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AN INTEGRATED VOICE-DATA TRANSMISSION SYSTEM FOR PRIVATE MOBILE RADIO.

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DECLARATION

I declare that this thesis has not been,nor is currently being submitted for the award of any other degree or similar qualification.

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ABSTRACT

Title:"An Integrated Voice-Data Transmission System for Private Mobile Radio Networks."

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Private mobile radio services have experienced for a number of years congestion of the available spectrum.Although the release of Band III has contributed significantly to alleviating the spectrum shortage,it will however,not be sufficient to accomodate the projected growth in private mobile radio (PMR).Consequently,new techniques of efficient spectrum utilisation are needed to ensure that future growth and enhancement of PMR services can be met.

The research work undertaken and presented in this thesis has been concerned with doubling the capacity of a mobile radio channel and providing a value added facility by developing an integrated voice and data transmission system.The multiplexed system developed and investigated is based on angle modulation and 12.5 kHz channeling as proposed for Band III mobile systems.

The thesis initially introduces and analyses a number of modulation strategies potentially suitable for the simultaneous transmission of speech and data over a common radio channel.Evaluation of a proposed scheme using a novel CAE simulation tool is described and relevant results presented.Prototype transmitter and receiver equipment developed to implement a speech-data multiplexed radio system is also described.Results of the system performance,obtained under simulated radio propagation conditions,are presented and the implications discussed.In particular,the interaction between speech and data message signals and other problem areas that influence the quality of the service and reliabilty of information transfer are identified and discussed.

Finally,proposals for future development work involving an alternative demultiplexing strategy are also presented.

DECLARATION.

The following related papers have been published prior to the submission of the thesis:-

1.M.O.Al-Nuaimi and M.Charnley,"An integrated voice and data transmission system for private mobile radio", Proceedings of the Online International Conference on Cellular and Mobile Communications, Wembley Conference Centre, 177,Nov 1985.

2.M.Charnley and M.O.Al-Nuaimi,"A functional modelling approach to the design and evaluation of radio systems", Proceedings of the 3rd IERE International Conference on Land Mobile Radio Systems, University of Cambridge, Dec 1985.

3.M.Charnley,"Astec 3:A model design approach to Electronic CAE", IEE Computer Aided Engineering Journal, 192, Dec 1985.

4.M.Charnley,"A cost-effective baseband waveform generator", Electronics and Wireless World, 55, Dec 1986.

5.M.O.Al-Nuaimi and M.Charnley,"Simultaneous voice and data transmission in private mobile radio using a narrowband FM channel",IEE Proceedings,Part F,In Press.

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CONTENTS

APPENDICES.

A: Comparison of FM and PM modulated signals.

B: Spectra of QNBFAM signals and their corresponding basebands.

C: The Distortion Factor DF(rms).

D: Distortion of the SINXOS baseband spectra.

E: Group Delay Distortion.

F: Fast Frequency Shift Keying.(FFSK)

G: A summary of important results derived from the "Scattering" model.

H: Transfer Function Modelling with Astec 3.

I: A Selection of User-defined Astec 3 Functional Models.

J: A Selection of Circuits for the QNBFAM Transmitter.

GLOSSARY

Abbreviations used throughout the main text

ADC	Analogue to Digital converter.
AM	Amplitude Modulation.
ASK	Amplitude Shift Keying.
AGC	Automatic Gain Control.
BER	Bit Error Rate.
BGC	British Gas Corporation.
BW	Bandwidth.
β	Modulation Index (Angle Modulation).
CAE	Computer Aided Engineering.
CTCSS	Continuous Tone Controlled Squelch System.
CDLC	Cellular Data Link Control.
CRAG	Cellular Radio Advisory Group.
D	Deviation Ratio (Angle Modulation).
DSB	Double Sideband.
DFT	Discrete Fourier Transform.
FDM	Frequency Division Multiplexing.
FM	Frequency Modulation
FFSK	Fast Frequency Shift Keying.
FSK	Frequency Shift Keying.
FFT	Fast Fourier Transform.

HDLC	High Level Data Link Control.
Hz	Hertz.
kHz	Kilohertz.
MHz	Megahertz.
GHz	Gigahertz.
IF	Intermediate Frequency.
ILS	Interactive Laboratory System.
ISB	Independent Sideband.
ISDN	Integrated Services Digital Network.
QMDSB	Quadrature Double Sideband Amplitude Modulation.
QNBFAM	Quadrature Narrowband Frequency Modulation with Composite AM.
QNBØM	Quadrature Narrowband Angle Modulation.
LCD	Liquid Crystal Display.
m	Modulation Depth (Amplitude Modulation).
MSK	Minimum Shift Keying.
ms	Millisecond.
μs	Microsecond.
NBFM	Narrowband Frequency Modulation.
NBPM	Narrowband Phase Modulation.
NBØM	Narrowband Angle Modulation.

pdf	Probability Density Function.
PM	Phase Modulation.
PMR	Private Mobile Radio.
PRBS	Pseudo Random Binary Sequence.
PSK	Phase Shift Keying.
PSTN	Public Switched Telephone Network.
RF	Radio Frequency.
s	Second.
S/N	Signal to Noise Power Ratio.
S/Xt	Signal to Crosstalk Power Ratio.
SSB	Single Sideband.
v	Velocity (metres/sec).
VHF	Very High Frequency.
VLSI	Very Large Scale Integration.
WBFM	Wideband Frequency Modulation.
WBPM	Wideband Phase Modulation.
λ	Wavelength (metres).

CHAPTER 1

INTRODUCTION

1.1 Background.

Mobile radio is undoubtedly one of the fastest growing sectors of the telecommunications industry and has a history of continuous growth in every country in the world.Britain in turn is in the middle of a telecommunications revolution, spearheaded by deregulation,the advent of competition and fast developing technology.The introduction of cellular radio telephones is only one manifestation of the change now under way,however,it has captured the imagination and galvanised the whole communications market.British Telecom (BT) with its Cellnet system and Racal with its rival Vodafone operation are locked in a fierce competitive battle which is fulfilling part of the Government aim of introducing competition in the market to chip away at the BT monopoly and provide greater customer choice.Arguably the cellular radio network has broken the logjam which had developed at the civil end of the mobile communications industry and market.

In its report on Mobile Communications in Western Europe 1987,communications analysts CIT Research indicated that the total market is experiencing an explosive growth of over 40% a year.In actual fact,there are now well over 500,000 cellular telephones in Western Europe and CIT believes that the market could double again over the next two years.

However, although Cellular radio has the potential of providing value added services not presently available in Private Mobile Radio (PMR), PMR is nontheless likely to continue to be the main provider of mobile communications to those users who currently employ PMR in their nationwide operations. This is particularly true for users in those areas for which cellular radio is not initially planned.

1.2 Private Mobile Radio and the Merriman Report.

Land mobile radio services have experienced, for a number of years, congestion of the available radio spectrum. However, following the "Merriman Report" (1) (September 1982 and July 1983) which reviewed the use of the radio spectrum in the range 30-960MHz, the portion of the spectrum previously allocated to the 405 line television broadcasting services (Band I and Band III) has, with effect from January 1985, been assigned in the main to land mobile services. The release of Bands I and III has indeed constituted one of the biggest ever additions to land mobile spectrum in the UK with at least 1000 channels being made available in London in Band III alone.

With the current growth of mobile radio services of around 8% annually likely to continue in the near future, it is predicted that the number of mobile radios in use will be more than doubled in the next ten years. Consequently, the relief in the current congestion will be short lived unless the available spectrum is used efficiently and economically.

Another significant aspect encouraged by the Merriman Report is that the release of the new block of spectrum should act as a stimulus to the development of new radio technologies. In response,the Government proposed that part of the Band III spectrum should be used exclusively for the introduction of new and advanced communications systems.

1.3 Benefits of Digital Transmission over PMR radio links.

In recent years PMR users have increasingly come to the realisation that digital transmission of routine information between the Control Centres and mobiles in the field results in considerable saving of the channel air time.

Consider the experience of the British Gas Corporation (BGC) which employ data transmission over their radio networks. British Gas have over 1000 vehicles equipped with mobile radio,therefore it is of considerable importance that they obtain greatest possible efficiency of channel usage. BGC introduced their first digital system,known as Gascord 300,in the last decade.The new prototype version, Gascord 1200,is a 1200 bit per second system which automatically controls access up to 100 mobile units in a single system.Tests have confirmed that when the channel is used for exchanging speech messages the average occupancy time is of the order of 60-90 seconds.In contrast,the Gascord 300 system reduces this to 20 seconds and using the newer 1200 bits per second system an average of only 6 seconds is required.This clearly corresponds to a tenfold saving in air time.Also,automatic requests for revalidation in case of error,are built into the system.Gascord 1200 automatically selects the best transmitter/receiver combination immediately the unit accesses the system.Another useful feature offered by a digital transmission facility is that the data can still be sent when a vehicle is unattended by transmitting the information directly to a printer or message store.This obviates air time wasted by abortive calls to an absent driver.

The overall reduction in air time per message using data means that the channel loading (number of mobiles served by the same channel) can be increased at peak times.This leads as a consequence to higher efficiency and better spectrum utilisation.

A digital transmission facility offers PMR users many other direct benefits which cannot be realised with a voice only mode of operation.For example,the accuracy of certain types of information sent by speech communications can be very doubtful.For instance,there are many applications when accuracy of information transfer is vital such as descriptions of hazardous chemicals sent to emergency services at a road accident.By using forward error correction and automatic re-transmission of erroneous data it is possible to achieve a high confidence in the fidelity of such information transmission.Another important advantage which cannot be ignored is the extra security offered by data transmission. If information is sent in a digital form the traffic is inherently less easy to intercept.In some applications encryption algorithms can readily be employed to scramble the transmitted data to ensure a higher degree of security. Also,if the data can be transmitted directly to a printer or message store in a mobile,this means that the driver does not have to be distracted from controlling the vehicle when in motion thereby improving safety of operation as a consequence.

However,the major benefit of data transmission as perceived by users of PMR systems is the opportunity to introduce value-added services not possible hitherto with conventional systems.The range of value-added services possible in the PMR field will be mainly determined by the solution of the fundamental problems associated with digital transmission in the mobile radio environment and the willingness of both users and manufacturers to agree jointly on a set of requirements and equipment standards.However,one important facility which would appear very attractive to PMR users is the inclusion of a "Prestel" type service in the mobile supported by a local memory capable of storing several pages of information which can be displayed on a LCD or Plasma display unit.This facility is thought to be particularly useful to maintenance staff when servicing equipment and systems in telemetry outstations,pressure reduction sites etc., where schematic diagrams of the systems involved and other relevant data are needed.

Also of particular interest to the field staff is the ability to access,remotely from a mobile,a database which is held in a central computer.For instance,to check the availabilty of spare parts and to reserve them in advance in order to carry out maintenance or repair of customers appliances,e.g., central heating systems.The examples quoted above,illustrate the benefits of digital transmission in the enhancement of services and the reduction of costs by eliminating unecessary travelling time to and from the work base.To summarise,the main benefits of digital transmission are considered to be:-

* Higher operating efficiency

* Better spectrum utilisation

* Accuracy of information transfer

* Security of information transfer

* Introduction of value-added services

1.4 The need for an Integrated Voice and Data Transmission facility.

Digital transmission with all its associated benefits is not considered by PMR users as a facility which should eventually replace speech communications.Indeed,these users see an important and continuing role of the speech link between the control centre and the staff deployed in the field.The speech link is vital in certain emergency situations and there are always those routine situations in which the field staff and the control centre staff need to communicate verbally.Because of these needs the users see the two forms of communications,speech and data,as complimentary and equally beneficial.

It is with all these considerations in mind that the research project,detailed in this thesis,was directed.The central aim of the research being the investigation and development of a suitable multiplexing strategy that is able to provide both data and speech communication between a control centre and mobiles in the field in a manner which enables the use of both these facilities simultaneously. Prior to starting any specific research activity,relevant discussions were held with BGC and other large PMR users regarding the preferred forms of services and systems.During these discussions it became clear that the relevant MPT specifications (2) namely,MPT 1303,MPT 1317 and MPT 1323 must be taken into account in the choice of the final system and the definition of its design parameters.In particular,the simultaneous transmission of both types of information, speech and data,must be achieved within the present frequency and bandwidth allocations of mobile radio channels.

In the system proposed and investigated,both data and speech signals are allowed to occupy the full bandwidth available in a 12.5kHz PMR channel.Allowing both information signals to occupy the full bandwidth available should,in principle,lead to better speech quality and higher data transmission rates than those possible using systems in which it is necessary to restrict the bandwidths of the baseband signals (3,4,5). Furthermore,the integrated speech and data system proposed make it possible for the operator to communicate orally while the data terminal is busy receiving the data message without having to switch to a different radio channel.

1.5 The Research Proposal.

The research proposal had the following initial objectives:

i. To investigate suitable modulation schemes having the capability of integrating the transmission of speech and data over a common private mobile radio channel.

The investigation to be conducted in the light of the MPT specifications relevant to Band III VHF (175-224MHz) and subject to the engineering and commercial constraints involved in equipment design and implementation.The purpose of the investigation to determine the optimal system suitable for simultaneous speech and data transmission.

ii. Based on the design criteria obtained from (i),an important objective was to construct a laboratory data-voice transmission system prototype to assess the system performance under simulated radio propagation conditions in the following modes of operation:

1. Speech transmission only.

2. Data transmission only.

3. Speech and Data transmission simultaneously.

In particular, the interaction between the independent speech and data channels as well as other problem areas likely to influence the quality of the service and the reliability of the information transmitted were to be thoroughly examined. Also, their influence on the design, implementation and performance of the system to be determined and quantified.

The following chapters detail the work undertaken in carrying out the investigation and present results obtained from both computer simulation studies and practical measurements conducted on a laboratory prototype integrated speech-data system. Based on these results a realistic assessment of the proposed multiplexing scheme and evaluation of its performance are provided.

1.6 Outline of the Thesis.

Following the introduction, Chapter 2 reviews relevant research previously undertaken in this area and presents three alternative schemes for consideration (QMDSB, QNBØM and QNBFAM). This chapter also establishes the theoretical basis for the three multiplexing schemes proposed and identifies potential operational problems or limitations that required further investigation. In contrast, Chapter 3 introduces and describes an experimental method,employed during the initial phase of the research,as a means of investigating the orthogonal narrowband angle modulated system referred to as QNBØM.The method,based on a CAE software package known as Astec 3,enabled high level investigations in the form of functional modelling to be successfully undertaken.After presenting relevant fundamental theory,associated with narrowband angle modulation,Chapter 4 describes the evolution of the QNBØM system.Detailed results of the performance of the QNBØM system,obtained by computer simulation,are also presented and their significance discussed.

Chapter 5 describes a laboratory prototype transmitter/receiver system which was designed and constructed to perform the basic multiplexing and demultiplexing processes required in a QNBFAM system implementation.The effects of system imperfections on the level of the interchannel crosstalk and the way this level of crosstalk is likely to influence the quality,intelligibilty and reliability of the speech and data messages conveyed by a practical QNBFAM system are detailed and discussed in Chapter 6.The QNBFAM multiplexing scheme has also been investigated under dynamic channel fading conditions similar to those experienced by a mobile receiver travelling at different speeds.The results of this investigation as well as the actual implementation of the fading channel simulator,are described in Chapter 7.

Chapter 8 discusses possible improvements to the QNBFAM laboratory prototype system before outlining an alternative and potentially better receiver architecture.This alternative architecture represents a radical departure from traditional mobile radio receiver techniques and is recommended as a new direction and initiative for the research effort to be continued in the future.Chapter 9 represents the final chapter of the thesis and reflects on the main successes achieved during the execution of the research project.This chapter closes by providing a summary of the exciting developments that have recently taken place in mobile radio and outlines the potential for mobile communications of the future.

CHAPTER 2

SIMULTANEOUS TRANSMISSION OF SPEECH AND DATA

2.1 Review of Previous Work Undertaken.

The problem of providing a data facility without worsening the spectrum congestion has been investigated by a number of researchers.R.C.French published a number of papers (6,7,8) which analysed the problems involved in digital transmission in the urban environment.Using both prediction and measurements he was able to show that the bit error rate over a mobile radio channel was relatively high,typically 10^{-2} - 10^{-3} in the normal range of received signal levels.This was clearly considered too high for a working system which prompted J.D.Parsons (9) and others to investigate effective techniques of reducing the error rates.One technique proposed consists of using certain types of diversity. Switched diversity was found to be potentially economical and effective in improving the performance of data communications. Another technique investigated employed "predetection combining" which effectively increases the signal to noise ratio prior to detection.In an FM system this technique produces an effect similar to threshold extension,which performs well in the presence of uncorrelated Gaussian noise.

J.R.Edwards (10,11) presented schemes of reliable data transmission and considered the design factors for modem error correcting systems required to overcome the effects of a fading channel.The simulation results obtained by Edwards indicated that an error probability of 10^{-10} may actually be achieved.His work also dealt with interference caused by ignition noise.

The work outlined previously,and that of many others,has demonstrated the feasibilty of digital transmission over a fading radio channel.However,the same amount of dedicated research effort would not appear to have been applied to determine whether a combined speech and medium speed data channel can co-exist within the confines of a single 12.5 kHz channel allocation.

One serious attempt at providing both data and speech within a 12.5 kHz mobile radio channel is due to the mobile radio research group at the University of Bath (12).The system proposed and investigated relies on transmitting data and speech in adjacent bands within the 12.5 kHz bandwidth on the outward path (base to mobile) and using Independent Sideband Transmission (ISB) for the mobile to base direction.The success of such a scheme and other systems,based on a form of Frequency Division Multiplexing,depends largely on whether PMR users and manufacturers are prepared to accept single sideband (SSB) transmission in mobile radio communications.Although a considerable amount of research effort has been directed toward improved implementations of SSB for the mobile radio environment,in particular the work of McGeehan and Bateman (13,14),so far there would still appear to be a reluctance to move away from Narrowband Angle modulation and adopt radio systems based on SSB techniques.

An alternative multiplexing scheme has also been proposed and investigated by a research team from Italy (15).Although specifically designed for air-traffic control applications, nevertheless,the technique proposed is also applicable for other mobile radio systems.The system is based on a form of mixed or hybrid modulation whereby the voice signal modulates the amplitude of the RF carrier,while the data signal modulates the phase of the carrier.The results, obtained through a computer simulation,show that systems using AM-FSK and particularly AM-MSK give acceptable performance in contrast to systems using AM-PSK.Moreover,it was determined that amplitude modulation of the carrier and bandwidth limitations in the system do not cause a significant deterioration in the data channel performance. However,the problem with adopting such a scheme is that it is not compatible with current mobile radio standards and any existing transceiver equipment would have to be modified significantly in order to exploit the multiplexing format proposed.

2.2 Alternative Modulation Schemes for Multiplexing Speech and Data.

The multiplexing schemes reviewed in the following sections emerged as a result of a colleagues experience of mobile radio systems used by the British .Gas Corporation (BGC) and,also subsequently,discussions with BGC and other major PMR users regarding the preferred forms of services and systems.The modulation schemes,to be outlined,are all aimed at leaving the present mobile radio channel of 12.5 kHz unchanged and to use some form of multiplexing to enable data and speech transmissions to take place simultaneously. In these systems both data and speech signals are allowed to occupy the full bandwidth obtained from the 12.5 kHz channel which,in principle,should result in better speech quality and higher baud rates than those resulting from using 6.25 or 5 kHz channel bandwidth proposals. Further,the proposed schemes offer a degree of compatibility with the present 12.5 kHz AM systems, or,narrowband angle modulated systems recommended for Band III.The following sections outline three modulation schemes suitable for multiplexing speech and data,namely :-

(i) Orthogonal Multiplexing using Double Sideband (QMDSB).

(ii) An Orthogonal Angle Modulation System with Synchronous Detection (QNBØM).

(iii) An Orthogonal Angle Modulation System with Conventional Detection (QNBFAM).

2.2.1 Orthogonal Multiplexing using Double Sideband (QMDSB).

This technique,exploited in the PAL Colour TV system for the transmission of R-Y and B-Y colour difference signals,is based on the following theory.

Let s(t) and d(t) represent the baseband speech and data signals respectively, both assumed to occupy the same bandwidth (300-3300 Hz). At the transmitter, two quadrature DSB signals $f_1(t)$ and $f_2(t)$ are generated and then added to a pilot carrier $a \cos\omega_c t$ to form the multiplexed signal f(t) as shown in equations 2.1-2.3.

$$f_1(t) = s(t) \cos\omega_c t \qquad 2.1$$

$$f_2(t) = d(t) \sin\omega_c t \qquad 2.2$$

$$f(t) = a \cos\omega_c t + s(t)\cos\omega_c t + d(t)\sin\omega_c t \qquad 2.3$$

The multiplexed signal f(t) occupies the same bandwidth as that of a conventional amplitude modulated (AM) signal and so can be transmitted over a mobile radio channel of 12.5 kHz.

In principle, a mobile radio with a conventional front-end can be used to receive the multiplexed signal to yield an intermediate frequency (I.F.) signal which is linearly related to f(t). At the I.F. stage, the pilot carrier can be extracted by means of a narrow band-pass filter. The carrier and its quadrature-phase version $a\sin\omega_c t$ are derived and subsequently used to synchronously demodulate the I.F. signal as shown below:-

$$f(t).a\cos\omega_c t = a^2\cos^2\omega_c t + s(t)a \cos^2\omega_c t + \tfrac{1}{2}d(t)a \sin2\omega_c t \qquad 2.4$$

$$f(t).a\sin\omega_c t = \tfrac{1}{2}a^2\sin2\omega_c t + \tfrac{1}{2}s(t)a\sin2\omega_c t + d(t).a\sin^2\omega_c t \qquad 2.5$$

After low-pass filtering and ignoring the d.c. term, equations 2.4 and 2.5 reduce to quantities which are proportional to s(t) and d(t) respectively.

The previous analysis shows that,in principle,QMDSB enables the simultaneous transmission and reception of two distinct information signals.In practice,the detection of the two signals without significant cross-talk interference between the two messages depends to a large extent on ensuring that both amplitude and phase errors in the modulation and demodulation processes are made sufficiently small.It is readily shown that a slight error in the phase of the carriers results in a reduction of the recovered wanted signal amplitudes as well as unwanted crosstalk between the two channels.Similar difficulties arise when the locally derived carrier frequency is in error.In addition,unequal attenuation of the channel will destroy the symmetry of the magnitude spectrum about +/- ω_c and the received signal component x(t) from s(t) will have a form:-

$$x(t) = s_c(t)\cos\omega_c t + s_s(t)\sin\omega_c t \qquad 2.6$$

Instead of:- $s(t)\cos\omega_c t$ 2.6a

Demodulation of this signal will result in $s_c(t)$ in the speech channel and $s_s(t)$ in the data channel.Both $s_c(t)$ & $s_s(t)$ are derived from s(t),hence,the speech channel will receive the desired signal in a distorted form,and the data channel will receive an unwanted interference signal. Similarly,the component y(t) from d(t) will yièld the distorted version of the desired signal in the data channel and an interference signal in the speech channel.

2.2.2. An Orthogonal Angle Modulation System with Synchronous Detection.(QNBØM)

This scheme was devised as a result of discussions with certain PMR users and the proposed introduction of angle modulation for Band III (174-225 MHz).The modulation process may be considered as a narrowband angle modulation (NBØM).The general expression for a NBØM signal modulated with a single tone is given by:-

$$f_c(t) = A \cos\omega_c t - \beta A \sin\omega_m t.\sin\omega_c t \qquad 2.7$$

where ω_c and ω_m are the carrier and modulation frequencies respectively and β is the modulation index (peak phase deviation).

Equation 2.7 contains a DSB component similar to the signals given in equations 2.1 and 2.2 which suggests that an orthogonal multiplexing strategy may be appropriately applied here.

With s(t) and d(t) as previously defined,let:-

$$s_1(t) = \int s(t)dt \quad \text{and} \quad d_1(t) = \int d(t) \qquad 2.8$$

Two NBFM signals $f_{c_1}(t)$ and $f_{c_2}(t)$ can be formed as follows:-

$$f_{c_1}(t) = A\cos\omega_c t - \beta_1\ s_1(t)\sin\omega_c t \qquad 2.9$$

$$f_{c_2}(t) = A\sin\omega_c t + \beta_2\ d_1(t)\cos\omega_c t \qquad 2.10$$

A combined signal f(t) which is the sum of the two signals is given by:-

$$f(t) = \sqrt{2}\ A\sin(\omega_c t + \pi/_4) - \beta_1 s_1(t)\sin\omega_c t + \beta_2 d_1(t)\cos\omega_c t \qquad 2.11$$

f(t) occupies the same bandwidth as that available from a 12.5 kHz channel.At the receiver,the carrier term can be filtered using a narrow band-pass filter.This is used to derive an in-phase and a quadrature-phase carrier output. Using synchronous detection,as explained in section 2.2.1.,it can be seen that s1(t) and d1(t) can be detected separately from which the original signals s(t) and d(t) can subsequently be obtained.

In principle,the above scheme achieves the same objective as the QMDSB system but has the added advantage of being compatible at the transmitting end with the Band III specifications.In practice,equipment design constraints give rise to certain problems in the modulator implementation. These are discussed in more detail in Chapter 4 when results of investigations undertaken using computer simulation techniques are detailed and reviewed.

