer filter. 

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Figure 5.12: The AM/AM and AM/PM characteristics of the TWTA used as the model for practical application (a,b) and the corresponding curves (c,d) made by curve fitting with 14 or 15 points of measured data in Table. The table also shows the optimized 4 coefficients of amplitude/phase model.

Results of simulation
nonlinear test using the former combination is also compatible with the result we got when the square root raised-cosine Nyquist filter pair is used. This fact tells the reason why the relatively simple structured Non-Nyquist type filter such as Butterworth filter or the Chebyshev filter are frequently used in practical systems.

Fig.5.14 through Fig.5.17 show the results of nonlinear test simulated with the elements in the block diagram of Fig.5.10. In Fig.5.14, we can see the results of the input back-off case of QPSK and Fig.5.16 shows the output back-off cases of QPSK. Fig.5.15 and Fig.5.17 show us the input and the output back-off cases for MSK respectively. Finally the required $E_b/N_0$ of QPSK to maintain the specific BER when the input or output-back off are provided are depicted in Fig.5.18.a and Fig.5.18.b respectively. From the results of TWTA input back-off case in QPSK, the additionally required $E_b/N_0$ is about 4 dB at 3 dB input back-off case to get the BER of $1 \times 10^{-4}$. Comparing with the generally accepted degradation of 2 to 3 dB, the result is degraded more than 1 dB and degradation is more severe in the MSK case. In the case of output back-off, QPSK requires an additional 3.3 dB compared with the ideal value and MSK needs more than 5 dB to maintain the same BER for the 1 dB output back-off. From these result, even though MSK and QPSK exhibit almost identical performance in the linear channel, QPSK shows better performance than that of MSK in the nonlinear channel if tight bandwidth is
Figure 5.13: The result of modem back to back test with the practical data described in Sec. 5.4.

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used. This relative better performance of QPSK in comparison with MSK is natural in the sense that MSK requires more bandwidth (1.5 times more) than QPSK. But the degradation we got above will be more severe when there exists a limiter in the satellite transponder because the limiter will distort the signal applied to the TWTA input. In addition, the insertion of the earth station HPA will make the result worse.
Figure 5.14: BER vs $E_b$/No curves of QPSK for various TWTA input back-off.

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Figure 5.15: BER vs $E_b$/No curves of MSK under various TWTA input back-off.

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Figure 5.16: BER vs $E_b/No$ curves of QPSK under various TWTA output back-off.

Results of simulation
Figure 5.17: BER vs E_b/No curves of MSK under various TWTA output back-off.

Results of simulation
Figure 5.18: Required $E_b/N_0$ for fixed BERs for QPSK under various TWTA back-off. (a) input back-off cases. (b) output back-off cases.

Results of simulation
CHAPTER 6 CONCLUSION

The computer simulation of the satellite communication system has been regarded as an economical tool especially when there exist nonlinear devices such as an earth station HPA and a TWTA in the satellite transponder. The simulation program which can simulate the linear and the nonlinear single satellite communication channel was developed and tested. The developed program has the capabilities such as pseudo random sequence generation, QPSK and MSK modulation and demodulation, filtering with the raised-cosine Nyquist filter, Butterworth filter, Chebyshev filter and Lowpass(brick-wall) filter, nonlinear amplification with a TWTA in the amplitude/phase model, and decision making to evaluate the system performance in terms of average BER. With the developed program, some tests were done such as the modem back to back test, and the nonlinear channel test incorporating the TWTA. Finally the same tests were repeated with practical data to produce realistic results from the program.

In the modem back to back test performed to get the reference for further tests, the result for the QPSK case, expressed in average BER deviation from the theory, was less than 0.2 dB and that of MSK was less than 0.4 dB; e.g. at a BER of $1 \times 10^{-6}$, the deviation is 0.2 dB in QPSK and 0.4 dB in MSK. In the nonlinear test, the degradation at the 3dB input
back-off is around 3 to 4 dB in the QPSK case, and 4 to 7 dB in the MSK case for average BER from $1 \times 10^{-4}$ to $1 \times 10^{-8}$. When the extra tests are done with the practical model simulated with some modification, the results of the modem back to back test is compatible with the results of the former test with almost the same amount of deviation from the theoretical values. The results from the the nonlinear test using practical TWTA model, simulated with measured data sets, exhibits a little more degradation, about 2 dB, more than the conventional simulation results listed in papers. Unfortunately there is no proper explanation why this degradation occurs. Because the simulation program developed lacks the method of simulating more than one nonlinearity, and that of simulating the group delay, it can not be a complete simulation at this time. However, under the restricted environmental situation such as limited computer run time, money, and storage area, the results from the simulation are acceptable. Something desirable is the modification of the simulation program into interactive form to facilitate the adaptability of the simulation and easy access.


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