2.2.3. An Orthogonal Angle Modulated Scheme with Conventional Detection (QNBFAM).

Consider two quadrature carriers whose frequencies are modulated by baseband signals s(t) and d(t). The resultant signal maybe written as:-

$$f(t) = \cos(\omega_c t + \beta_1\phi_1(t)) + \sin(\omega_c t + \beta_2\phi_2(t)) \qquad 2.12$$

For Narrowband FM:-

$$\phi_1(t) = \int s(t)dt \qquad 2.12a$$

$$\phi_2(t) = \int d(t)dt \qquad 2.12b$$

For Narrowband PM:-

$$\phi_1(t) = s(t) \qquad 2.12c$$

$$\phi_2(t) = d(t) \qquad 2.12d$$

The modulation process maybe narrowband or wideband depending on the values of β_1 and β_2 .However,it is reasonable to let $\beta_1 = \beta_2 = \beta$ i.e. by arranging for both modulation processes to have the same peak phase deviation or modulation index.Equation 2.12 can be re-written in an alternative form:-

$$f(t) = 2\cos\{\pi/_4 + \beta/_2[\phi_1(t) - \phi_2(t)]\}\sin\{\omega_c t + \pi/_4 + \beta/_2[\phi_1(t) + \phi_2(t)]\} \qquad 2.13$$

Consider now the envelope term:-

$$E(t) = 2\cos\{\pi/_4 + \beta/_2\ [\phi_1(t) - \phi_2(t)]\} \qquad 2.14$$

This is of the form cos(A+B) where $A = \pi/_4$ & $B = \beta/_2[\phi_1(t) - \phi_2(t)]$
This in turn can be expanded into:-

$$E(t) = 2\cos(\pi/_4).\cos\{\beta/_2[\phi_1(t) - \phi_2(t)]\}$$

$$- 2\sin(\pi/_4)\sin\{\beta/_2[\phi_1(t) - \phi_2(t)]\} \qquad 2.15$$

If the modulation process is restricted to being narrowband then the envelope term reduces to:-

$$E(t) = \sqrt{2}\ \{1 - \beta/_2[\phi_1(t) - \phi_2(t)]\} \qquad 2.16$$

As indicated, the amplitude modulation of the composite signal is related to the "difference" between the two independent baseband signals. Consequently, a conventional AM detector used to detect the envelope of the composite signal f(t) would give rise to an output which is proportional to equation 2.16. If this output is subsequently differentiated the result would be given by equation 2.17.

$$\frac{dE(t)}{dt} = K_1 \frac{d}{dt}[\phi_1(t) - \phi_2(t)] \qquad 2.17$$

And referring to equations 2.12a and 2.12b, for the NBFM case, this would produce a baseband signal corresponding to the difference between the two signals s(t) and d(t) as shown in equation 2.18.

$$\frac{dE(t)}{dt} = K_1[s(t) - d(t)] \qquad 2.18$$

Consider now the phase term of the composite quadrature angle modulated signal f(t) represented by equation 2.13.

$$\theta(t) = \pi/_4 + \beta/_2[\phi_1(t) + \phi_2(t)] \qquad 2.19$$

Equation 2.19 confirms that the resulting angle modulation of the composite signal is in fact related to the "sum" of the baseband signals.

Now, a conventional FM detector gives an output which is proportional to the derivative of $\theta(t)$. Therefore, if an FM detector were driven by the composite signal f(t), and the initial angle modulation was NBFM (Eqns:2.12a and 2.12b), the following baseband signal would be recovered as indicated in equation 2.20.

$$\frac{d\theta(t)}{dt} = K_2[s(t) + d(t)] \qquad 2.20$$

Therefore, the use of conventional AM and FM detectors yield outputs which can be combined to recover the original baseband signals s(t) and d(t). Figure 2.2.3. shows a schematic diagram of the required demultiplexing system.

This system has the principal advantage over the QMDSB and QNBØM systems of requiring only conventional detectors for recovery of the two baseband signals instead of synchronous detection. Clearly, the amplitude scaling factors K1 and K2 need to be set to ensure minimum crosstalk interference between the two halves of the system. This requirement is given detailed consideration in Chapters 5 and 6.

2.3 Effects of Frequency Multiplication.

At the transmitter,conventional NBØM mobile equipment design relies on the use of frequency multiplication (factor of 16 typically) to arrive at the eventual VHF carrier frequency and also the maximum frequency deviation value of 2.5 kHz which is permitted by the adopted standard for 12.5 kHz channeling.However,the current practise for achieving such frequency multiplication relies on a non-linear process which inevitably affects the orthogonality of a quadrature multiplexed narrowband angle modulated signal as well as seriously distorting the envelope information of the composite signal.Both these effects have significant consequences for the multiplexing schemes discussed in sections 2.2.2. and 2.2.3. and will be dealt with in due course,in relevant chapters.However,as far as the QNBFAM system is concerned the undesirable distortion of the envelope can be circumvented.In order to avoid the envelope distortion,without precluding the use of frequency multiplication,the system represented by Figure 2.3 can be employed.

In this system the envelope information is removed from the multiplexed signal at the I.F. stage prior to frequency multiplication.The envelope information,which is essential for the demultiplexing process,is superimposed back on the amplitude of the VHF carrier after frequency multiplication. In turn,the frequency multiplication not only increases the frequency deviation of the original multiplexed signal but converts the I.F. signal to the operational carrier frequency ready for transmission at VHF.The system outlined in Figure 2.3 also has the additional advantage that the modulation depth of the envelope can be controlled independently of the frequency deviation of the composite signal.

2.4 Summary.

The QMDSB and QNBØM schemes,previously reviewed (2.2.2 and 2.2.3),both require synchronous detection to recover the independent baseband signals from the multiplexed RF signal.In order to keep the crosstalk interference between the two separate information bearing channels to a minimum, precautions must be taken to ensure that amplitude and phase errors are sufficiently small at the receiver. Consequently, these systems become somewhat less attractive to the user unless some means can be found to overcome the effects of the crosstalk interference on the reliabilty of transmitted information,such as better error control strategies for data signals.Although the QMDSB system has the advantage of being relatively straightforward in its implementation,it does lack compatibility with Band III recommendations.On the other hand,QNBØM is an angle modulated system and therefore conforms to the recommendations.However,the system implementation suffers from the problems highlighted in sections 2.2.2. and 2.4 and dealt with in more detail in Chapter 4.

The QNBFAM system evolved as a result of preliminary investigations conducted to determine the performance of an operational QNBØM system. The QNBFAM system offers compatibility as well as simplicity of information recovery at the receiver. The implementation of frequency modulation to the required frequency deviation can also be achieved by standard techniques commonly used in modern mobile transceiver equipment. At the receiver conventional AM and FM detectors can be utilised in the demultiplexing process to recover the independent baseband signals. This feature offers the opportunity of using existing mobile receiver equipment, with suitable modifications, to enable the correct reception of such a multiplexed signalling scheme.

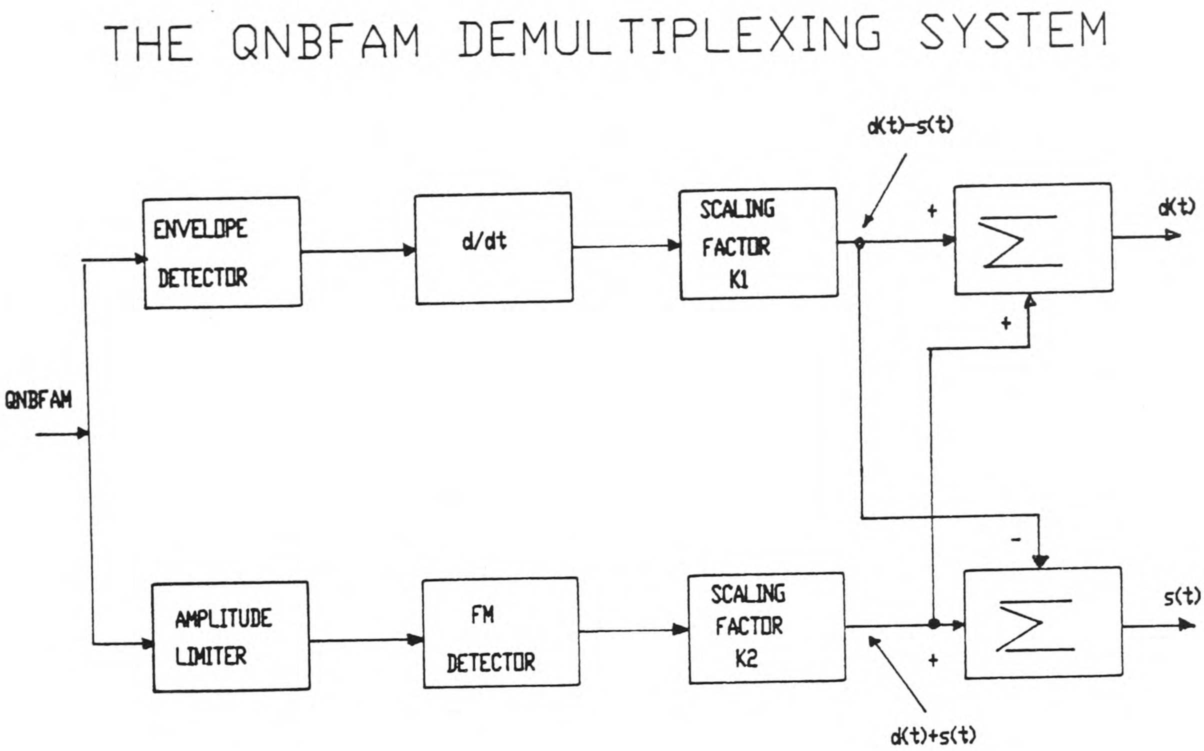


Figure 2.2.3 : The QNBFAM Demultiplexer

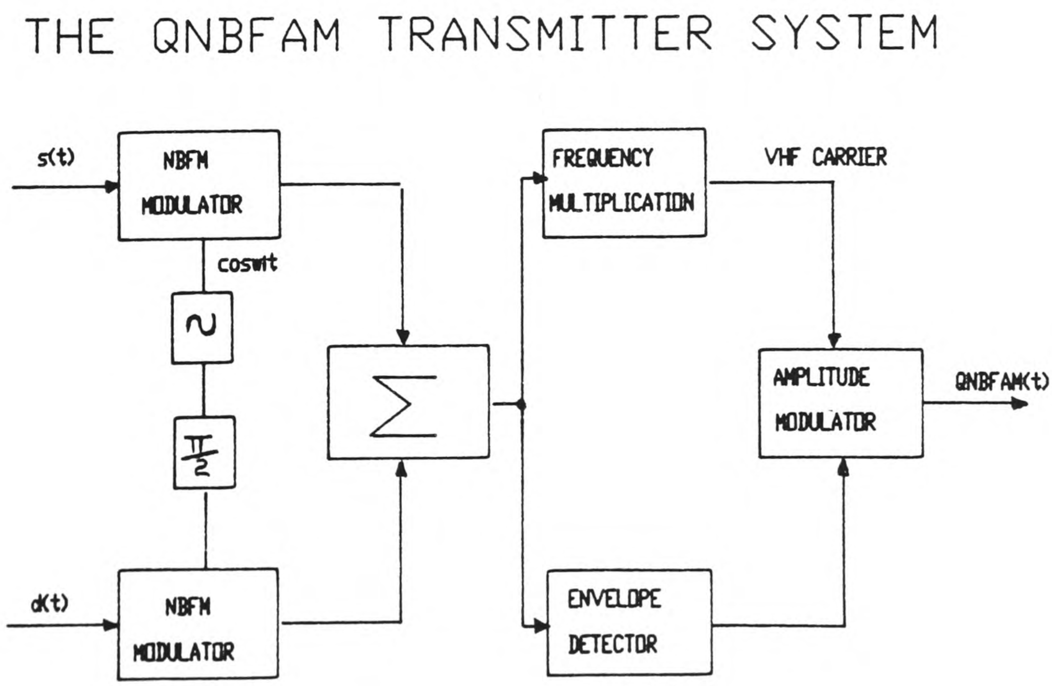


Figure 2.3 : The QNBFAM Transmitter

CHAPTER 3

A FUNCTIONAL MODELLING APPROACH USING ASTEC 3

3.1 Introduction.

The first scheme considered worthy of further investigation was the QNBØM orthogonal multiplexing strategy based on narrowband angle modulation.However, it was realised from the outset that this approach was not without its inherent problems.The problems,summarised below,clearly required further investigation in order to determine whether or not they could render the system unsuitable for use in a mobile radio application.

As indicated in Chapter 2,and given further detailed consideration in Chapter 4,the major limitations identified were associated with the frequency multiplication process employed in practical narrowband angle modulated radio systems.Namely,the undesirable effects of frequency multiplication on the orthogonality of the quadrature multiplexed signal and the resultant envelope of the composite waveform.The extent of the distortion of the recovered baseband signal and the magnitude of the interchannel crosstalk,produced as a result of a demultiplexing strategy which relies on synchronous detection,also needed to be determined.

Although a purely theoretical analysis of the system could have provided certain pertinent answers to the problems identified it was eventually discarded in favour of a more appropriate engineering approach.The investigative method adopted was based on a high level design strategy,employing functional modelling,which could also be used to introduce practical implementation problems such as component tolerance and other relevant imperfections such as drift.The functional modelling technique was achieved by employing a CAE software package known as Astec 3.In addition to the usual features expected of an electronic circuit simulator Astec 3 provided a number of facilities which were harnessed and refined to provide a very efficient and elegant solution to the investigations to be undertaken.Chapter 4 will demonstrate the suitabilty and convenience of the functional modelling approach and will illustrate how a computer terminal can be converted into an investigative tool capable of presenting detailed results in the time and frequency domains.

3.2 Astec 3 : A Strategy for System Modelling and Evaluation.

Astec 3 is a general purpose analogue circuit and system simulation package capable of performing DC,AC and transient analysis.The software was originally developed by the French Atomic Energy Authority and has been available in the UK since 1983.

The main attributes of Astec 3 include the following features:-

(i) System modelling at a functional description level as well as the physical component level.

(ii) The ability to create a library of user defined models.

(iii) Powerful model nesting facilities.

(iv) Excellent time domain graphical output.

(v) Comprehensive system library of electronic devices.

The input language of the Astec 3 program consists of a collection of straightforward commands and sub-commands.They describe the elements of the circuit to be simulated,the topological aspects of the circuit and sub-circuits,as well as the element values which in turn may be any specified function of other network variables.They also allow the user to specify the type of simulation to be executed and the choice of results to be derived as well as defining the format in which the output results are to be presented.

Unlike other simulation languages,Astec 3 permits the modelling of electronic systems on a function by function basis which encourages the engineer to break down complex designs into smaller more manageable and simpler tasks.It is this approach which permits an initial schematic block diagram specification to be transformed into a representation suitable for dynamic testing under computer simulation.Also,many electronic systems such as filters are defined in terms of an "s" domain relationship between output and input variables.Such transfer functions can be translated into sets of first order differential equations which are then in a form suitable for direct simulation under Astec 3 (Appendix H).

3.3 Circuit and System Modelling with Astec 3.

Astec 3 allows the description of sub-circuits or processing modules,called models,which can be used individually or nested within other more complex models many times,without the need to write out the sub-circuit description each time it is employed.User defined models can also be permanently stored on whatever medium the user so desires in the form of a personalised library.The models stored in the users library can be used in any future circuit in the same way as if it had been defined as a model in the description sequence of a particular Astec 3 run.

The !MODEL sequence is used to describe a sub-circuit separately from the main circuit description.This must be done before any use of the model.

Syntax:

!MODEL model-name(node-node...node):

Sub-circuit description.

Model-name...A name supplied by the user,consisting of one to six alphanumerical characters.This name identifies the model and must be unique within the $DESC (description) command.This name is used to refer to the model at its time of use.

node-node...Set of node names,separated by a delimiter (- or ,) which identify the external connections of the model.

Sub-circuit description...List of instructions describing the sub-circuit.

The model description can call on all of the following basic instructions of the description language.

- Branch definitions
- Local parameter definitions
- Differential equation definitions
- Electrical variables
- Definitions of tables and distributions
- Conditionals
- Models and types already defined

Elements in a model need not be allocated a value,such elements are called "formal arguments".These formal arguments must be supplied with numerical constant values when the model is actually used.Of course,different numerical constants can be supplied for each different usage of the model.All the elements (nodes,branches,parameters etc) defined in such a !MODEL sequence,are local to it and as such will not be confused with any element of the same name defined in another sequence.

3.3.1. Examples of the !MODEL Sequence.

```
!MODEL MIXER(IN1-IN2-OUT-COM):
PKMIX;
UIN1(IN1-COM);
UIN2(IN2-COM);
EOUT(OUT-COM)PKMIX*UIN1*UIN2;
```

With reference to Figure 3.1,the basic function of this model is to mix two input signals together by performing the mathematical operation of multiplication.PKMIX is a formal argument to be given a numerical value when the model is used and effectively scales the amplitude of the output signal resulting from the multiplication of the two input signals.

The prefix U employed before the IN1 and IN2 sequences define the input voltage generators as external to the model,whereas,the prefix E before the OUT sequence identifies the output voltage generator as internal to the model.

The external nodes of a model provide communication between the model and the circuit and/or model in which it is to exist.When the model is used,there must be the same number of node names supplied,but these node names now refer to nodes in the circuit or outer model.An outer node name may be thought of as being electrically connected to the model node name in the same position.Consequently,the order of the node names is significant.

If the ground-node is needed in the model it must be passed through to the model by means of a connection to an external node of the model.(Use of a "common" terminal is recommended).Any previously defined model,possibly retrieved from a library,used in the circuit description sequence can be used in a !MODEL or !CIRCUIT sequence as many times as is necessary.When the model is used,it is implanted into the circuit by connecting the model connecting nodes to specified nodes in the circuit.Consequently,a copy of the model becomes part of the circuit or model in which it is implanted.The sub-circuit represented by the model within the main circuit is actually called a Macro.A Macro in this context therefore becomes a circuit element,but with two or more nodes associated with it.

Another example of a model sequence is included to illustrate how a user may define a model in terms of physical electronic components.This is an alternative to employing a mathematical function to establish the relationship between the input and output node voltages,as employed in the mixer configuration previously described.The model sequence listed below corresponds to a simple "LC" low pass filter (Figure 3.2).The relevant values of each component are required to be supplied as formal arguments when the model is used as part of a more extensive circuit or system configuration.

```
!MODEL LCLPF(IN-OUT-COM):
PC1;
PC2;
PL1;
UIN(IN-COM);
L1(IN-OUT)PL1;
C1(IN-COM)PC1;
C2(OUT-COM)PC2;
```

This model can now be inserted into a cicuit description sequence in the following manner:-

```
MFILT(FILTIN-FILTOP-GND)LCLPF:
PC1=0.067MC,
PC2=0.067MC,
PL1=48ML;
```

Specific values for the filter components are allocated to the parameters PC1,PC2 and PL1 which make up the formal argument list for the model.(Further examples of more sophisticated user defined models are given in Appendix I.)

3.4 System Simulation Stages in Astec 3.

The simulation of a system via Astec 3 proceeds in three well defined stages:-

(a) Problem description.

(b) The simulation run itself.

(c) The presentation of results.

Each of these phases may proceed independently and maybe repeated as required.For instance,after problem description and simulation the results obtained may be presented in a number of different ways,graphical or tabular,without the need to repeat the description or simulation phases.This saves considerably on computing resources especially when Monte-Carlo simulations involving statistical analysis of the effects of component spreads are concerned.It is beyond the scope of this discussion to present a description of the Astec 3 system,details of which are available elsewhere (16,17).However,as a simple example consider the modelling of the orthogonal narrow band angle modulated signalling system as represented by Figure 3.3.This can be readily converted into a form suitable for testing using Astec 3.The waveforms obtained during a 2 milli-second transient simulation sequence at important nodes of the modulator system are shown in Figures 3.4a and 3.4b,which clearly indicate how the graphical output capabilities of Astec 3 can transform a VDU screen into an effective multi-channel cathode ray oscilloscope.The complete Astec 3 data file for modelling the QNBØM multiplexing system is listed in Figure 3.5.

The previous example demonstrates that an Astec 3 simulation need only consist of three distinct phases following relevant command keywords.

3.4.1. Phase 1 : The $DESC Command.

This command is used to identify all models to be employed in the circuit or system description,as well as describing all the necessary interconnections,branches and local parameters involved in defining the network to be investigated.Also required during this sequence is a statement to specify all dynamic values to be made available for the signal data output display procedure.This is achieved by using the !OUTPUT: keyword which follows the

!CIRCUIT: description sequence.Only the elements following the declaration will be accessible by the $EDIT command.As indicated,user defined models are recalled from the users personal library by simply declaring their names following the !LIB: sequence keyword.As shown,the model of the QNBØM system contains macros referring to other models,which in turn could if required contain a macro referring to yet another model,and so on.This is known as model "nesting" and there is no restriction on the complexity of the nesting except that the nesting must not be recursive.

3.4.2. Phase 2 : The $TRAN Command.

The $TRAN command performs a transient simulation of the network described during Phase 1.The circuit description defined during the $DESC command is converted into a set of first order differential equations.These are solved by numerical integration from a set of initial conditions defined by the !INIT sequence.The !PARAM: sequence permits the user to specify the value of various program environmental parameters such as algorithmic method,convergence criteria,maximum number of iterations,initial dc steady state parameters,number of cases for statistical simulations,number precision of quantities etc.All of these parameters,with the exception of of TMAX which determines the duration of the transient simulation,possess default values which are chosen as a compromise between accuracy and calculation time for the type of circuits encountered in electronic engineering.In general,the user need only become involved in using these parameters when numerical difficulties are encountered.Although the numerical integration algorithm calculates results at unequal intervals of time,results can be stored on file at equal intervals by allocating a suitable value to the HOUT parameter.The simulation phase is terminated by the !EXEC sequence keyword.

3.4.3. Phase 3 : The $EDIT Command.

The final phase follows the $EDIT command which allows the user to gain access to the results of the previous simulation(s) that were stored on file during execution of the earlier phases.These results contain the values calculated for the elements and variables mentioned in the !OUTPUT: sequence of the $DESC command.The $EDIT command allows the declared entities to be printed out in tabular or graphical form at a standard or graphics terminal.

An alternative method of handling output data is to use the !PRINT sequence which enables simulation results to be presented in the form of a tabular representation.This type of representation is very suitable for performing further analysis on simulation data such as obtaining frequency domain information by performing a Discrete Fourier Transform on relevant sets of time domain responses.

To assist the user in performing such analysis it is possible to transfer pertinent simulation results to a separate data file for further access rather than to the VDU or graphics display.This is readily achieved by executing an

$EDIT sequence after running the corresponding $DESC and $TRAN phases of the simulation. An $EDIT sequence suitable for storing 1024 uniformly spaced samples of the resultant modulated output signal, over the time period one to two milli-seconds of the simulation sequence, follows:-

```
$EDIT
!PRINT(TITLE=QNBØM MODULATED WAVEFORM DATA),
VERSUS &T(RENAME=TIME,LINSCALE=1ML/2ML,NX=1024):
VNQNBOUT(RENAME=OUTPUT);
!EXEC
$END
```

Table 3.1 represents the format of the actual data transferred to the output data file after executing the $EDIT sequence, previously listed, following a transient simulation of the multiplexing scheme represented by Figure 3.3. Obviously, to perform any particular analysis on such output data the file must be accessed by a separate program, specifically written to perform the required computation. The instruction sequence listed in Figure 3.6 corresponds to part of a Fortran program which prompts the user to enter the name of the data file to be interrogated and to identify the name of the specific data variable to be accessed. The called sub-routine scans the identified file to extract the data corresponding to the specified variable and then transfers the relevant data into a "real" array defined as DATA(I) while completely ignoring all other entries in the output data file.

3.5 Final Remarks.

Experience has shown that Astec 3 is well suited to the type of investigations involved in assessing modulation methods considered for this specific application.In particular,the ability to translate a schematic block diagram which represents a functional description of a modulation scheme into a form suitable for dynamic testing has been a very attractive and useful feature.The power to model and simulate theoretical or idealised systems,where appropriate,as well as being able to introduce practical implementation problems,such as component tolerance and drift,have confirmed its suitability as an investigative strategy,as well as being an effective design aid.

The decision to adopt Astec 3 as an experimental method has been vindicated on a number of occasions.The ease with which modifications can be effected as well as the efficiency with which results are derived have proved very beneficial to the research effort and progress that has been achieved.The ability of Astec 3 to mix component and functional models means that component level sub-circuit simulations maybe substituted,one by one,for the functional models to verify the correctness of the circuit designs prior to building a hardware prototype.

Astec 3 was also used successfully in the design of a complex baseband signal generator which was needed during the experimental investigations on the QNBFAM multiplexing system laboratory prototype (Chapters 5 & 6).The generator, described in more detail in (18), was employed during the transmitter/receiver alignment procedure as well as the crosstalk and distortion investigations undertaken on the laboratory prototype.

The next chapter details the experimental work undertaken using Astec 3 to investigate the QNBØM orthogonal narrowband angle modulated multiplexed signalling systems.

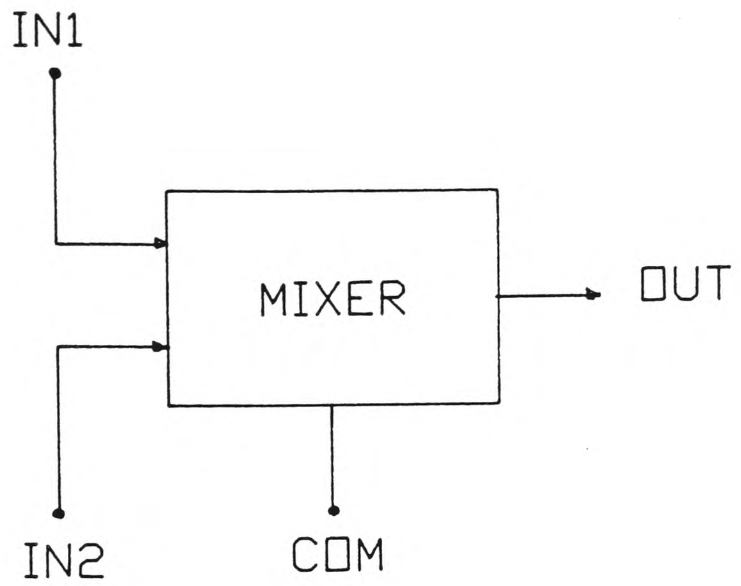


Figure 3.1 : Astec 3 Model of a Mixer

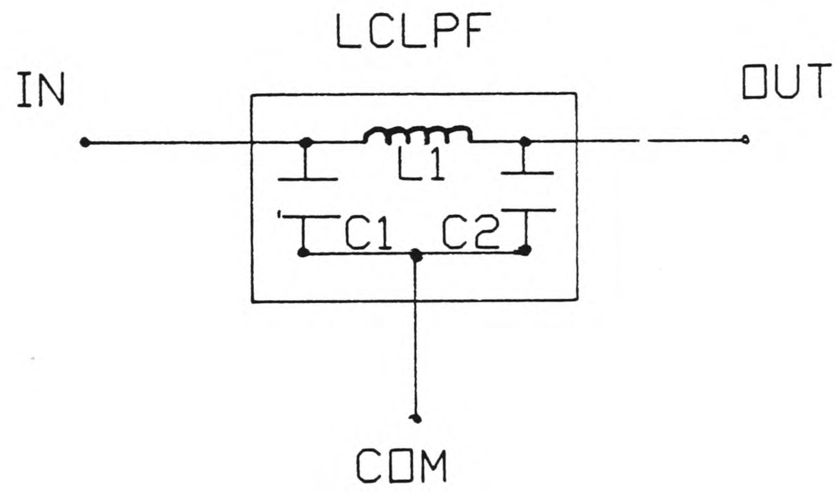


Figure 3.2 : Astec 3 Model of a Low Pass Filter

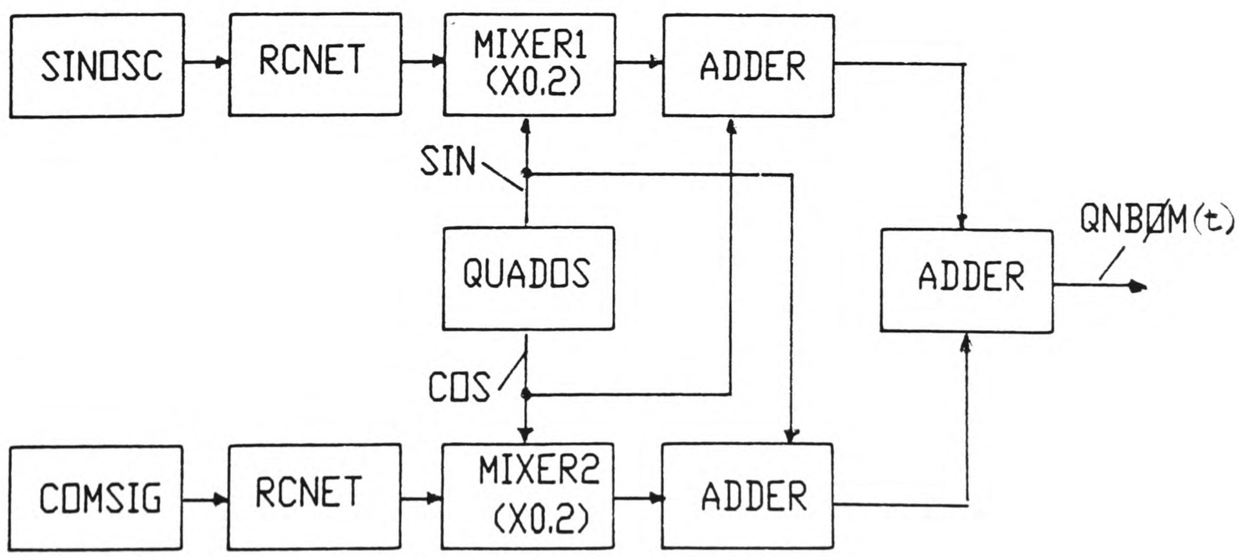
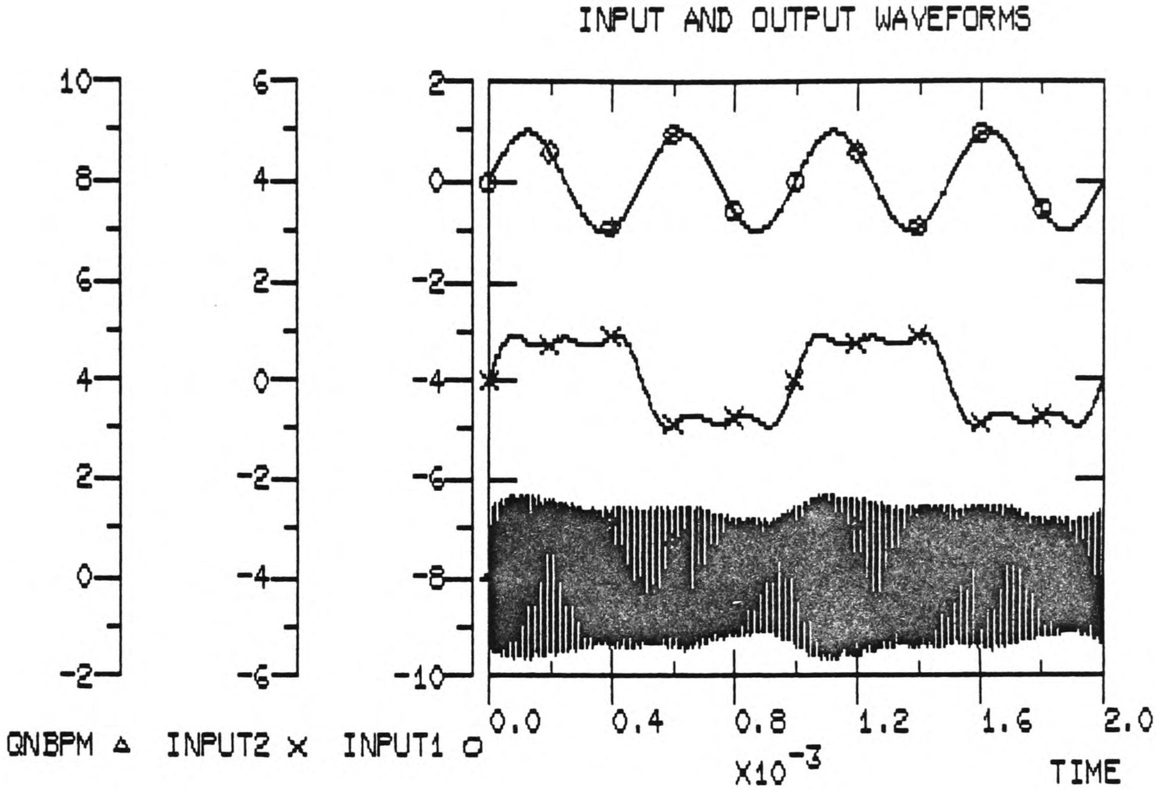


Figure 3.3: Astec 3 Model of a QNBØM Modulator (Low Deviation)

(a)



(b)

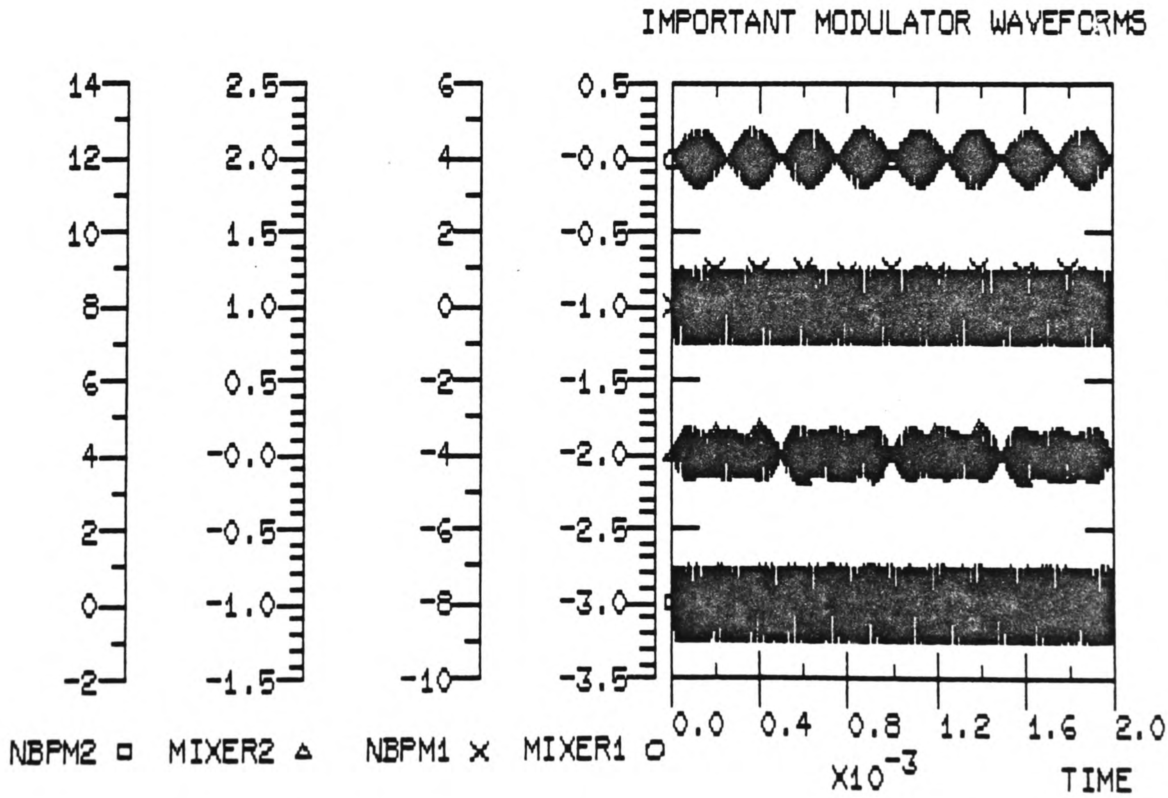


Figure 3.4 : Modulator waveforms produced by Astec 3

```
$DESC

!LIB:QUADOS;MIXER;ADDER;SINOSC;RCNET;COMSIG;

!MODEL QNBMOD(IN1-IN2-OUT-COM):
MIXER1 (IN1-SINOUT-MIX1OP-COM)MIXER:PKMIX=0.2;
MIXER2 (IN2-COSOUT-MIX2OP-COM)MIXER:PKMIX=0.2;
MQUCARR (SINOUT-COSOUT-COM)QUADOS:PA=1,PFREQ=100K,PERROR=0
MADD1 (MIX1OP-COSOUT-NBMOD1-COM)ADDER;
MADD2 (MIX2OP-SINOUT-NBMOD2-COM)ADDER;
MADD3 (NBMOD1-NBMOD2-OUT-COM)ADDER;

!CIRCUIT(GND):
MSIG1 (SIG1-GND)SINOSC:PA=1,PFREQ=2K;
MSIG2 (SIG2-GND)COMSIG:PA1=1,PFUND=1K;
MCR1 (SIG1-IN1-GND)RCNET:PCAP=0.1MC,PRES=100K;
MCR2 (SIG2-IN2-GND)RCNET:PCAP=0.1MC,PRES=100K;
MOD (IN1-IN2-QNBOUT-GND)QNBMOD;

!OUTPUT:VNSIG1;VNSIG2;VNQNBOUT;
        MOD.VNMIX1OP;MOD.VNMIX2OP;MOD.VNNBMOD1;MOD.VNNBMOD2

!EXEC

$TRAN
!INIT RESET
!PARAM:TMAX=2ML;HOUT=0.5MC;
!EXEC


$EDITTK
!PPLOT(TITLE=INPUT AND OUTPUT WAVEFORMS),
VERSUS &T(RENAME=TIME):
VNSIG1(RENAME=INPUT1,LINSCALE=-10/2);
VNSIG2(RENAME=INPUT2,LINSCALE=-6/6);
VNQNBOUT(RENAME=QNBPM,LINSCALE=-2/10);

!PPLOT(TITLE=IMPORTANT MODULATOR WAVEFORMS),
VERSUS &T(RENAME=TIME):
MOD.VNMIX1OP(RENAME=MIXER1,LINSCALE=-3.5/0.5);
MOD.VNNBMOD1(RENAME=NBPM1,LINSCALE=-10/6);
MOD.VNMIX2OP(RENAME=MIXER2,LINSCALE=-1.5/2.5);
MOD.VNNBMOD2(RENAME=NBPM2,LINSCALE=-2/14);

!EXEC

$END
```

Figure 3.5 : Astec 3 Data File Listing

```
PRINT  - 1

I-----------------------------------------------------------------------I
I QUADRATURE ANGLE MODULATED WAVEFORM DATA                              I
I-----------------------------------------------------------------------I
I                                                                       I
I    SYMBOL      MINIMUM      MAXIMUM      ---- VARIABLE NAME ----      I
I                                                                       I
I      A        -1.64E+00    1.65E+00      OUTPUT                       I
I                                                                       I
I-----------------------------------------------------------------------I

TIME          1.000E-03   1.001E-03   1.002E-03   1.003E-03   1.004E-03
OUTPUT        9.991E-01   1.374E+00   1.220E+00   6.429E-01  -2.286E-01

TIME          1.005E-03   1.006E-03   1.007E-03   1.008E-03   1.009E-03
OUTPUT       -1.001E+00  -1.384E+00  -1.154E+00  -6.488E-01   2.353E-01

TIME          1.010E-03   1.011E-03   1.012E-03   1.013E-03   1.014E-03
OUTPUT        1.034E+00   1.383E+00   1.296E+00   6.555E-01  -2.429E-01

TIME          1.015E-03   1.016E-03   1.017E-03   1.018E-03   1.019E-03
OUTPUT       -1.044E+00  -1.400E+00  -1.314E+00  -6.617E-01   2.462E-01
```

Table 3.1

```
      WRITE(5,1000)
1000  FORMAT(1X, 'PLEASE INPUT FILENAME:'$)
      READ(5,1001)FILENAME
1001  FORMAT(A12)
      OPEN(UNIT=1,FILE=FILENAME,FORM='FORMATTED',STATUS='OLD',
       READONLY)
      WRITE(5,1002)
1002  FORMAT(1X,'PLEASE INPUT DATA REQUIRED:'$)
      READ(5,1003)SELECTED__DATA
1003  FORMAT(A12)
      CALL READ__INFO__FROM__FILE(SELECTED__DATA)
      CLOSE(1)


C     ...........................................................................
      SUBROUTINE READ__INFO__FROM__FILE(SELECTED__DATA)
C     ...........................................................................

      CHARACTER*12 SIGNAL__NAME
      CHARACTER*12 SELECTED__DATA
      REAL DATA__SAMPLE(5)
      REAL DATA(1024)
      INTEGER*4 DATA__ARRAY__POS
      COMMON/ARRAY__INFO/DATA,DATA__ARRAY__POS
      DATA__ARRAY__POS=1
10    READ(1,1000,ERR=10,END=999)SIGNAL__NAME,(DATA__SAMPLE(I),I=1,5)
      IF (SIGNAL__NAME.EQ.SELECTED__DATA) THEN
      DO I=DATA__ARRAY__POS,DATA__ARRAY__POS+4
      DATA(I)=DATA__SAMPLE(I-DATA__ARRAY POS+1)
      ENDDO
      DATA__ARRAY__POS=DATA__ARRAY__POS+5
      ENDIF
      GO TO 10
999   RETURN
1000  FORMAT(A12,4(F10.0,2X),F10.0)
      END
```

Figure 3.6 : Transfer of Astec3 output data - Fortran Listing

CHAPTER 4

THE QNBØM MULTIPLEXING SCHEME

4.1 Basic Theory of Narrowband Angle Modulation.

Angle modulation is in general a non-linear process and the spectral composition of an angle modulated waveform is not related in any simple way to the message spectrum.Furthermore,superposition does not apply and the bandwidth of an angle modulated signal can be much greater than the baseband message spectrum.Also,with angle modulated signals both Phase (PM) and Frequency Modulation (FM) are produced and consequently PM and FM are closely related.(The relationship between PM and FM is given greater consideration in Appendix A.)However,for convenience,the following treatment is restricted to considering Frequency Modulated signals.

FM signals can be classified according to their value of Deviation Ratio "D",where:-

$$D = \frac{\text{Maximum Frequency Deviation}}{\text{Maximum Modulating Frequency}}$$

If $D < 0.5$,the FM signal is considered to be narrowband.(NBFM.)

If $D>1$,the FM signal is considered to be wideband.(WBFM.)

4.1.1 Similarity of NBFM to DSB AM Signals.

The basic expression for an FM signal is given in Equation 4.1:-

$$f_c(t) = A_c \cos(\omega_c t + \phi(t)) \qquad 4.1$$

With $$\phi(t) = \beta \int_o^t x(\tau)d\tau \qquad 4.2$$

Where β corresponds to the Peak Phase Deviation or Modulation Index and $x(\tau)$ the baseband message signal.

For NBFM, the maximum value of $|\phi(t)|$ is much less than unity (an alternative definition for NBFM) and therefore, when the message signal is restricted to a single tone the NBFM signal can be represented by:-

$$f_c(t) = A_c \cos\omega_c t - \beta A_c \sin\omega_m t.\sin\omega_c t \qquad 4.3$$

(As given in Equation 2.7 in Chapter 2.)

By inspection, Equation 4.3 indicates that a NBFM signal contains a carrier component and a quadrature carrier linearly modulated by the baseband signal. This is conveniently represented in the phasor diagram of Figure 4.1.1. The addition of the modulation in quadrature with the carrier in NBFM, in contrast to that which is in phase in the case of DSB AM, is emphasised when the equations of NBFM and AM are compared using the exponential form.

$$f_c(t)_{NBFM} = A_c e^{j\omega_c t} [1 + j\beta \sin\omega_m t] \qquad 4.4$$

$$f_c(t)_{AM} = A_c e^{j\omega_c t} [1 + m \cos\omega_m t] \qquad 4.5$$

Taking the real parts of the exponential equations:-

$$\text{Re}\,\{f_c(t)_{NBFM}\} = Ac\,[\cos\omega_c t - \beta \sin\omega_m t \sin\omega_c t] \qquad 4.6$$

$$\text{Re}\,\{f_c(t)_{AM}\} = Ac\,[\cos\omega_c t + m \cos\omega_m t \cos\omega_c t] \qquad 4.7$$

4.1.2. Limits on β for True NBFM.

Considering the NBFM phasor diagram of Figure 4.1.1.,the phase deviation introduced is given by:-

$$\theta(t) = \text{Tan}^{-1} [\beta \sin\omega_m t] \qquad 4.8$$

Now,for a true FM signal,the instantaneous frequency deviation from the carrier should be equal to:-

$$\Delta\omega \cos\omega_m t = \beta\omega_m \cos\omega_m t \qquad 4.9$$

However,from Equation 4.8 the instantaneous frequency deviation is given by:-

$$\Delta\omega = \frac{d\theta(t)}{dt} = \frac{\beta\omega_m \cos\omega_m t}{1 + \beta^2 \sin^2\omega_m t} \qquad 4.10$$

This expression approximates to Equation 4.9 providing:-

$$\beta^2 \sin^2\omega_m t \ll 1 \qquad 4.11$$

Also, the amplitude of the resultant phasor should be a constant for an FM signal. Again referring to the phasor diagram, the resultant amplitude is given by:-

$$|f_c(t)| = Ac\sqrt{1 + \beta^2\sin^2\omega_m t} \qquad 4.12$$

However, Equation 4.12 also approximates to a constant value providing the condition identified in Equation 4.11 is met.

Since $\sin^2\omega_m t$ is less than or equal to unity, the approximations referred to previously are therefore valid if $\beta^2 \ll 1$. Choosing $\beta^2 < 0.1$, then $\beta < 0.316$ is a reasonable bound for a NBFM approximation.

4.2 Generation of NBØM Signals and Evolution of the QNBØM System.

Figure 4.2.1 represents schematically how AM, NBPM and NBFM signals can be generated by using balanced modulators. The only difference between the NBPM system and that of the NBFM system being the presence of an integrator prior to the multiplication process provided by the balanced modulator (Appendix A). As discussed in Chapter 2, and given further consideration in 4.1, the structure of NBØM signals contain sidefrequency components similar to DSB AM signals which suggests that an orthogonal or quadrature scheme based on

NBØM signals maybe appropriate for multiplexing speech and data baseband signals.Figure 4.2 represents a block diagram of such a multiplexing configuration which as can be seen involves two NBØM modulators but with the respective carrier components in quadrature with each other.

As discussed in 4.1.2.,in order to ensure a linear relationship between the angle modulation introduced,(both frequency and phase deviation),and the input signal amplitudes presented at both channel inputs,the initial value of modulation index chosen for the system has to be restricted to a value less than 0.3.However,if a 1 kHz component is considered,the maximum frequency deviation obtainable from such an arrangement is only of the order of 300 Hz which is significantly less than the 2.5 kHz deviation permitted by the existing MPT transmission standard.This is also true of modulators used in conventional angle modulated mobile radio equipment.The standard method employed in practice to increase the value of frequency deviation obtainable to 2.5 kHz is that of frequency multiplication,the principle of which is represented in Figure 4.3.

Although this technique is readily implemented there are some serious consequences that have to be taken into account when considering the orthogonal multiplexing strategy based on NBØM.Firstly,the process of frequency multiplication yields extra pairs of significant side frequency components as the effective value of the modulation index is increased.

Consider a narrow band angle modulated signal applied to a "squaring" circuit:-

$$Vo(t) = [\cos(\omega_c t + \beta m(t))]^2 \qquad 4.13$$

After suitable band pass filtering:-

$$Vo(t)' = K\{\tfrac{1}{2}\cos 2\omega_c t - \beta m(t)\sin(2\omega_c t) - \tfrac{1}{2}\beta^2 m^2(t)\cos 2\omega_c t\} \qquad 4.14$$

Taking each term of Equation 4.14 individually and neglecting any amplitude scaling or phase shift introduced by filtering:-

(i) $\frac{1}{2}\cos 2\omega_c t$

This represents the carrier component now at twice its original frequency.

(ii) $\beta m(t)\sin(2\omega_c t)$

Corresponds to the first set of sidefrequency components which are still in quadrature with the carrier component.

(iii) $\frac{1}{2}\beta^2 m^2(t) \cos(2\omega_c t)$

Yields extra sidefrequency components.Their number and distribution depending on the complexity of the modulating signal m(t).

The effect on the spectra of a narrowband angle modulated signal with tonal modulation and an initial value of β = 0.2,after successive stages of frequency doubling,is illustrated in Figure 4.4.The actual results,derived from relevant Astec 3 simulations are summarised in Figure 4.5.

4.3 Synchronous Detection of NBØM Signals after Frequency Multiplication.

The method originally proposed to recover both channels from the multiplexed orthogonal narrowband angle modulated signalling system is illustrated in outline form in Figure 4.6.As indicated,the demultiplexing technique relies on synchronous detection with quadrature coherent local carriers derived from the transmitted carrier component.However,the presence of any extra sidefrequency components produced by frequency multiplication will inevitably result in distorted baseband signals being recovered in both channels as the extra sidefrequencies are translated into the baseband as unwanted spurious components related to the original baseband frequencies.Another problem,as far as the QNBØM scheme is concerned,is that the extra sidefrequency components also represent an unwanted source of crosstalk as orthogonality between the two quadrature oriented modulations is no longer assured.This is dealt with in more detail in section 4.3.2.

In order to assess the relative magnitude of distortion obtained by synchronously detecting NBØM signals a series of investigations were undertaken using Astec 3 the results of which are summarised in Figures 4.7a and 4.7b.

4.3.1. Distortion of Baseband Signals Recovered by Synchronous Detection of NBØM.

Figure 4.7a shows the relative levels of harmonic distortion, at integer multiples of the sinusoidal component employed as the modulating frequency, for different values of frequency multiplication. In contrast, Figure 4.7b represents the results obtained for a more complex modulating signal whose actual waveform shape and power spectrum are given in Figure 4.8a and 4.8b respectively. As indicated, the spectrum of this complex baseband signal consists of four distinct frequency components which are odd and prime number multiples of a 200 Hz fundamental, which is itself excluded. i.e:-

$$m(t) = \sum_{\substack{i \\ i \neq 1 \\ i \neq \text{even} \\ i = \text{prime}}}^{11} \sin(\omega i t + \phi i) \qquad 4.15$$

Where i can take the values 3,5,7 and 11.

Consequently, the complex test signal conveniently covers a range of frequencies from 600 Hz to 2.2 kHz within the speech baseband. The amplitudes of the individual sinusoidal components are all equal, however, their relative phase values ϕi are chosen to ensure a composite waveform with a controlled peak-to-rms value (19). The distortion factor employed during the analysis of the results presented in Figure 4.7b is defined in Appendix C.The results represented graphically in Figure 4.9 correspond to actual baseband signals,recovered by synchronous detection,as obtained from relevant Astec 3 simulations.

4.3.2. Synchronous Detection of QNBØM Signals and Interchannel Crosstalk.

The QNBØM system relies on the orthogonality between the in phase and quadrature carrier modulations to enable the subsequent recovery of the two independent baseband signals.As previously indicated,this orthogonality is no longer assured after frequency multiplication which means that unwanted interchannel crosstalk is inevitably introduced.

Consider the following simple analysis for a QNBØM multiplexed signal:-

$$f(t) = \sqrt{2}\sin(\omega_c t + \pi/_4) + \beta_1 f_1(t)\sin\omega_c t + \beta_2 f_2(t)\cos\omega_c t \qquad 4.16$$

For convenience let $f_2(t) = 0$.Therefore:-

$$f(t) = \sqrt{2}\ \sin(\omega_c t + \pi/_4) + \beta_1 f_1(t)\sin\omega_c t \qquad 4.17$$

Applying this signal to a frequency doubler:-

$$[f_2(t)]^2 = \{[\sqrt{2}\sin(\omega_c t + \pi/_4)]^2 + 2\ (\sqrt{2}\sin(\omega_c t + \pi/_4).\beta_1 f_1(t)\sin\omega_c t) + \ldots$$

$$\ldots + (\beta_1 f_1(t)\sin\omega_c t)^2\} \qquad 4.18$$

After bandpass filtering:-

$$[f^2(t)]' = K\{\sin(2\omega_c t) + \sqrt{2}\ \beta_1 f_1(t)\sin(2\omega_c t - \pi/4) + \ \ldots$$

$$\ldots + \frac{\beta_1{}^2 f_1{}^2(t)}{2}\ \sin(2\omega_c t - \pi/2)\} \qquad 4.19$$

Again taking each term individually:-

(i) $\sin(2\omega_c t)$

Which corresponds to the resultant of the orthogonal carrier components which now have the form:-

$$\frac{1}{\sqrt{2}}\ \sin(2\omega_c t - \pi/4) + \frac{1}{\sqrt{2}}\ \sin(2\omega_c t + \pi/4)$$

(ii) $\sqrt{2}\ \beta_1 f_1(t)\ \sin(2\omega_c t - \pi/4)$

This term represents the wanted sidefrequency components of Channel 1.

(iii) $\frac{\beta_1{}^2 f_1{}^2(t)}{2}\ \sin(2\omega_c t - \pi/2)$

This term yields unwanted sidefrequencies which contribute to the distortion of the recovered Channel 1 baseband signal,as previously shown,but also leads to crosstalk being translated into Channel 2 by synchronous detection.

Since a more thorough analysis for complex baseband signals is particularly cumbersome,relevant Astec 3 simulations were used to investigate the potential distortion and crosstalk problem associated with using synchronous detection in demultiplexing and recovery of baseband signals in QNBØM.

The baseband waveform employed for this investigation is shown along with its power spectrum in Figure 4.10.The waveform was designed to have a specific amplitude spectrum,namely,a sinc(x) distribution from zero to 1.6 kHz with a peak value occuring at 800 Hz.As indicated by the power spectrum of Figure 4.10b,the signal consists of a fundamental at 200 Hz and harmonics up to and including the seventh.Spectral zeroes of the distribution are chosen to occur at d.c. and the eighth harmonic.The amplitudes of each harmonic are weighted according to the expression defined in Equation 4.20.

$$an = K\,\frac{\sin(x)}{x} \qquad 4.20$$

Where:- K = An amplitude scaling constant.

$x = \pi(n\tau/T-1)$ for n=1,2,3,4,5,6,7.

T = Waveform period (5ms)

τ = T/4

The phases of the individual components were again selected to yield a waveform with a well defined peak factor.

Having such a well defined spectral composition the extent of the distortion,introduced by synchronous detection,can be readily assessed by comparing the recovered baseband spectrum with the original amplitude spectrum.The three signals on the left hand side of Figure 4.11 (a,b and c) show the recovered signal which represents the wanted "speech-like" signal of Channel 1.On the right hand side,4.11d,4.11e and 4.11f show traces of the crosstalk observed on Channel 2.Ideally,the output on Channel 2 should be zero,however,Figure 4.11 clearly shows that the interference level gets progressively larger as the multiplication factor is increased.A useful figure of merit for assessing the degree of interference introduced is the "Signal to Crosstalk Ratio" (S/Xt dB) which is indicated on the three traces.The actual analysis of the results involving the extent of distortion and crosstalk introduced is dealt with in Appendix D.

4.4. Envelope Considerations of the QNBØM Signal.

Another significant factor,that emerged from the experimental work undertaken to investigate the effects of frequency multiplication on the QNBØM signal,was the magnification of the amplitude disturbance of the resultant RF signal.This is illustrated in Figure 4.12 which shows how the depth of envelope modulation of the transmitted signal is a function of the frequency multiplication factor employed and hence the frequency deviation of the composite signal.This situation is undesirable for the transmission of RF energy over a mobile radio channel where efficient use of available transmitter power is a very important consideration.

The results of this particular experimental work prompted a closer examination of the envelope characteristics of the resultant multiplexed signal and further theoretical analysis (Chapter 2) revealed that:-

(a) Amplitude limiting of the resultant signal is not a viable proposition to restrict the degree of envelope modulation produced since orthogonality between the two channels is destroyed.

(b) An alternative multiplexing scheme involving a novel hybrid form of amplitude/angle modulation (QNBFAM) is possible where the depth of AM can be controlled to ensure a reasonable degree of RF transmission efficiency.Also,the demultiplexing technique need not rely on synchronous detection but conventional envelope and frequency demodulators as dealt with in Chapters 2 and 5.

4.5 Concluding Remarks

The extent of the recovered baseband distortion,as well as the level of inter channel crosstalk inherent in the demultiplexing process,makes the QNBØM system rather unattractive as a suitable multiplexing scheme for the simultaneous transmission of speech and data over a mobile radio link.Also,since the scheme is reliant upon synchronous detection it means that the system implementaion requires coherent local quadrature carriers to be derived at the receiver from the incoming RF signal.Consequently,the system is prone to further problems of baseband signal distortion and interchannel crosstalk as dealt with in section 2.2.1 of Chapter 2 when discussing the QMDSB signalling technique.

The fact that the system is reliant on synchronous detection also means that existing transceiver equipment would require extensive modifications in order to receive QNBØM multiplex transmissions successfully.This feature would undoubtedly discourage PMR operators from adopting such a system.Large PMR operators would only appear to be interested in a system that could exploit existing equipment and also provide a measure of forward and/or reverse compatibilty with current operational systems.

Although the problems associated with implementing a QNBØM system mean that it is not a serious contender as a potential multiplexing scheme,nevertheless,the investigations undertaken have proved to be very worthwhile.In fact,the results obtained from the Astec 3 simulations led to the evolution of the QNBFAM system (Section 2.2.3) which is described in the following chapters.

(a) A.M. (b) NBFM

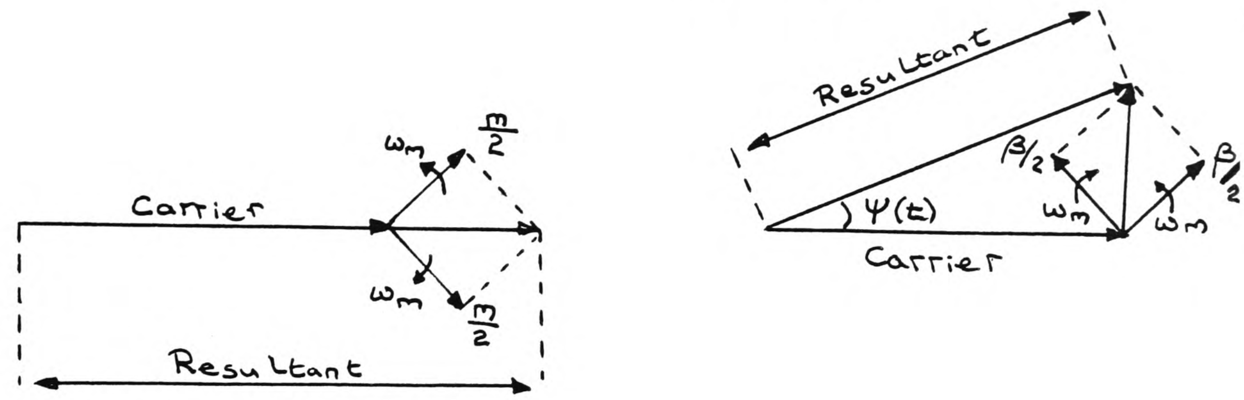
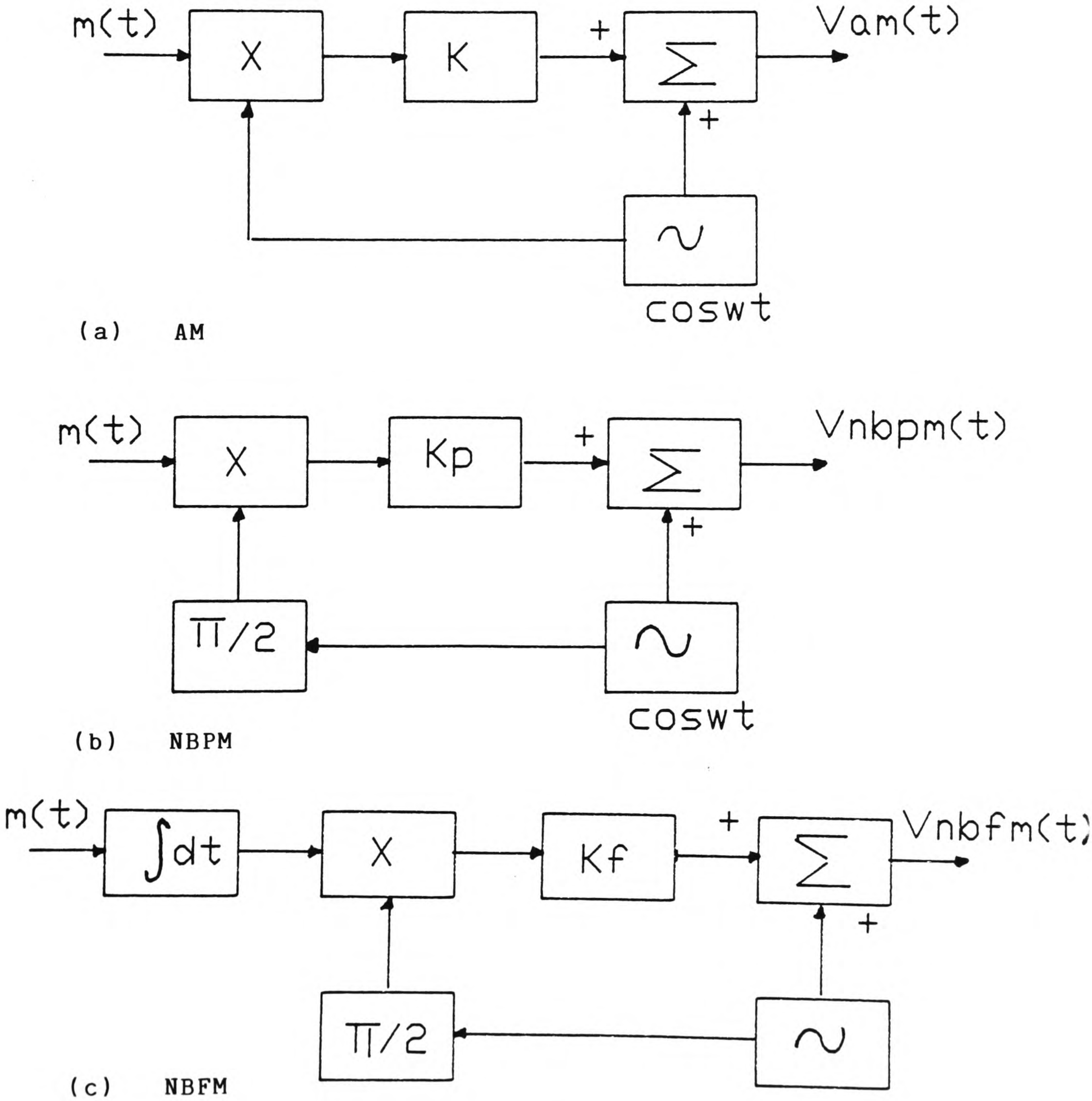


Figure 4.1.1 : Phasor Diagrams



(a) AM

(b) NBPM

(c) NBFM

Figure 4.1 : Generation of Modulated Signals

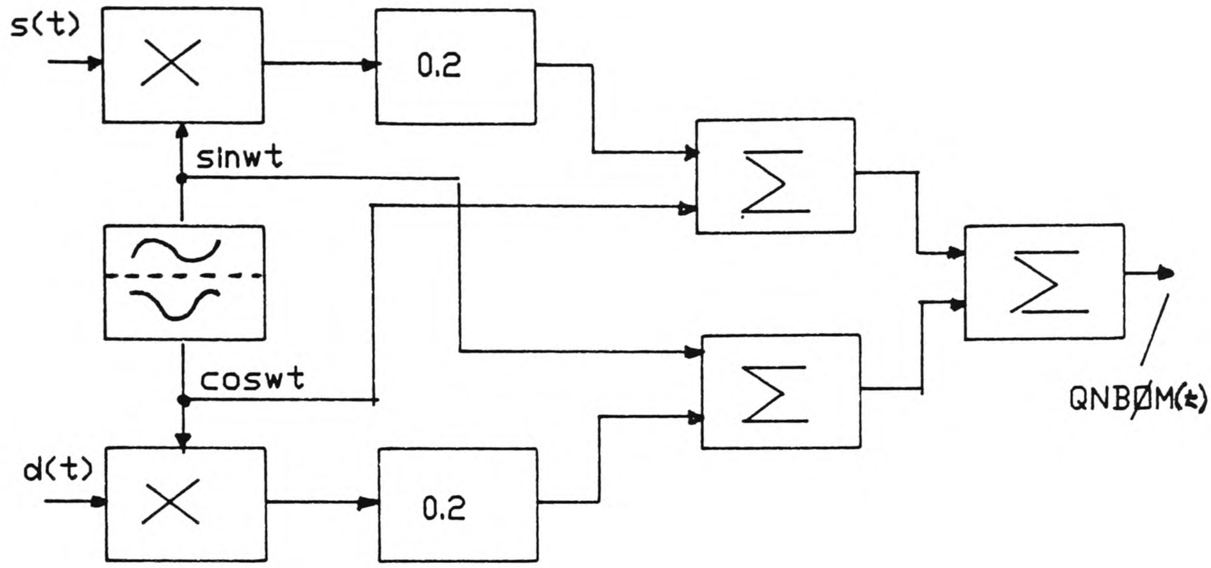


Figure 4.2 : The QNBØM Multiplexer (Low Deviation)

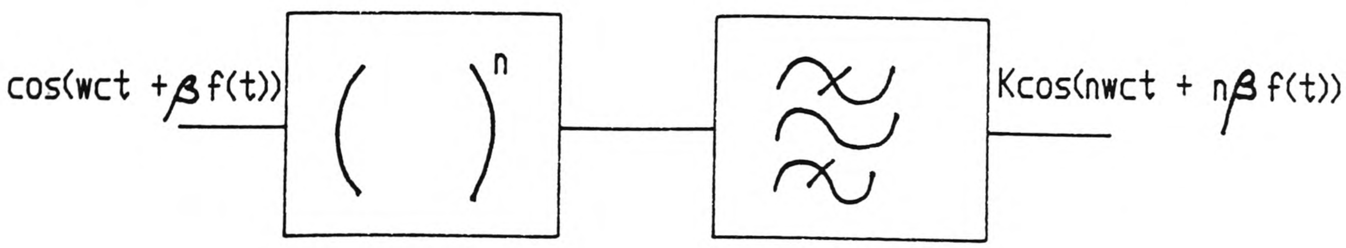


Figure 4.3 : Frequency Multiplication

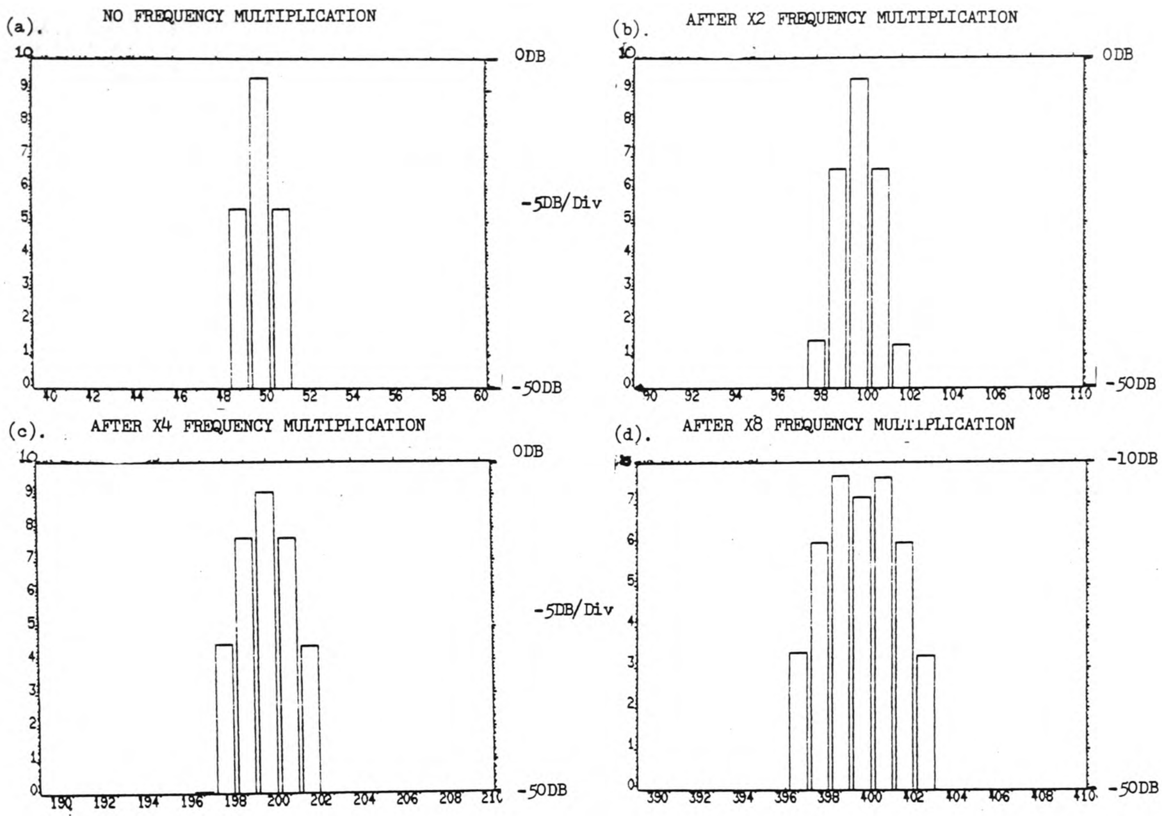


Figure 4.4. : Effect of Frequency Multiplication

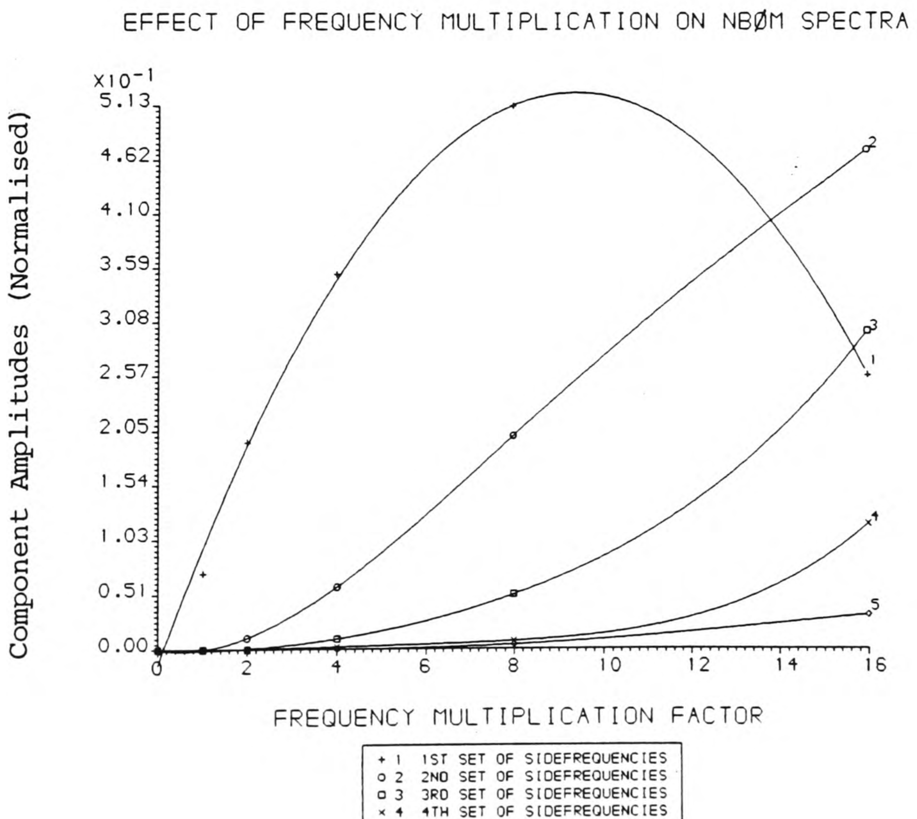


Figure 4.5 : Summary of Results

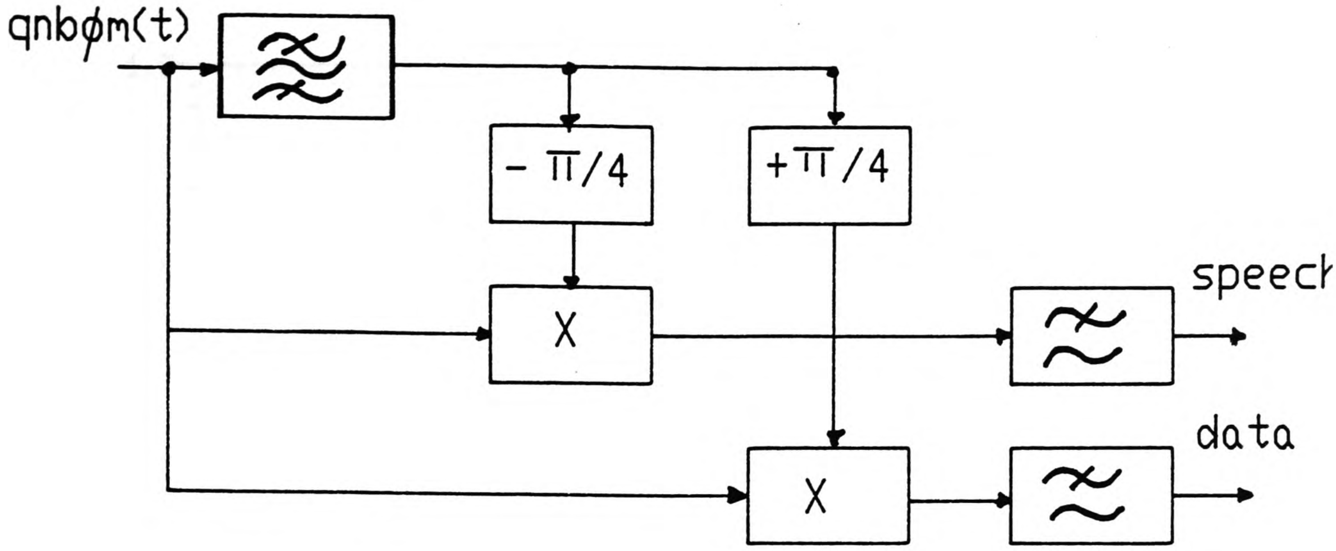


Figure 4.6 : Schematic of the QNBØM Demultiplexer

(a) (b)

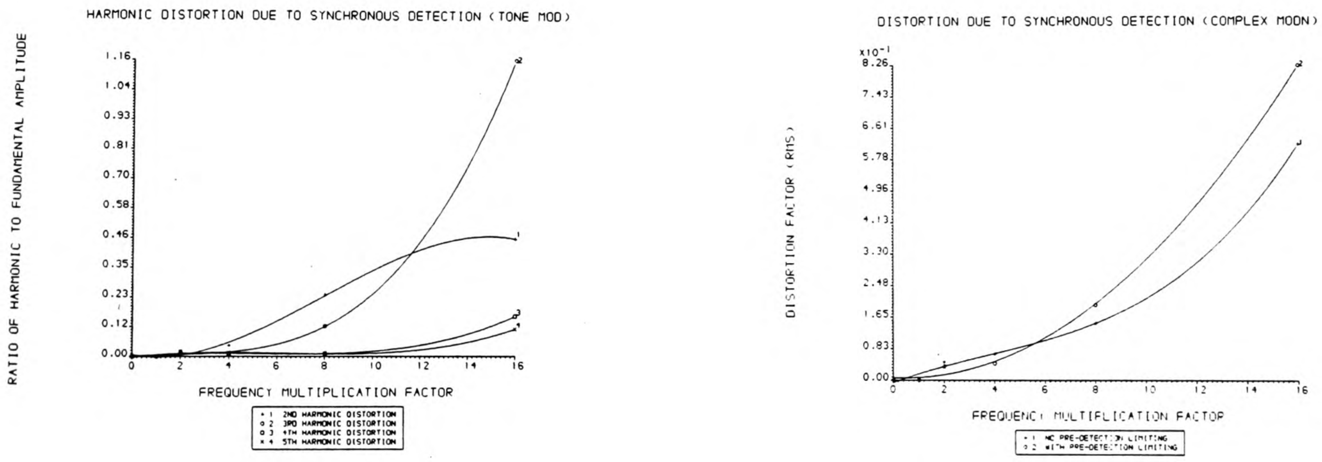
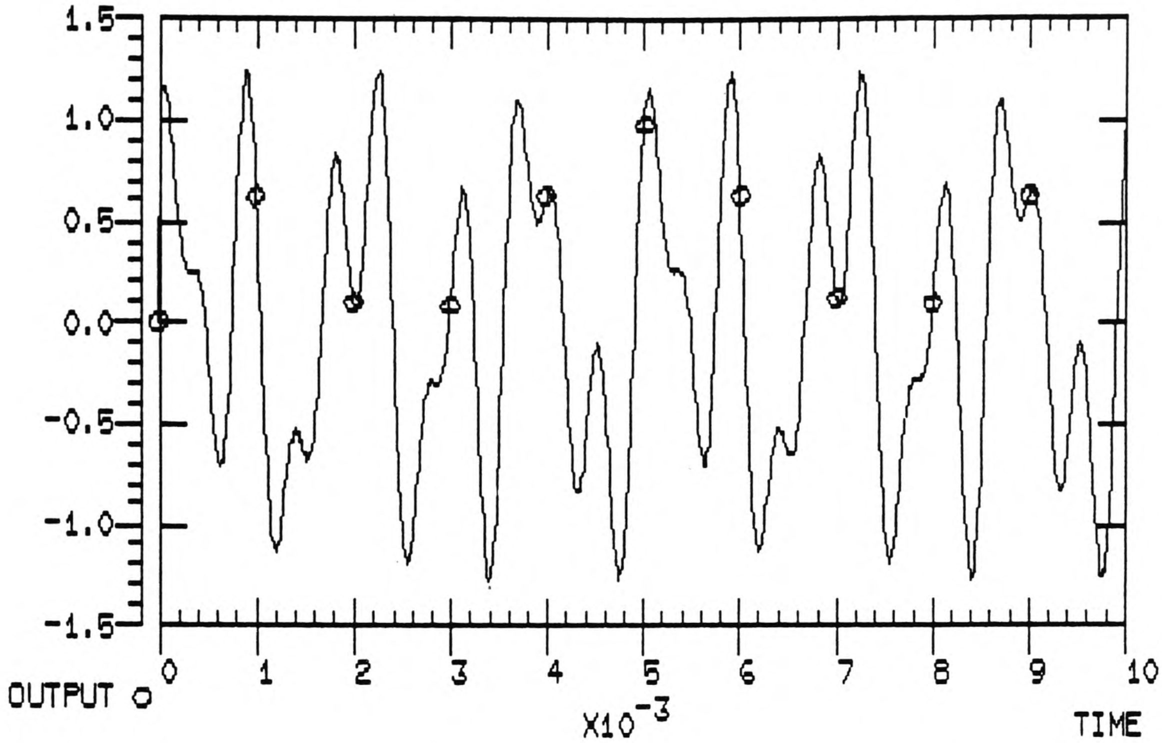


Figure 4.7 : Distortion of the recovered baseband signals

(a) The DISTOS Baseband Signal



(b) DISTOS Power Spectrum

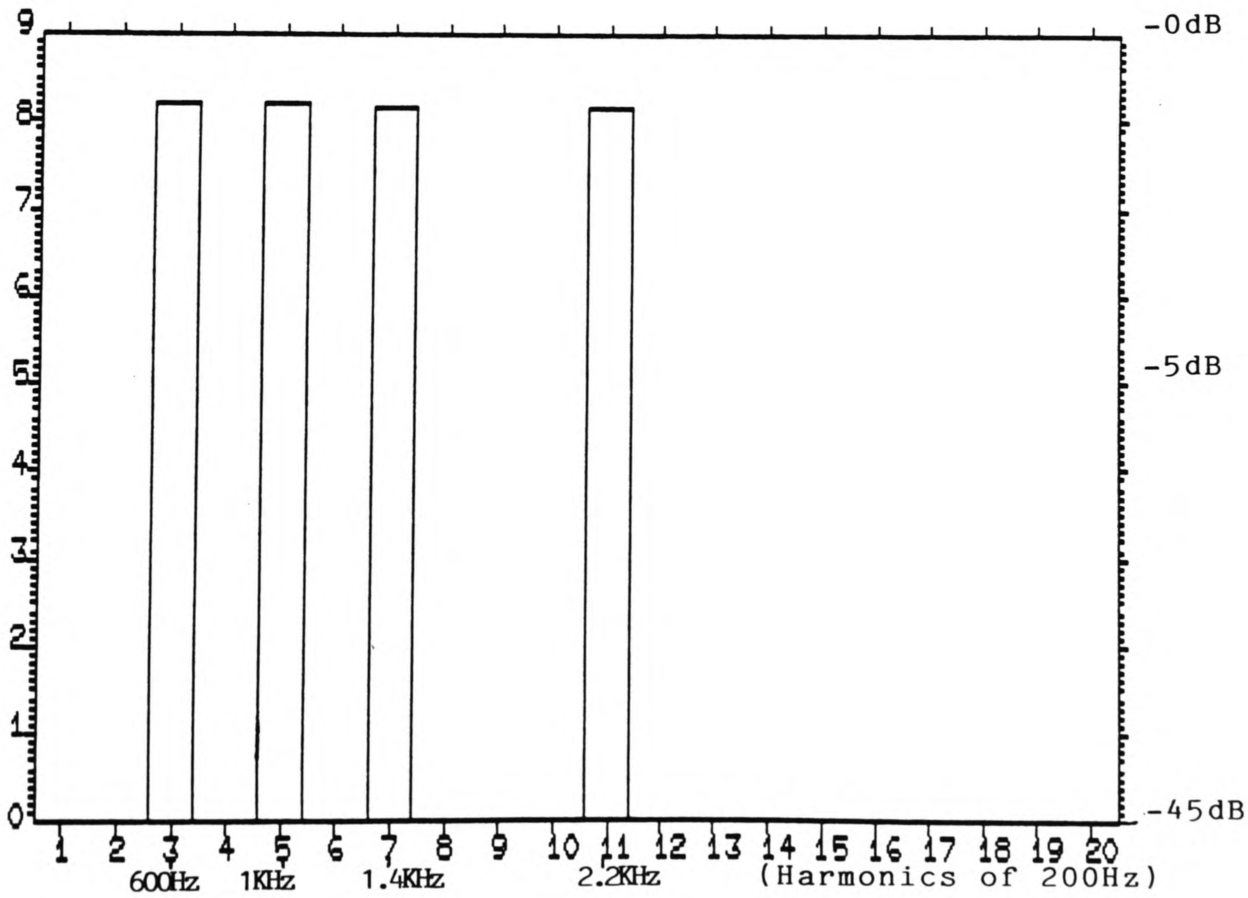
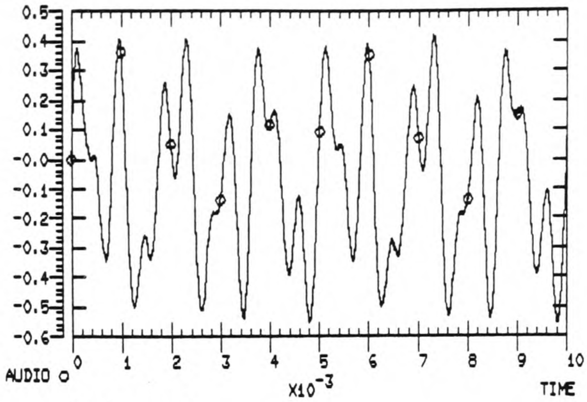
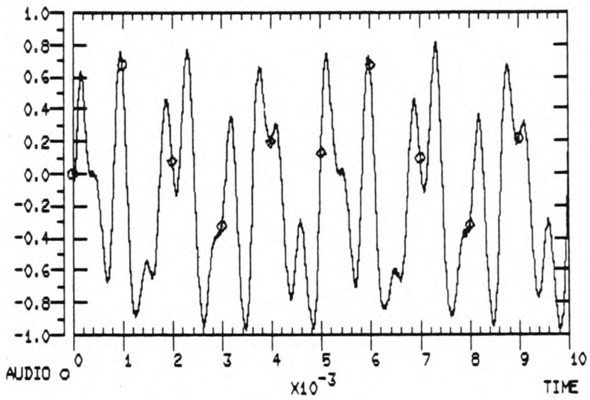


Figure 4.8

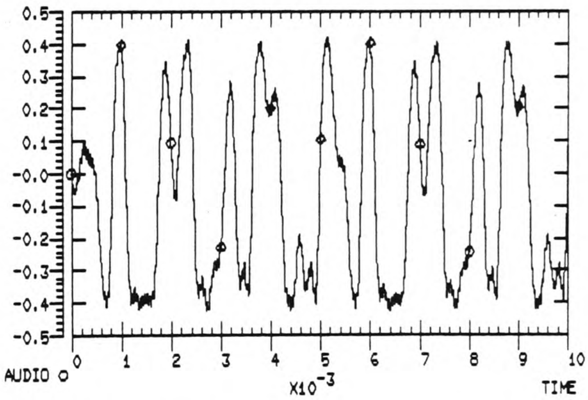
(a) x 2



(b) x 4



(c) x 8



(d) x 16

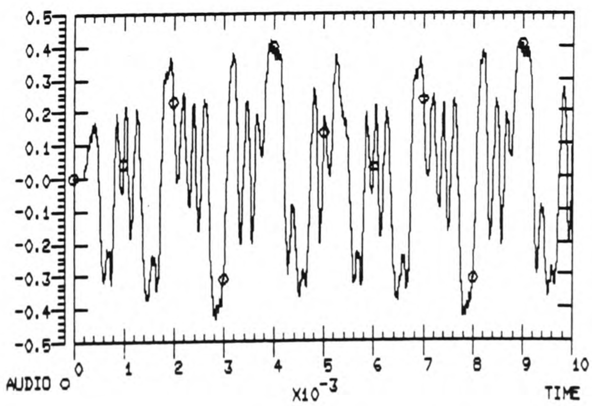
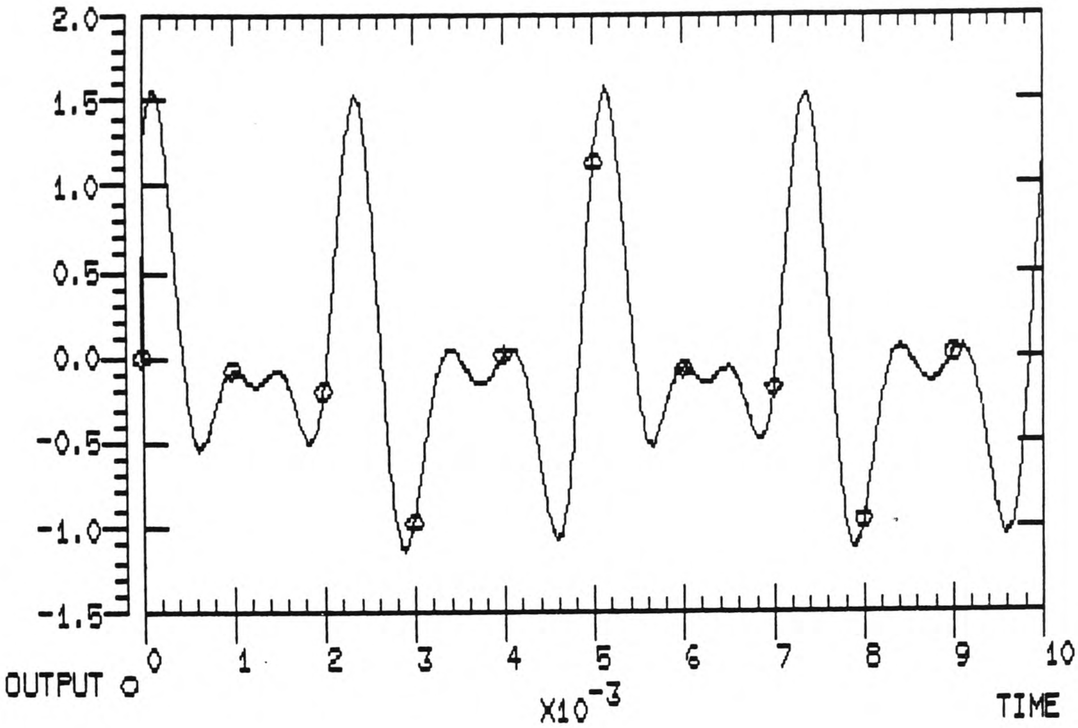


Figure 4.9 : Recovered Baseband Signals after Frequency Multiplication

(a) The SINXOS Baseband Signal



(b) The SINXOS Power Spectrum

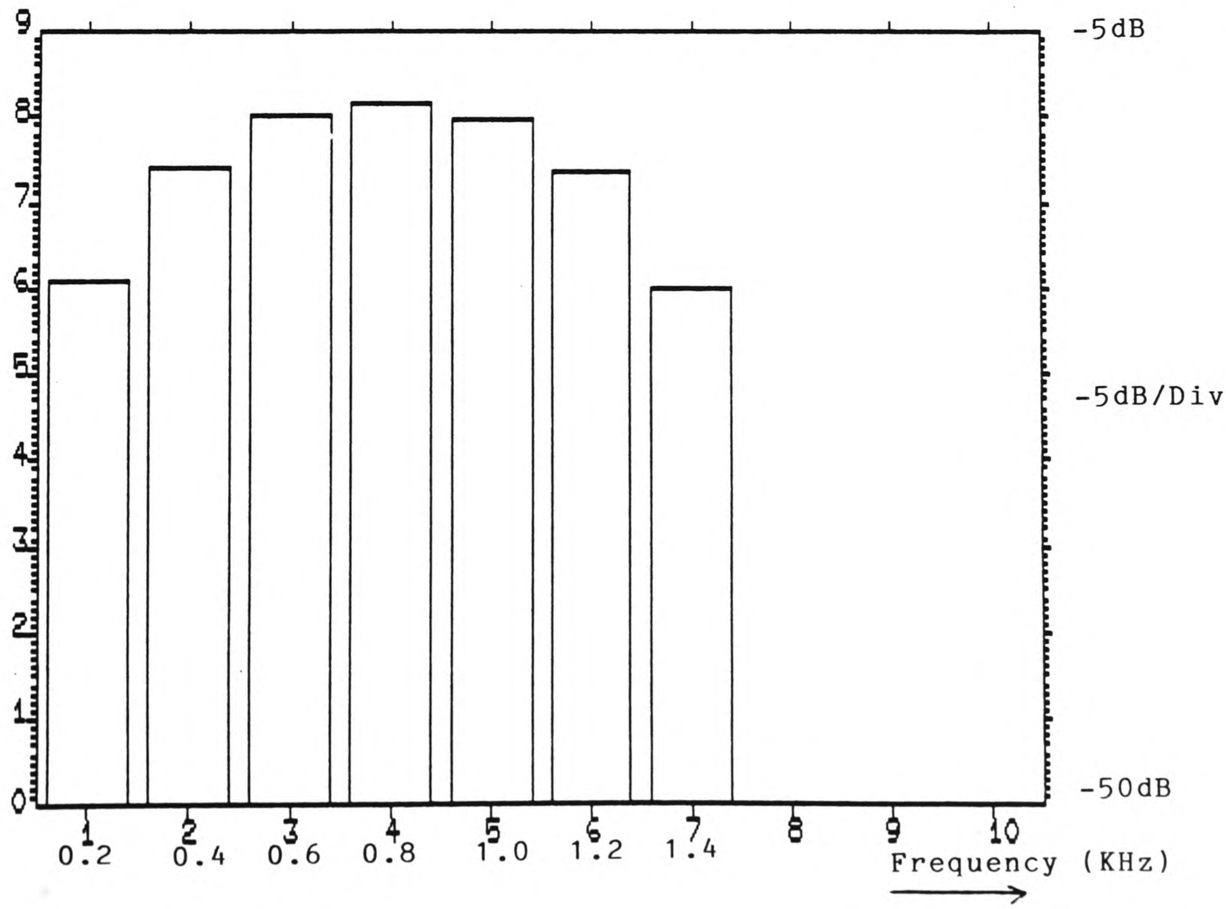


Figure 4.10

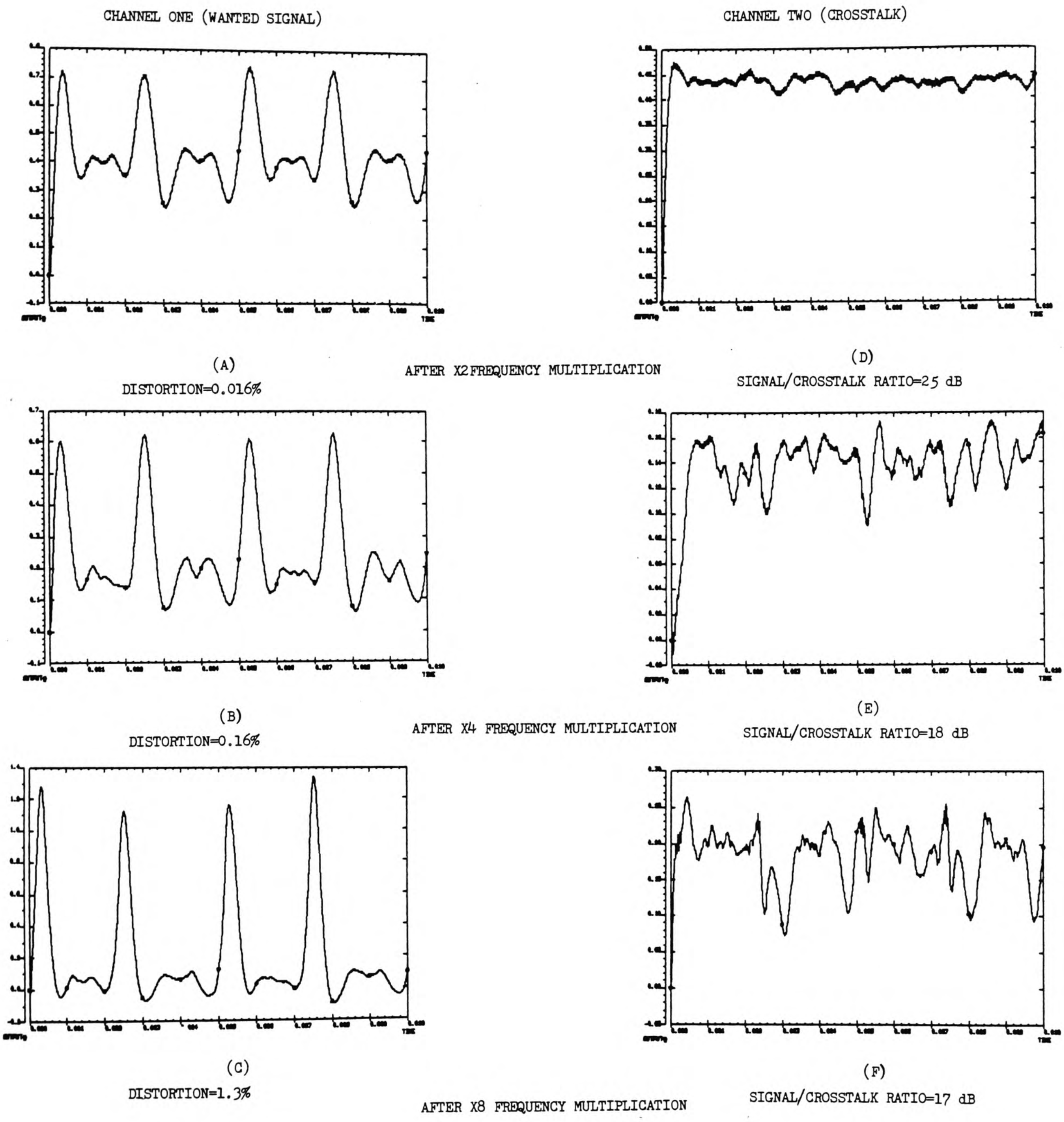
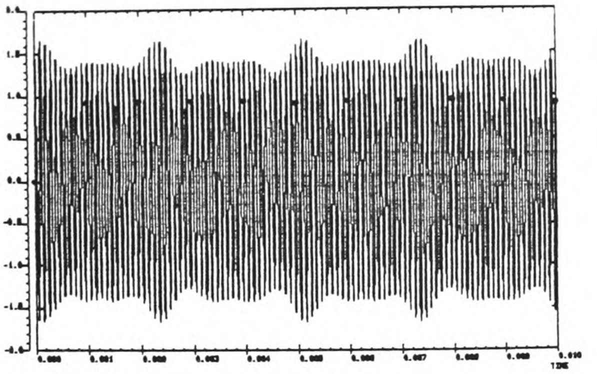
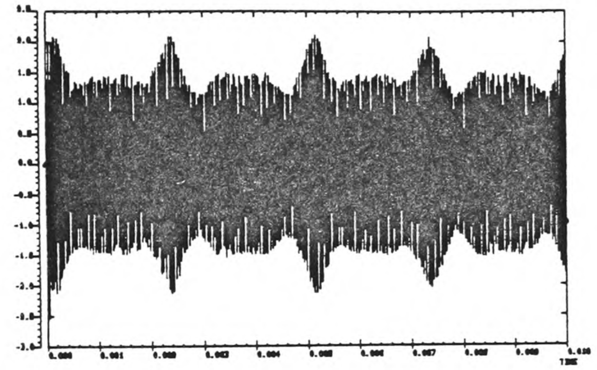


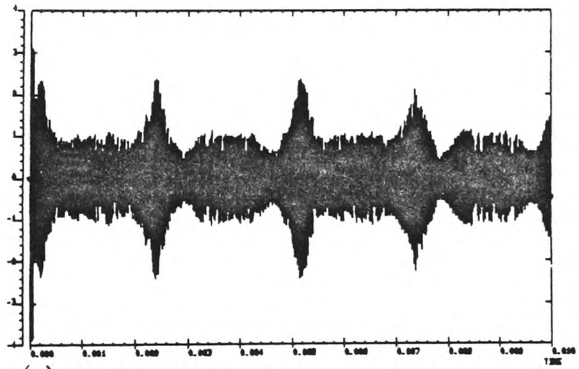
Figure 4.11 : Recovered Signals after Frequency Multiplication



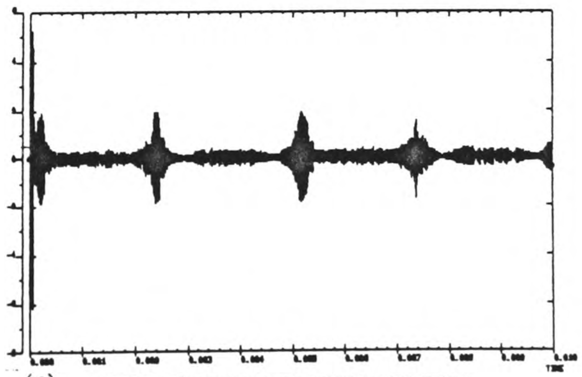
(a). NO FREQUENCY MULTIPLICATION



(b). AFTER X2 FREQUENCY MULTIPLICATION



(c) AFTER X4 FREQUENCY MULTIPLICATION



(d). AFTER X8 FREQUENCY MULTIPLICATION

Figure 4.12 : Envelope of the QNBØM signal after Frequency multiplication

CHAPTER 5

THE QNBFAM MULTIPLEXING SYSTEM

5.1 Implementing a Laboratory Prototype of the QNBFAM Multiplex System

As previously discussed,the QNBFAM multiplexing system emerged as a result of pertinent investigations conducted to assess the performance of the QNBØM system.Unlike other multiplexing strategies,considered in Chapter 2,the QNBFAM system offers a significant degree of forward and reverse compatibility with existing narrowband angle modulated radio systems.For example,the receiver configuration designed and developed for the QNBFAM system will actually permit the reception of conventional NBFM signals without any unecessary deterioration in quality.Also,with only one channel "active",either speech or data,the relevant quadrature carrier component can be disabled resulting in a transmitted signal which is actually a true narrow band angle modulated signal and can therefore be received by conventional VHF mobile radio equipment.

Throughout the development of the transmitter and receiver equipment,designed for the purpose of realising a QNBFAM simultaneous voice-data transmission,due consideration was given to the need to adopt standard techniques and approaches wherever possible.Such a design strategy was also encouraged by the various PMR operators involved in the initial discussions held concerning the desirabilty and feasibilty of a multiplex speech and data transmission facility.The attraction of a multiplex system which can be overlayed onto existing systems without the need for radical and fundamental equipment changes are obviously apparent.The actual transmitter/receiver equipment designed goes a significant way to meeting the system compatibilty requirements.In fact,in many instances,the signal processing functions required for the QNBFAM system and included in the laboratory prototype have actually been taken directly from commercially available transceiver modules.However,it is inevitable in order to successfully implement a novel integrated system which can support simultaneous transmissions of speech and data messages,without undue degradation of the baseband channels,that extra circuit functions and equipment need to be introduced which are not normally found in current VHF mobile radiotelephone transceivers.

5.2 The Transmitter Configuration.

Figure 5.1 represents in outline form the basic schematic of the QNBFAM transmitter.As indicated,the transmitter can be broken down into two distinct sections,namely,the IF and RF signal processing stages.In turn,these two separate stages can be further subdivided into several smaller functional operations which will be highlighted in due course.However,the design of the transmitter is relatively straightforward and readily implemented without the need for complicated and awkward alignment procedures.In addition,the transmitter is configured so that it can operate in either a single channel mode (speech or data) or a multiplex mode (speech and data simultaneously).This is achieved using the two switches S1 and S2.

With either switch S1 or switch S2 open,the QNBFAM IF signal appearing at the output of the summing circuit does not possess any AM since a carrier which is colinear with the sidefrequency components will not be present. Consequently,under these conditions the QNBFAM RF signal is a conventional narrow band angle modulated signal with no AM and a maximum carrier deviation of 2.5 kHz available for each individual baseband channel.However,with both S1 and S2 switches closed and modulation present in one channel only,the resultant RF signal will be a composite AM/NBFM signal with a frequency deviation half the value it would have if the single mode of operation were active.As dealt with theoretically in Chapter 2,with both baseband channels present the envelope of the QNBFAM signal is proportional to the difference between the speech and data signals,whereas,the angle modulation and hence frequency deviation is proportional to the sum of the two independent message channels.This means that the permitted 2.5 kHz maximum frequency deviation must be shared between both independent channels during a simultaneous transmission. Therefore,the maximum deviation per channel is now only 1.25 kHz when both quadrature carriers are present at the summing circuit.

The following sections identify and briefly describe the signal processing functions associated with the IF and RF sections of the QNBFAM transmitter.(Circuit diagrams for certain sections of the transmitter configuration are presented in Appendix J.)

5.2.1 The Carrier Source.

The transmitter carrier stability,required of mobile radio telephony equipment operating in the VHF band,means that the carrier source must be derived from a crystal oscillator arrangement.The oscillator circuit employed for this application is of traditional design,based on a JFET Colpitts configuration and operates in the parallel resonant mode (20).Although the exact frequency of oscillation can actually be adjusted slightly,by a series connected trimmer capacitor,careful setting of this frequency is necessary as it actually determines the final RF carrier frequency after it has passed through four stages of frequency doubling.Therefore,the choice of crystal (fXO) determines the channel on which the system operates in the VHF band (fRF=16*fXO).

The oscillator circuit is buffered by an emitter follower stage in order to avoid any unwanted interaction with successive stages which could otherwise disturb the oscillator performance.

5.2.2 Derivation of the I and Q Quadrature Carrier Components.

The I and Q quadrature carrier components,required for orthogonal multiplexing,are derived from the buffered crystal oscillator output signal.Various circuit arrangements were investigated for deriving the quadrature carrier sources,however,the most straightforward and effective approach was finally adopted.This technique relies on simple RC lead and lag circuits which,driven from a low impedance source and buffered at the output by JFET source followers,permit the carrier signal to be phase advanced and retarded by 45 degrees respectively resulting in two carrier components which are 90 degrees apart.With careful alignment this simple arrangement can be used to provide the required quadrature carrier sources with equal amplitudes from the stable crystal oscillator source.The high to low impedance output buffers provide the necessary isolation from the phase modulators which are driven by the separate quadrature carrier sources.

5.2.3 The Narrow Band Angle Modulators.

The circuit adopted to achieve narrowband angle modulation of the I and Q quadrature oriented carriers is based on a phase modulator configuration employed in many commercial VHF mobile transceivers.The principle of operation of this arrangement is described elsewhere (21) and is based on a technique referred to as "AM-C Vector Sum Phase Modulation".Basically,the IF carrier from the source oscillator and the buffered baseband modulating signal are combined together at the base-emitter junction of a transistor amplifier (Figure 5.2.3).The resultant signal at the collector of the transistor corresponds to the phasor sum of a direct IF carrier component,fed forward by a low value capacitor (typically of the order of 30 pF) and an IF carrier component phase inverted by the transistor amplifier which is actually amplitude modulated by the baseband signal.Consequently,the phasor sum of the in-phase and anti-phase IF components at the collector vary in phase by an angle determined by the amplitude of the baseband signal and at a rate proportional to the frequency of the baseband signal.The residual AM inherent with this technique of achieving angle modulation inevitably has to be removed by amplitude limiting.The actual limiter employed being based on a CA3028 linear high gain IC amplifier with a tuned collector output.

Figure 5.2 represents the performance characteristics obtained from the angle modulators.The two separate modulators built to accomodate both channels having characteristics which tracked each other sufficiently closely as to be considered identical.

5.2.4 Addition of the I and Q Modulated Carriers.

The QNBFAM IF signal is obtained by combining the two narrowband angle modulated signals whose carriers are in phase quadrature.As previously discussed,this combined signal will have an envelope related to the difference between the two independent information channels and an angle modulation proportional to the sum of the same.The configuration adopted to achieve the addition of the two quadrature oriented angle modulated signals is based on a summing amplifier which employs a transistor operating in the common base mode.The common base amplifier offers linear,in phase,wideband voltage amplification coupled with a low input impedance which provides a measure of isolation between the two signal sources to be combined.The excellent linearity offered by the common base amplifier means that spurious outputs are not a problem providing the input signals do not drive the transistor into saturation.The resultant "sum" signal is extracted from the collector and again buffered before being coupled to the RF section of the transmitter which in turn consists of three separate subsections.

5.2.5 The Frequency Multiplier.

As described in Chapter 4,the function of the multiplier is twofold.Firstly it is needed to increase the frequency deviation of the QNBFAM signal delivered from the IF section of the transmitter and,secondly,to translate the IF frequency to the final RF carrier frequency ready for transmission in the VHF band.The actual multiplier used was taken directly from an existing mobile radiotelephone, namely, the "PYE Europa" (Type MF5FM).This configuration employs four frequency doublers resulting in a frequency multiplication factor of sixteen.The first two doublers employ JFET frequency doublers which permit the use of untapped coils to be used in the drain tuned circuits.The remaining pair of doublers use bipolar junction transistors which because of their lower input and output impedance require tapped coils in the interstage tuned coupled circuits.The multiplier terminates in a single stage transistor pre-driver which in turn provides the RF carrier drive to the AM modulator of the transmitter.

5.2.6 The Envelope Detector and Amplitude Modulator.

Frequency multiplication of the QNBFAM IF signal is achieved by transistor amplifier circuits which are operated on a non-linear part of their transfer characteristic.This means that the envelope information associated with the QNBFAM signal is destroyed by the frequency multiplier stage.However,this information plays a vital role in the separation of the speech and data baseband channels at the receiver and must therefore be preserved.This preservation is achieved by first detecting the envelope of the QNBFAM IF signal prior to frequency multiplication.Then restoring this information later by modulating the amplitude of the RF signal,delivered by the frequency multiplier,with the waveform derived from the IF signal envelope.

Since RF power was not an operational feature required of the QNBFAM system laboratory prototype,a low level AM modulator based on a double balanced mixer was developed.The double balanced mixer used is a standard +7dBm LO input power mixer (Hatfield RF components Type 1763) and is specified for operation up to 500 MHz.It utilises an array of Schottky barrier (hot carrier) diodes which are matched for maximum performance and reliabilty.Amplitude modulation is readily achieved by applying the low level RF output of the frequency multiplier stage to the RF port of the balanced mixer and a bias current,which is actually controlled by the waveform derived from the IF envelope information,to the IF port (22).The resultant QNBFAM signal with its envelope now restored is available at the remaining LO port of the mixer.

The bias current to the IF port needs to be carefully set in order to ensure that the operating point of the mixer is established on a linear part of the "Attenuation/IF port dc current" characteristic of the mixer (23).The bias current is derived from a current source which is driven by a transconductance amplifier.The transconductance amplifier is designed in such a way as to permit a dc control current to be established which is then made to vary in a linear fashion in accordance with the amplitude of the low frequency waveform corresponding to the envelope information.In this way low level amplitude modulation has been successfully achieved.Several practical investigations were conducted to determine the optimum operating conditions of the amplitude modulator.These were found to be an RF carrier drive level of 0dBm to the RF port with a dc bias current to the IF port of 0.5mA.The gain of the transconductance amplifier being such that an input modulating signal of approximately 2 volts r.m.s. produces a modulation depth of 100%,however,this extreme depth of modulation is not envisaged for the system in question.

Although the linearity of the AM modulator designed proved to be very satisfactory,its baseband response was found to require frequency equalisation.This was needed in order to ensure the depth of modulation introduced was determined solely by the amplitude of the modulating signal and not its frequency.Practical measurements revealed that without equalisation the depth of modulation fell by a factor of 10% over the frequency range of 300 Hz to 3 kHz. Fortunately,introducing a passive pre-emphasis network between the transconductance amplifier,which controls the bias to the balanced modulator,improved the situation significantly.The practical performance characteristics of the final AM modulator are summarised conveniently in the graphs of Figure 5.3.

The original circuit employed for detecting the envelope of the QNBFAM IF signal was based on a conventional diode detector arrangement.However,this was subsequently replaced by an integrated circuit version,the Plessey SL623C (24),which offered a significant improvement in terms of linearity of performance and ease of interface with the pre and post detection signal processing circuitry.

The final circuit of the QNBFAM transmitter,which delivers the RF signal output,is an elliptic-function filter which is employed to minimise unwanted spurious outputs.The network used was taken directly from the configuration employed in a PYE Europa VHF mobile radiotelephone which normally performs the function of coupling the transmitted signal from the RF power amplifier to the antenna.

5.3 Introducing the QNBFAM Receiver.

Figure 5.4 represents in schematic form the configuration developed to receive and demultiplex a QNBFAM simultaneous transmission of speech and data. Essentially the receiver conforms to a standard double conversion superhet with a first IF of 10.7 MHz and a second IF of 455 kHz with the receiver selectivity being defined by a crystal block filter following the first mixer stage. The laboratory prototype design does not include an RF amplifier as sensitivity was not a parameter initially considered a limiting factor in the successful operation of the demultiplexing strategy proposed. In fact, initial investigations with a commercial VHF front end were hampered somewhat by over sensitivity because of the significantly high radiation emanating from the transmitter prototype. These problems were obviated by not employing an RF amplifier prior to the receiver first mixer. However, careful measurements were required in order to calibrate the system in terms of relative input signal levels which could in turn be referred to the front end of a commercial VHF mobile receiver which does employ an RF amplifier. The calibration chart used to refer the front end attenuator settings to effective RF input level is shown in Figure 5.5.

Although the receiver conforms to a standard double conversion superhet the overall system requires three distinct functional blocks not normally found in standard VHF mobile radiotelephone equipment. Namely:-

(i) A "linear-phase" crystal filter at the first IF.

(ii) Delay equalisers and a channel decoder matrix.

(iii) A "linear" gain controlled IF section which drives an AM detector.

The reasons for employing the networks,identified in (i) to (iii) respectively,will be established in Chapter 6 where practical performance measurements of the demultiplexing process will also be presented and discussed.

The remaining circuitry used in the receiver prototype employ modules taken directly from a PYE Europa transceiver with the exception of various buffer stages and the second mixer and FM IF strip.The second mixer and FM IF strip were implemented with a chip set produced by Mullard Ltd.,namely,the SA/NE602 double balanced mixer and oscillator as well as the SA/NE604 low power FM IF system.The SA/NE602 is a monolithic double balanced mixer with an integral oscillator and voltage regulator.The SA/NE602 can be used in applications up to 200 MHz and boasts low current consumption,excellent noise and third order intermodulation performance as well as low external component count.The SA/NE604 is a monolithic low power FM IF system incorporating two limiting intermediate frequency amplifiers,quadrature detector,audio muting,logarithmic signal strength indicator,and similar to the SA/NE602,a built in voltage regulator (25).

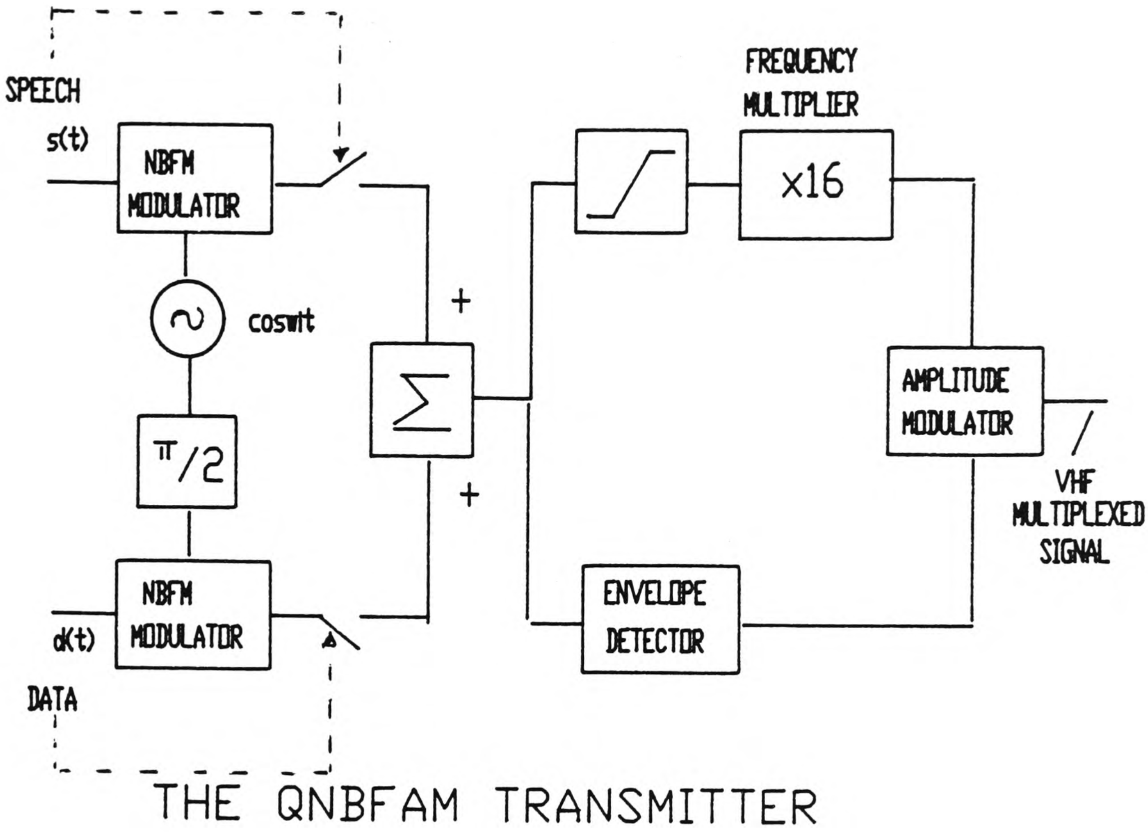


Figure 5.1

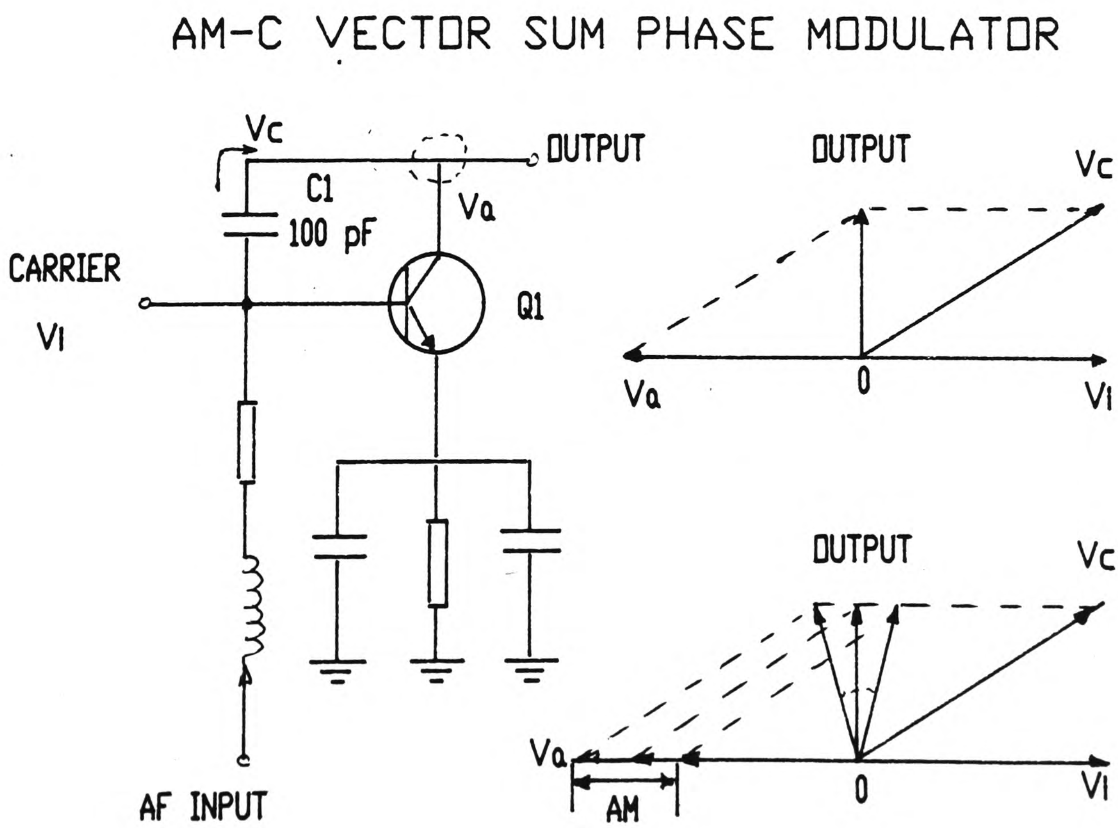
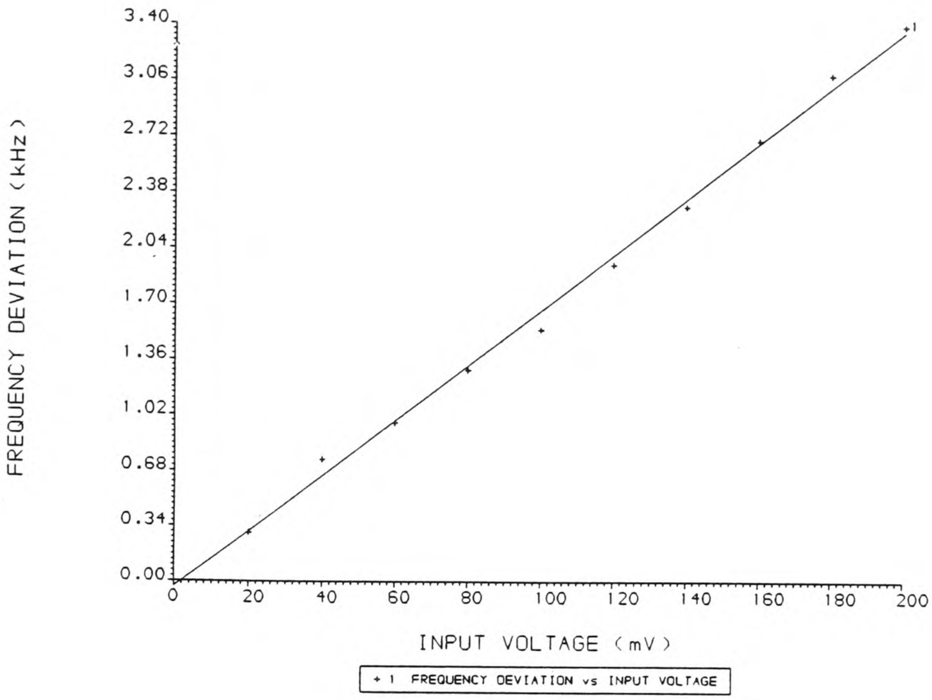


Figure 5.2.3

(a)



(b)

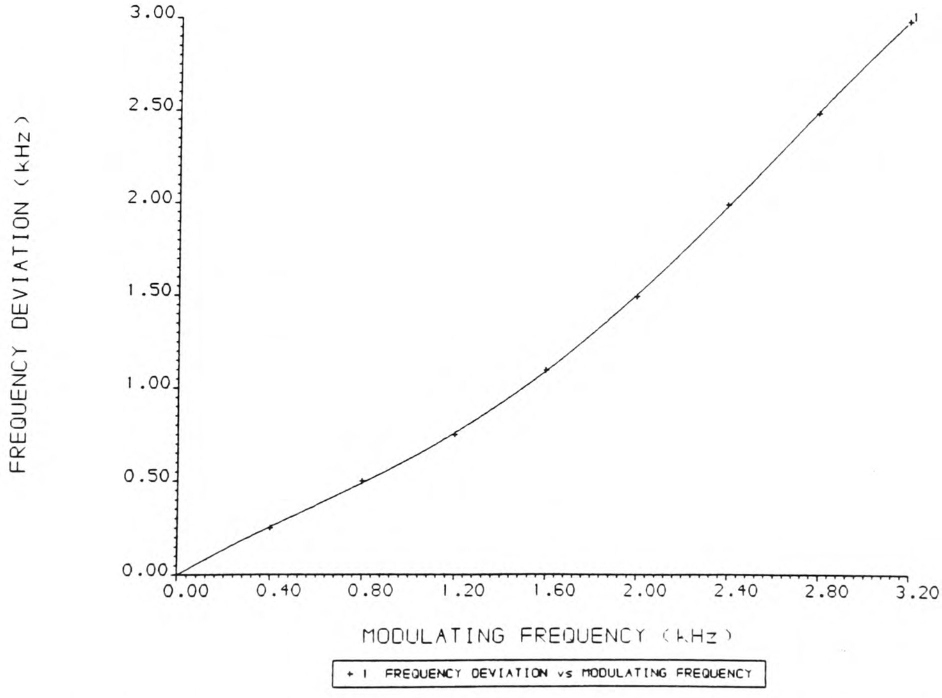
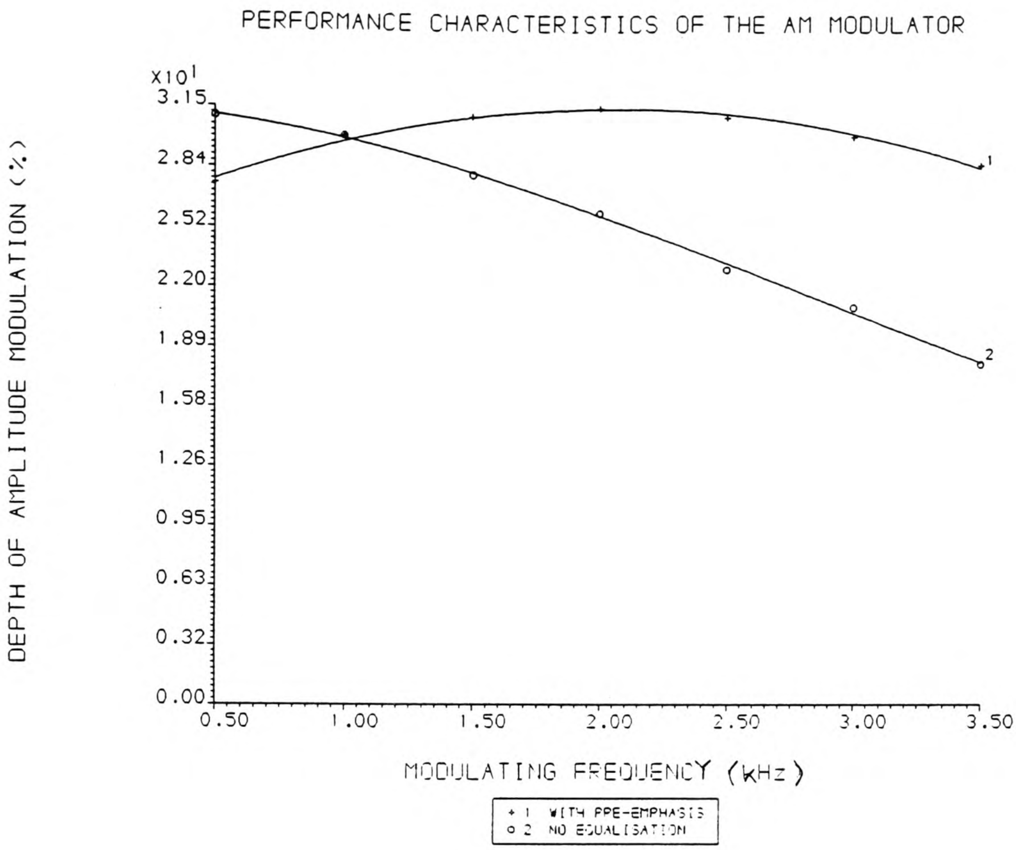


Figure 5.2 : Performance Characteristics of the Narrowband Angle Modulators (After x 16 frequency multiplication)

(a)



(b)

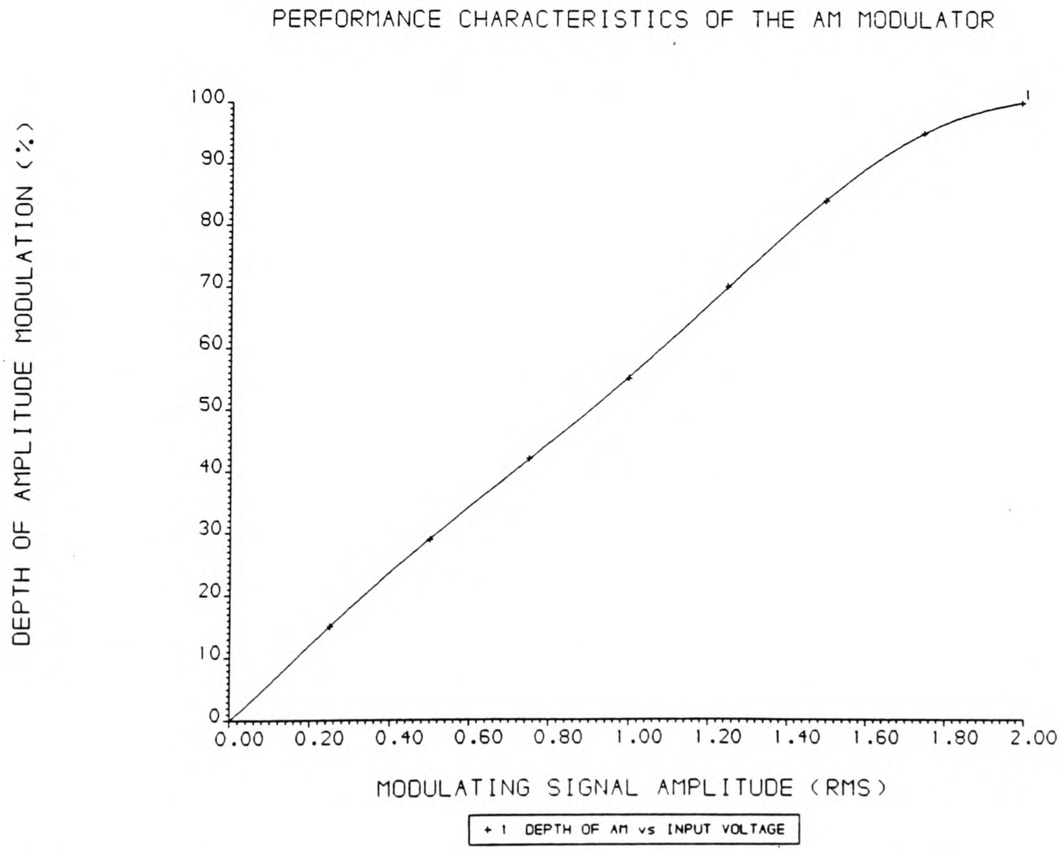


Figure 5.3 : Performance Characteristics of the Low-Level AM Modulator

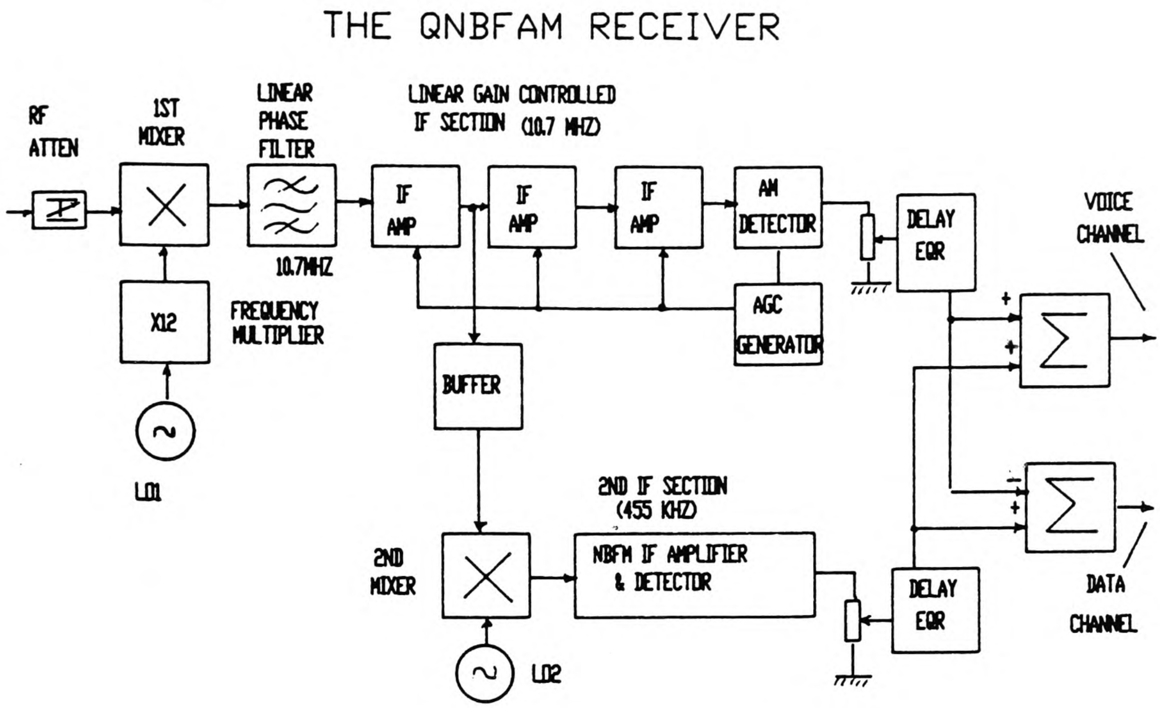


Figure 5.4 : The QNBFAM Receiver

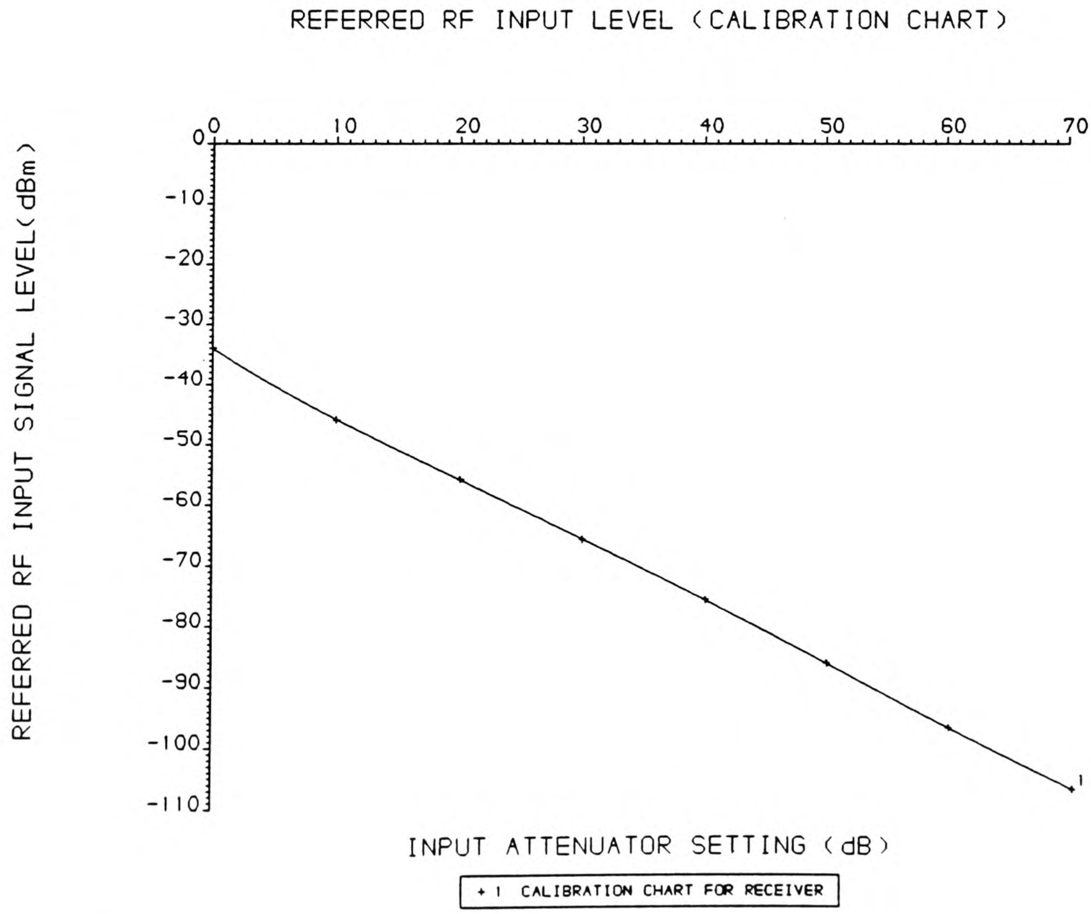


Figure 5.5 : Referred RF Input Level (Calibration Chart)

CHAPTER 6

PERFORMANCE OF THE QNBFAM RECEIVING SYSTEM

6.1 The Need for a Linear Phase Filter.

The adjacent channel selectivity requirements of conventional narrowband angle modulated receivers,operating in the VHF band,is normally achieved by employing a crystal block filter after the first stage of frequency conversion.The specification for a typical block filter is listed below:-

Pass bandwidth	=	3.0 dB at +/- 3.75 kHz
Stop bandwidth	=	90.0 dB at +/- 12.75 kHz
Passband ripple	=	2.0 dB max
Insertion loss	=	4.5 dB

As indicated,such filters are characterised by a very sharp attenuation skirt with a certain degree of amplitude response variation in the passband.However,one important parameter of the filter characteristic not normally specified for mobile radio applications involving speech telephony,but particularly important for the application in question,is the group delay characteristic.In order to achieve the desired amplitude response the phase characteristic obtained departs considerably from the ideal resulting in a significant group delay variation over the passband of the filter.A non-linear phase characteristic which produces a group delay variation results in the distortion of modulated signals which propagate through such filters.

The analysis contained in Appendix E shows that the distortion effects on modulated signals can be summarised as follows:-

(i) Modulation conversion.(PM to AM and AM to PM)

(ii) Change of modulation depth or index.

(iii) Non-linear envelope distortion.

(iv) Intermodulation.(When more than one modulating component is present)

For normal speech telephony applications the presence of the angle modulation results in the envelope of the signal acquiring a residual AM component (Figures 6.1 and 6.2) which can normally be eliminated by pre-detection amplitude limiting.Also, the degree of distortion produced by the non-linear phase characteristic is normally tolerated as intelligibilty rather than fidelity is the main performance criterion to be achieved.However,unfortunately this is not the case with the QNBFAM multiplexing scheme under consideration and a linear phase characteristic with negligible group delay distortion is an essential feature required of the IF filter.The reason for this,of course,is that the envelope of the multiplexed RF signal contains information essential to the demultiplexing process. Consequently,the envelope of the modulated signal must therefore be maintained throughout and any residual AM component produced by modulation conversion due to a non-linear phase characteristic will inevitably lead to unwanted crosstalk and inadequate channel separation.The presence of AM on the composite RF signal will also lead to unwanted angle modulation which will also contribute a crosstalk component which cannot be eliminated in the channel decoding process.

Therefore a crystal filter with a linear phase characteristic is clearly a pre-requisite for this particular application.The filter adopted for the laboratory prototype receiver is a FD800B obtained from ECM Electronics Ltd.The important filter specifications are listed below:-

Pass bandwidth	=	3.0 dB at +/- 5 kHz
Attenuation	=	60.0 dB at +/- 35 kHz
Passband ripple	=	0.5 dB max
Insertion loss	=	3.0 dB max
Group delay distortion	=	5.0 μs max (10.7 MHz +/- 5 kHz)

By inspection,the attenuation characteristics of this crystal filter are significantly poorer than the previous crystal filter reviewed,however,the FD800B was the only filter obtainable that could provide the linear phase response demanded by the system being investigated.

Practical measurements revealed that the amount of spurious AM produced by propagating a true NBFM signal through the FD800B crystal filter was negligible,being less than 3% with the maximum possible frequency deviation of 2.5 kHz.

6.2 The Demultiplexing Process.

The circuit functions required to implement the demultiplexing process enabling the independent speech and data channels to be recovered are actually very straightforward.The output from the AM detector yields a signal which is proportional to the difference between the two independent channels,whereas,the output from the FM detector is related to the sum of the two baseband channels.Therefore,by employing circuits that can perform simple addition and subtraction the speech and data channels can be separated.This process is conveniently summarised in Figure 6.3.

Of course,the successful operation of the channel decoder relies on the cancellation of the unwanted signal and enhancement of the wanted signal at the relevant channel output which in turn requires the demodulated signals to be matched in terms of their amplitude and relative phase relationships.In reality such channel cancellation and enhancement,which effectively separates the two channels,can never be absolute over the modulation baseband due to a number of different factors.Consequently, crosstalk between the speech and data channels is a parameter the system has to cope with when both channels are being transmitted simultaneously.However, results obtained in practice have indicated that data and speech channel performances can be obtained that do not suffer unduly from the crosstalk between the data and speech channels providing the demodulated signals are correctly equalised in terms of their amplitude and relative delay characteristics.

Figure 6.4 illustrates the operational performance of the channel decoder over the baseband of 300 Hz to 3 kHz for two different types of delay or phase equalisation network.As indicated,the better performance is obtained by an equaliser based on a SAD1024 "Bucket Brigade" analogue delay line integrated circuit (26).However,a satisfactory operational performance can still be obtained even when a straightforward all pass filter network is employed to effect the necessary delay equalisation (27).Figure 6.5 represents a more meaningful indication of the Signal to Crosstalk performance available from the system by displaying the actual spectral representation of the waveforms obtained at the decoder outputs when both speech and data signals are present simultaneously.In this particular case the data being transmitted corresponds to a FFSK waveform (Appendix F) modulated by a pseudo random binary data sequence at a rate of 1200 baud (28).In contrast,the speech like waveform corresponds to a band limited random noise signal.In both cases,each baseband signal modulates the transmitted carrier with a maximum frequency deviation of 1.25 kHz and a depth of AM restricted to 25% maximum.(A summary of the baseband signals used and the corresponding QNBFAM RF spectra is presented in Appendix B.)

The effect of the crosstalk from the speech channel to the data channel inevitably affects the bit error rate (BER) performance available.By controlling the amount of crosstalk between channels the effect on the error rate performance can be readily investigated.Figure 6.6 indicates that the important parameter which determines the overall BER attainable is the Signal to Crosstalk Ratio (S/Xt dB) and as illustrated a S/Xt(dB) greater than 12 dB is required to ensure that the error rate is less than one error in 100000 bits.These results were obtained with bandlimited random noise modulating the speech channel with a maximum frequency deviation of 1.25 kHz in each signalling channel.In contrast,assessing the performance of the speech channel,when subjected to crosstalk from the data channel,is ultimately a matter of opinion as far as quality of reproduction and level of impairment is concerned.However, when correctly aligned and employing the standard five point scale for subjective assessment of quality and impairment (29),scale factors of at least 4,in each category,are attainable.This corresponds to "good" quality sound with the impairment "perceptible but not annoying".

6.3 Need for a Linear Gain Controlled IF Strip.

In order to separate the independent speech and data channels the demultiplexer requires the baseband signals,recovered by the AM and FM detectors,to be matched in amplitude and relative phase.However,although the output from the FM detector is determined by the frequency deviation of the received carrier,the corresponding output from the AM detector is dependent upon the IF carrier drive to the demodulator as well as the modulation depth.Therefore,if the received signal strength varies so will the output from the AM detector and hence the demultiplexer performance will suffer in terms of channel separation attainable.This predicament is represented by Figure 6.7 which indicates how the demodulated outputs and resultant decoded channel output signal amplitudes vary with received input signal level.The nett effect of the AM detector output variation with received signal strength results in a serious deterioration in the signal to crosstalk ratio in each channel and hence an inferior speech and data channel performance.This situation is unacceptable particularly when one considers the mobile radio environment in which it is proposed to operate the system (30).

Figure 6.8 indicates the data channel performance obtained in terms of the measured Signal to Crosstalk available as a function of the received signal strength.Error rate measurements undertaken confirm that for a BER performance better than one error in 100000 bits the signal variations which can be tolerated,about the nominal calibrated level of the decoder,are actually only of the order of +/- 6 dB.These practical results confirm the need for employing an automatic gain control (AGC) system in order to maintain the IF drive level to the AM detector reasonably constant over the dynamic range of signal strength anticipated in practice.This is necessary to ensure adequate data and speech channel performances by the system over this range.

The linear gain controlled IF strip developed to determine whether a straightforward AGC control system would extend the operational dynamic range available to the system was based on a chip set from Plessey Semiconductors Ltd.Namely,the SL623 AM detector and AGC amplifier in conjunction with three SL612 IF amplifiers.The main features of the SL623 are negligible distortion,straightforward interfacing and a fast response time.The AGC voltage of this device is actually generated directly from the detected carrier signal and is therefore independent of the depth of modulation.Also,its response is considered to be fast enough to follow the most rapidly fading signals (24).In contrast,the SL612 is an RF voltage amplifier with a maximum voltage gain of 34 dB and an upper -3 dB cut-off frequency of 15 MHz.The main features of the SL612 are a wide AGC range of 70 dB,ease of interfacing and integral power supply RF decoupling (31).The gain control voltage range of the SL612 is matched to the AGC generator characteristics of the SL623 device emphasising their suitability for the configuration proposed.Figure 6.9 indicates how the presence of AGC extends the dynamic range of the system as required.The graph shows how the data channel performance,in terms of operational Signal to Crosstalk ratio,has been extended to cover a range of 70 dB with the performance only deteriorating below the required value when the RF input level falls below 1 μV.This is confirmed by the error rate profile which is superimposed on the graph of Figure 6.9.Figure 6.10 represents a selection of spectral plots obtained at the data channel output for different RF input attenuation values from 0 dB to 80 dB.As indicated by the BER profile of Figure 6.9,the crosstalk from the speech channel to the data channel increases significantly for RF input signal levels less than 1 μV which corresponds to an input attenuation setting of 80 dB.The reason for the increase in crosstalk at the other extreme,0 dB input attenuator setting,is due to the onset of amplitude limiting of the signal delivered to the AM detector due to front end overloading.However ,the severity of this effect does not actually reduce the Signal to Crosstalk ratio below that required to maintain a BER of better than 1 error in 100000 bits.

In a similar way to the data channel,the speech channel performance also deteriorates at the extremes of the RF input signal range.In order to assess this deterioration the data channel was modulated by a 1200 baud FFSK random data waveform with a maximum frquency deviation of 1.25 kHz and an AM depth of 25%.The maximum frequency deviation of the speech channel was also limited to 1.25 kHz.Recorded speech excerpts taken from Open University audio cassette programmes were used as the source of speech channel modulation and the speech quality,perceived audible effects of the crosstalk and listening effort required to understand the messages were assessed according to the following 5 point scales of subjective assessment:-

(i) Speech Quality:

5 Excellent
4 Good
3 Fair
2 Poor
1 Bad

(ii) Perceived Impairment:

5 Imperceptible
4 Perceptible but not annoying
3 Slightly annoying
2 Annoying
1 Very annoying

(iii) Listening Effort:

A Complete relaxation possible
(No effort required)
B Attention necessary
(No appreciable effort required)
C Moderate effort required
D Considerable effort required
E Message not understandable

The subjective assessment results obtained at various input Signal and Crosstalk levels are tabulated in Table 6.1.It can be seen from the tabulated results that provided the input signal level is higher than 1μV,the performance of the speech channel can be considered to be very satisfactory.Even at the 1μV level,which can occur in pockets of a coverage area,the performance can be considered satisfactory apart from the annoying effects of the crosstalk caused by the presence of the data signal modulating the other baseband channel.

6.4 Concluding Remarks.

To summarise,the practical investigations undertaken to assess the operational performance of a laboratoty prototype receiver capable of demultiplexing a QNBFAM RF transmission,have confirmed that the sensitivity of the system and its overall dynamic range of operation are simply matters for conventional receiver design techniques to accomodate.However,the major cause for concern is that the adjacent channel performance of the system falls somewhat short of operational system requirements because of the need for the linear phase IF filter characteristic.As yet there would appear to be no commercially available crystal filters which can meet both the amplitude and phase response requirements demanded for a practical QNBFAM system.The results presented in this Chapter have been obtained under controlled laboratory conditions and with constant RF input signal levels applied to the receiver front end.However,for the system to be considered suitable for a mobile radio environment it must be able to function effectively even when subjected to the rather severe channel impairments associated with such an operational environment.The following Chapter presents results obtained when the QNBFAM receiver is operated with a radio signal subjected to fading.

Footnote:

The subjective assessment of the speech channel performance was undertaken by three independent groups of listeners. The composition of the groups comprising final year Electronic Engineering undergraduates and members of the academic staff with a total sample size of thirty listeners. The results obtained from the three separate groups proved to be consistent and in general agreement in terms of assessing subjectively the three performance parameters "Speech Quality", "Perceived Impairment" and "Listening Effort" as summarised in Table 6.1.

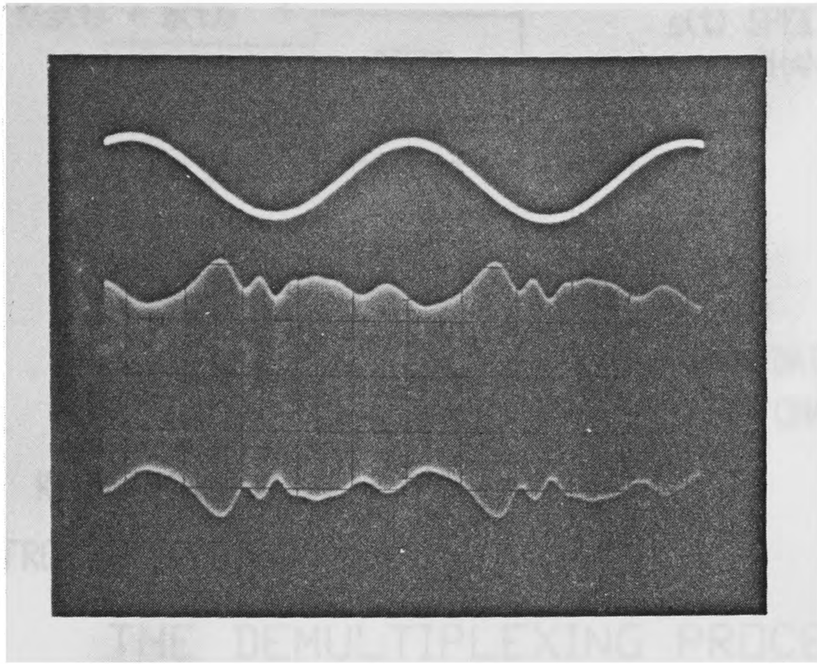


Figure 6.1 : Residual AM produced by Group Delay Distortion

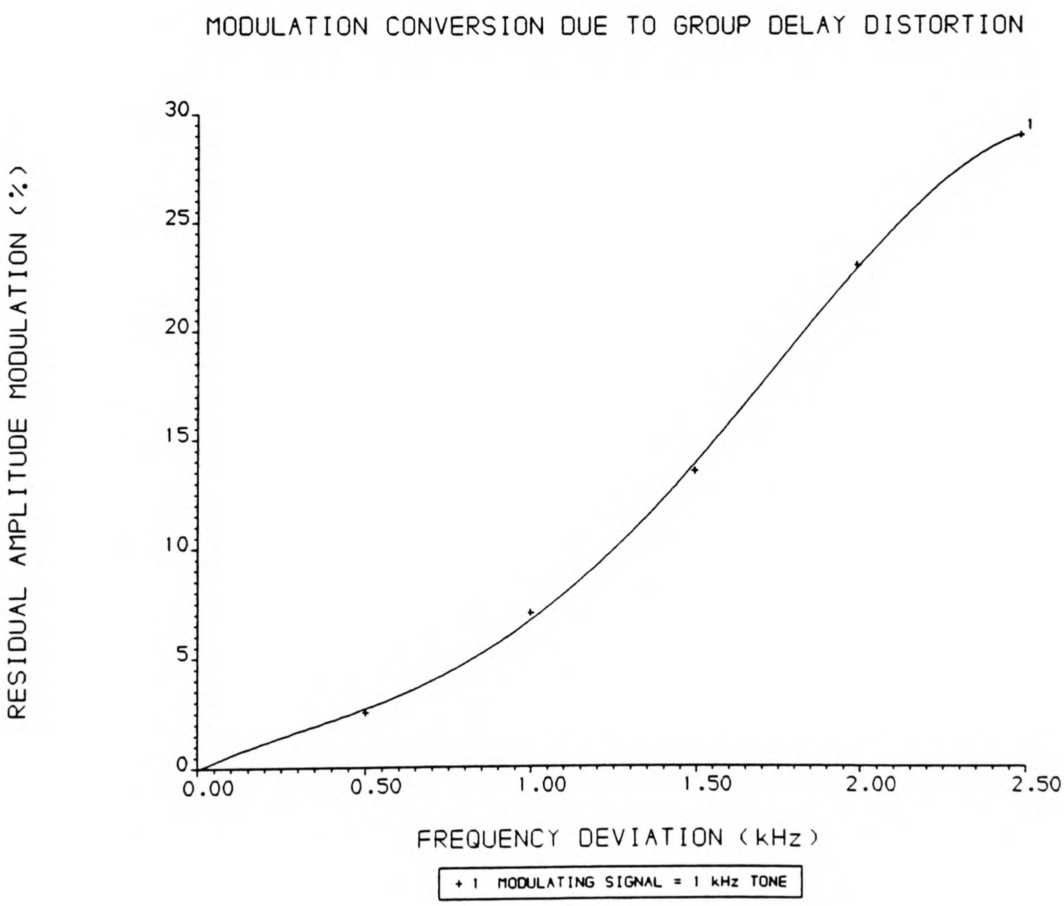


Figure 6.2 : Residual AM vs Frequency Deviation

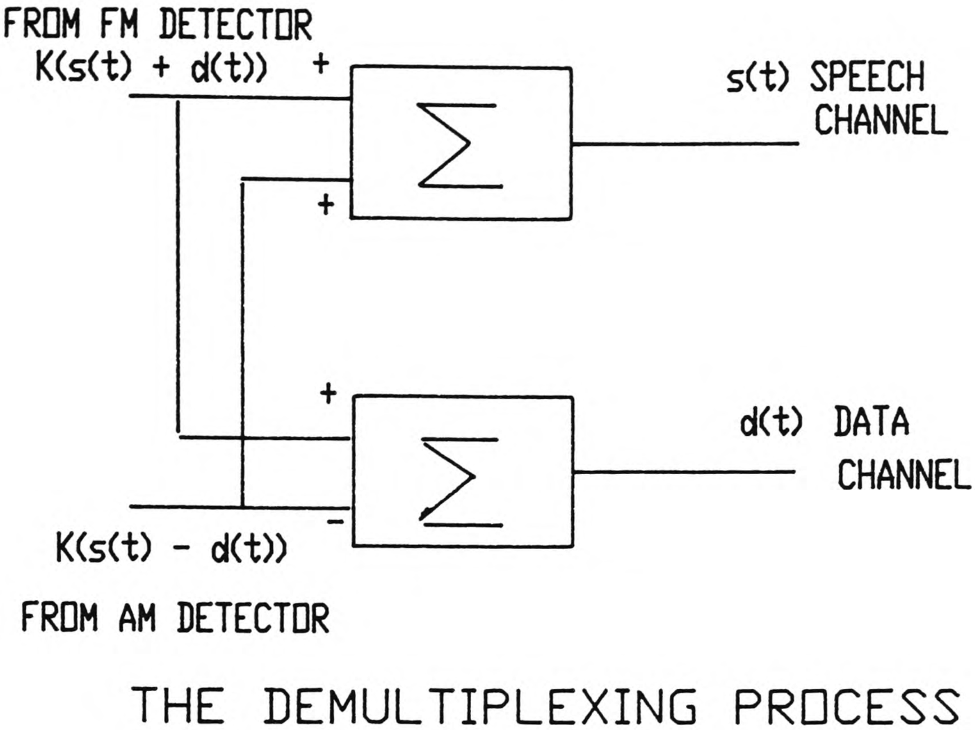


Figure 6.3 : Decoding the Speech and Data Channels

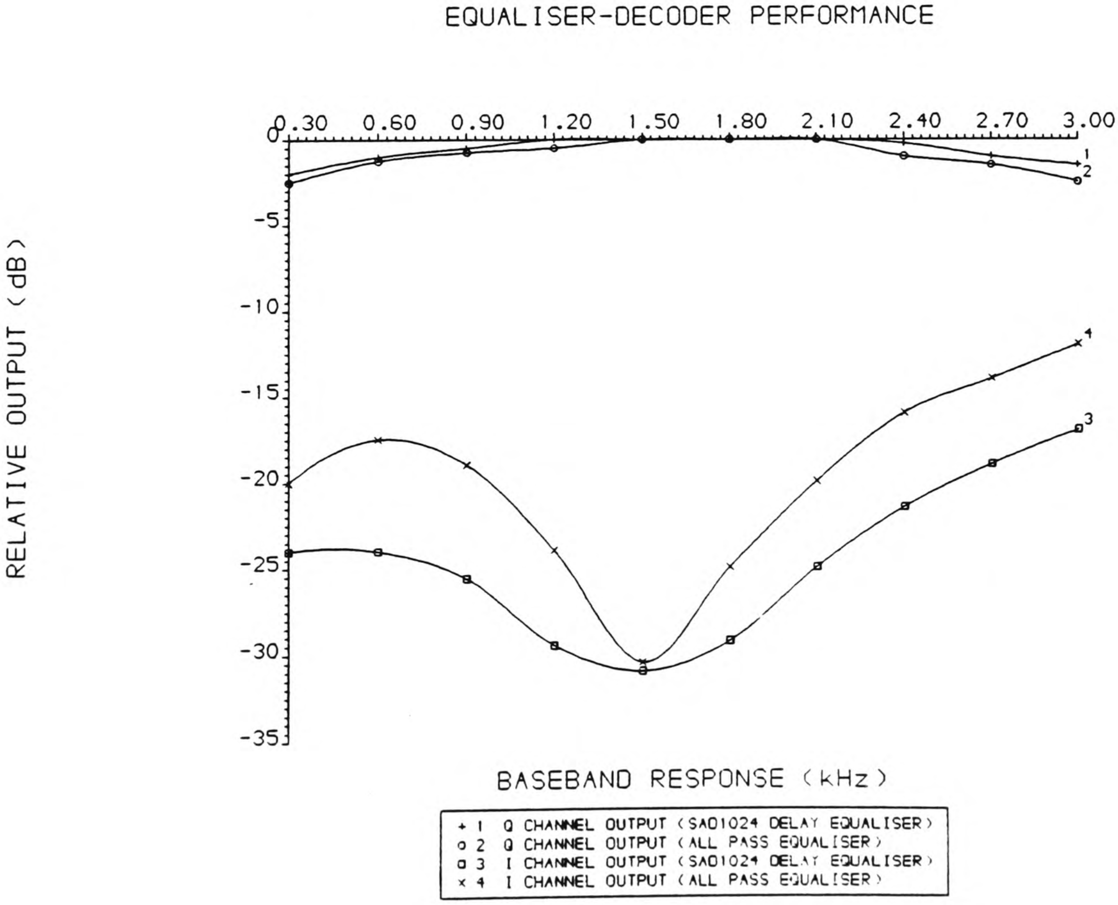
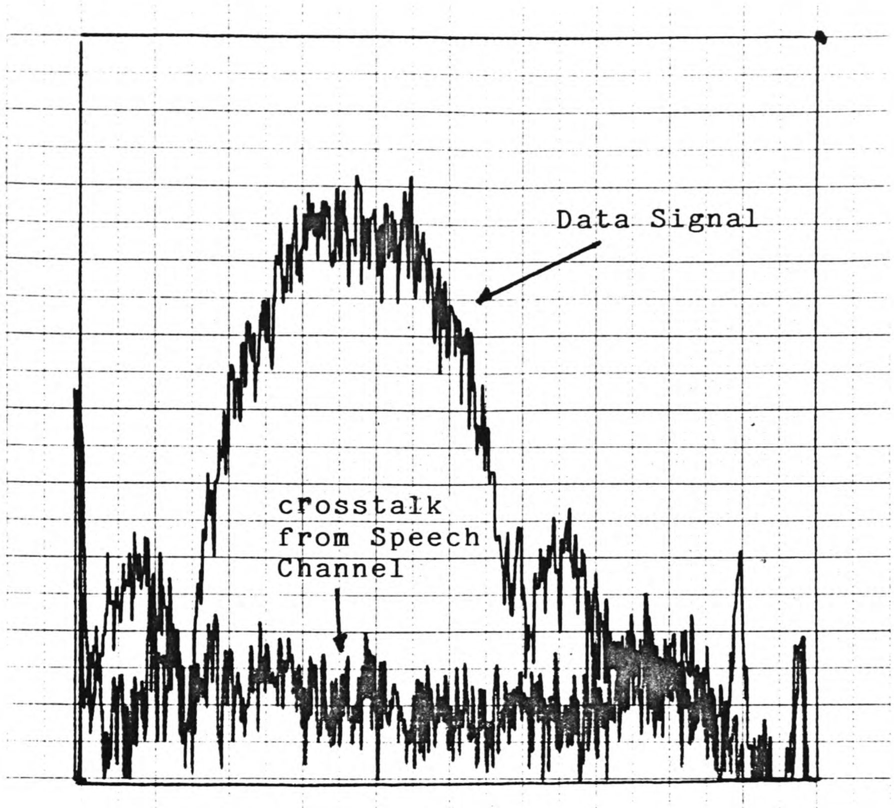


Figure 6.4 : Operational Performance of the Channel Decoder

(a) Data Channel



<u>SPECTRUM ANALYSER SETTINGS</u>:-

Reference Level = 10dBm : 5 dB/Div
Frequency Span = 4kHz

(b) Speech Channel

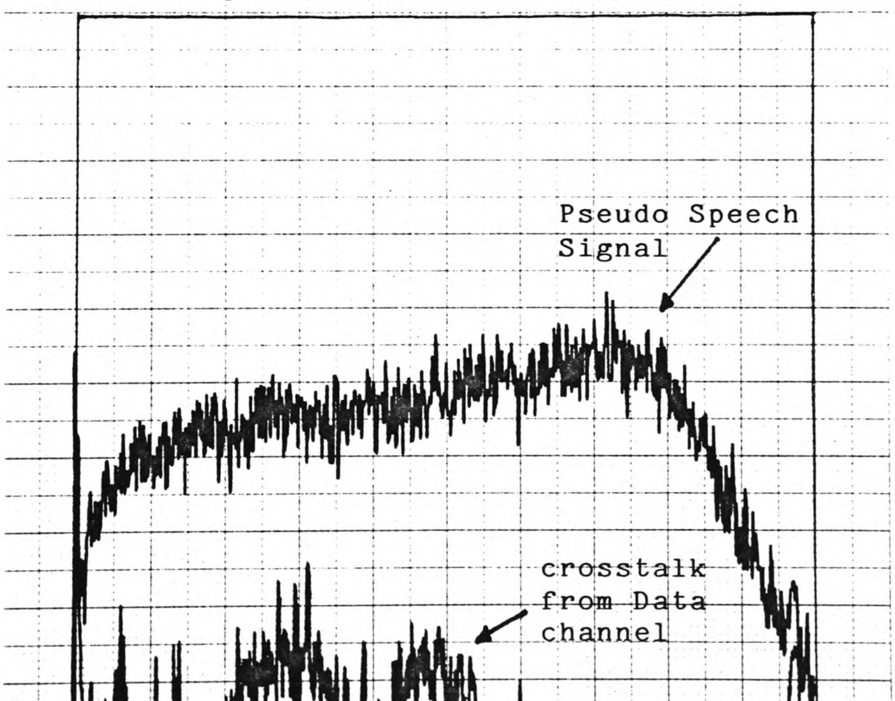


Figure 6.5 : Signal to Crosstalk Performance of the Speech and Data Channels

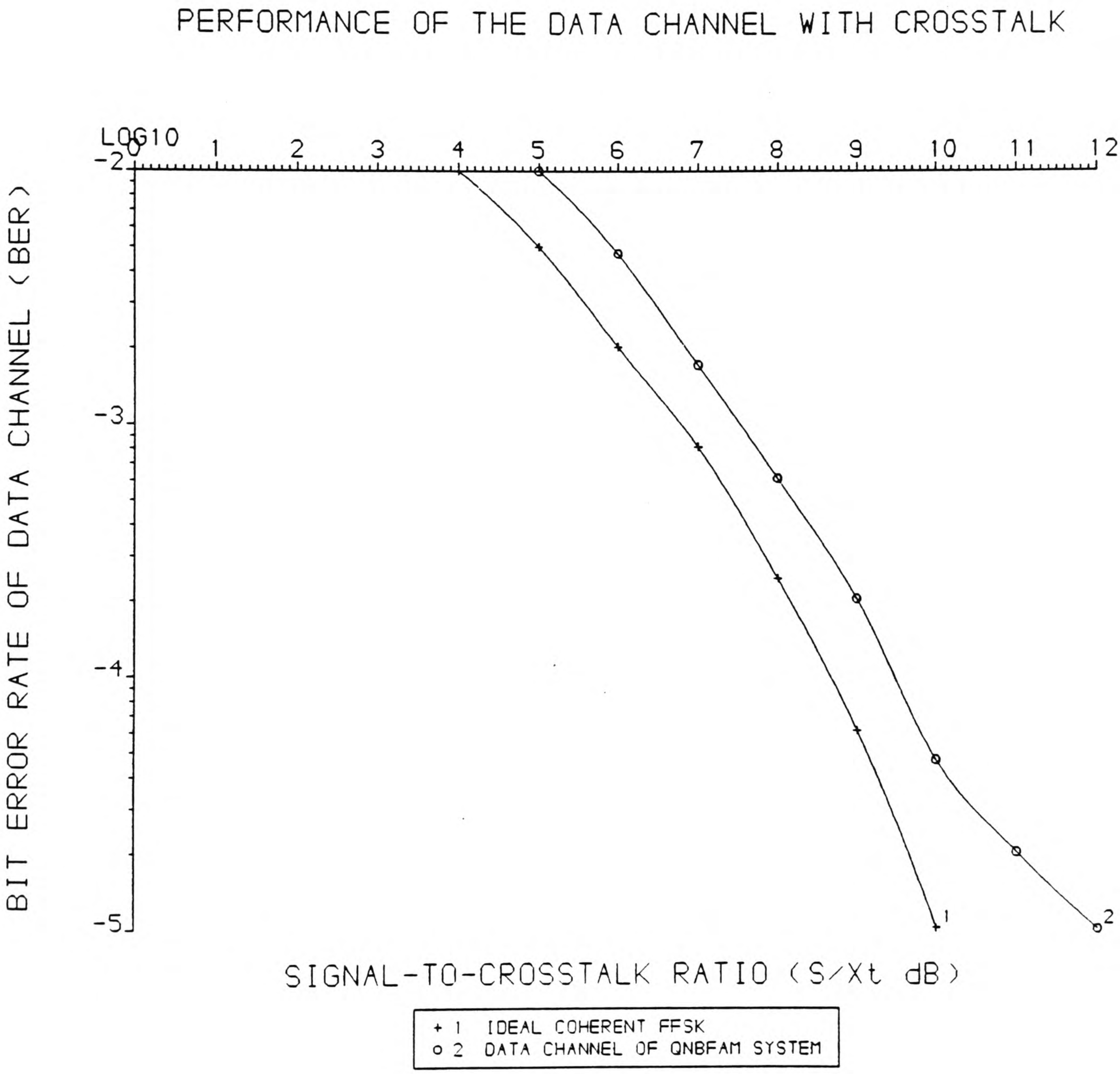


Figure 6.6 : Bit Error Rate as a function of Signal-to-Crosstalk Ratio (dB)

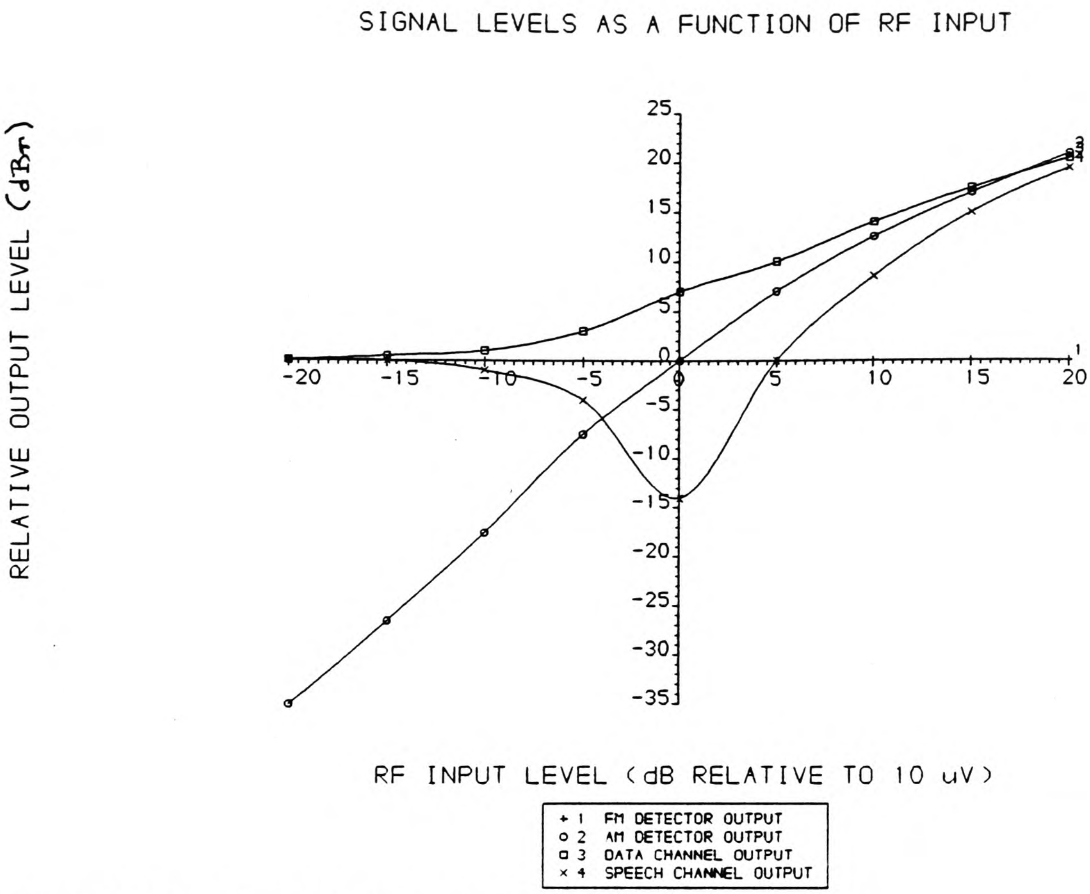


Figure 6.7 : Signal Levels as a function of RF input

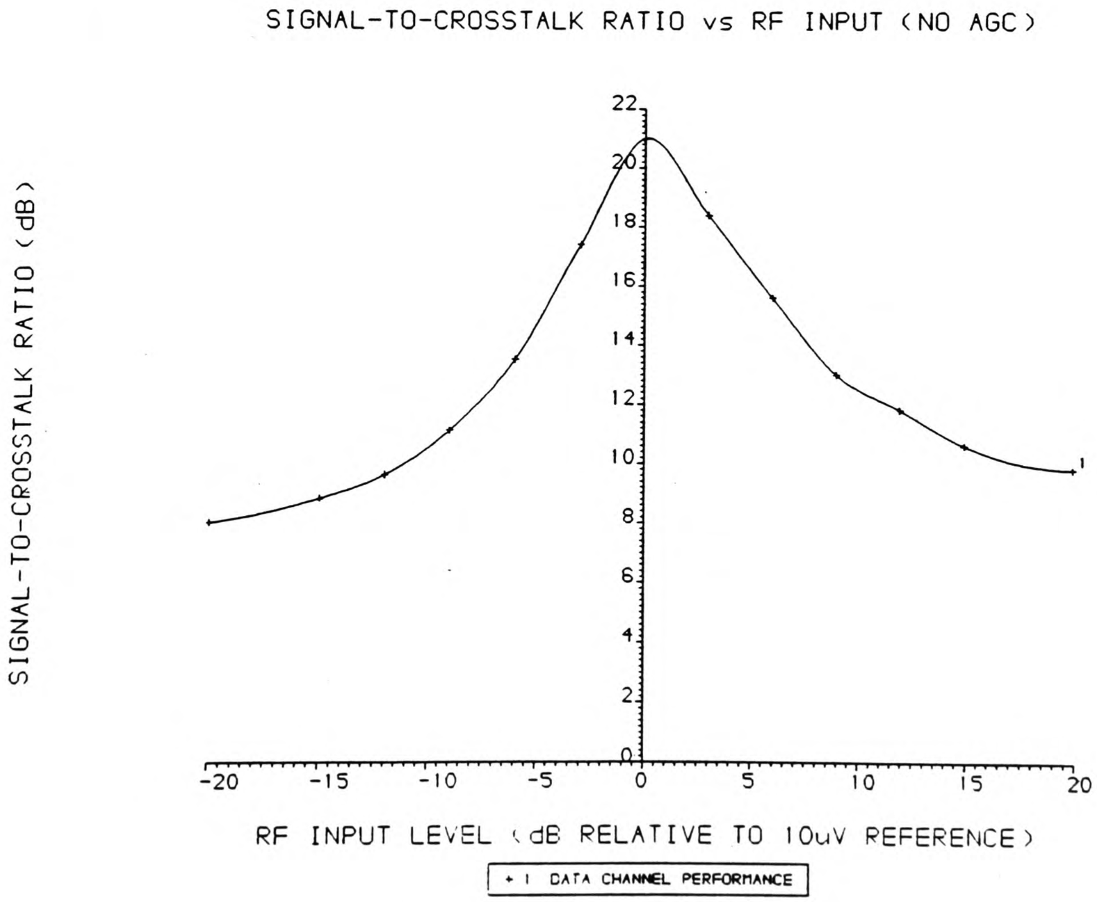
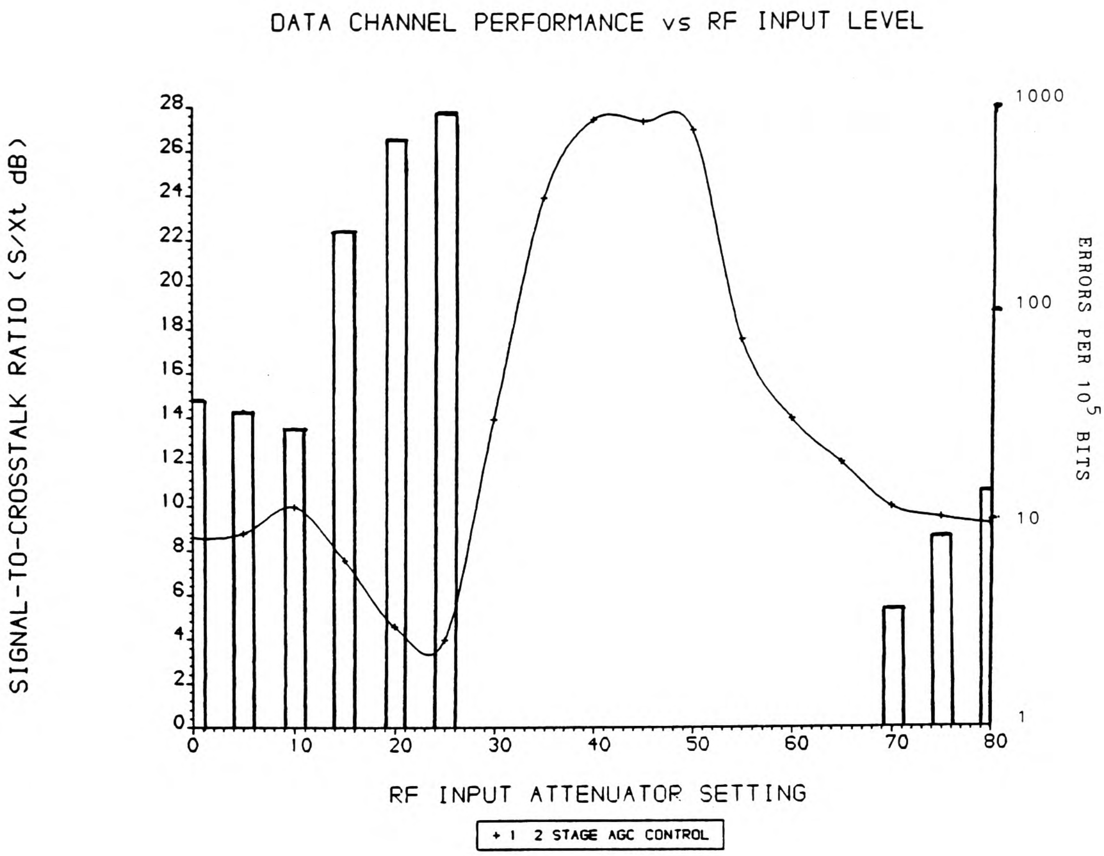


Figure 6.8 : Signal-to-Crosstalk without AGC

(a)



(b)

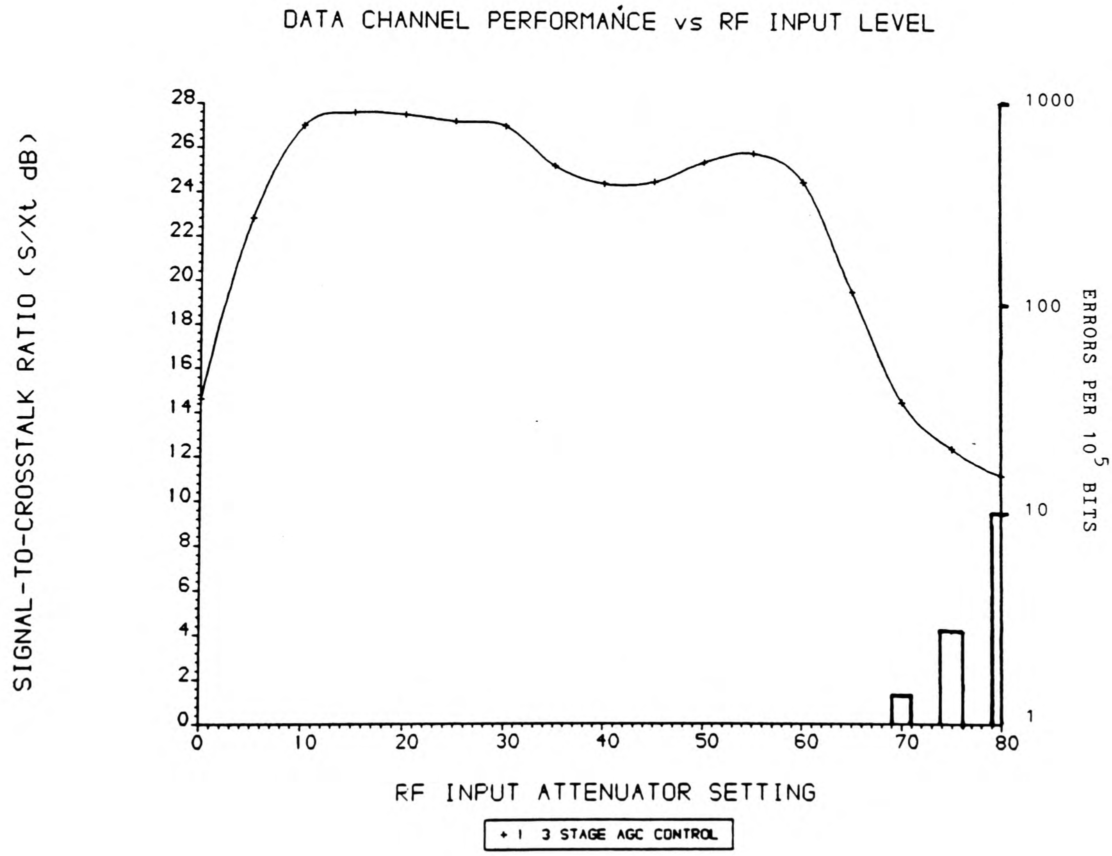
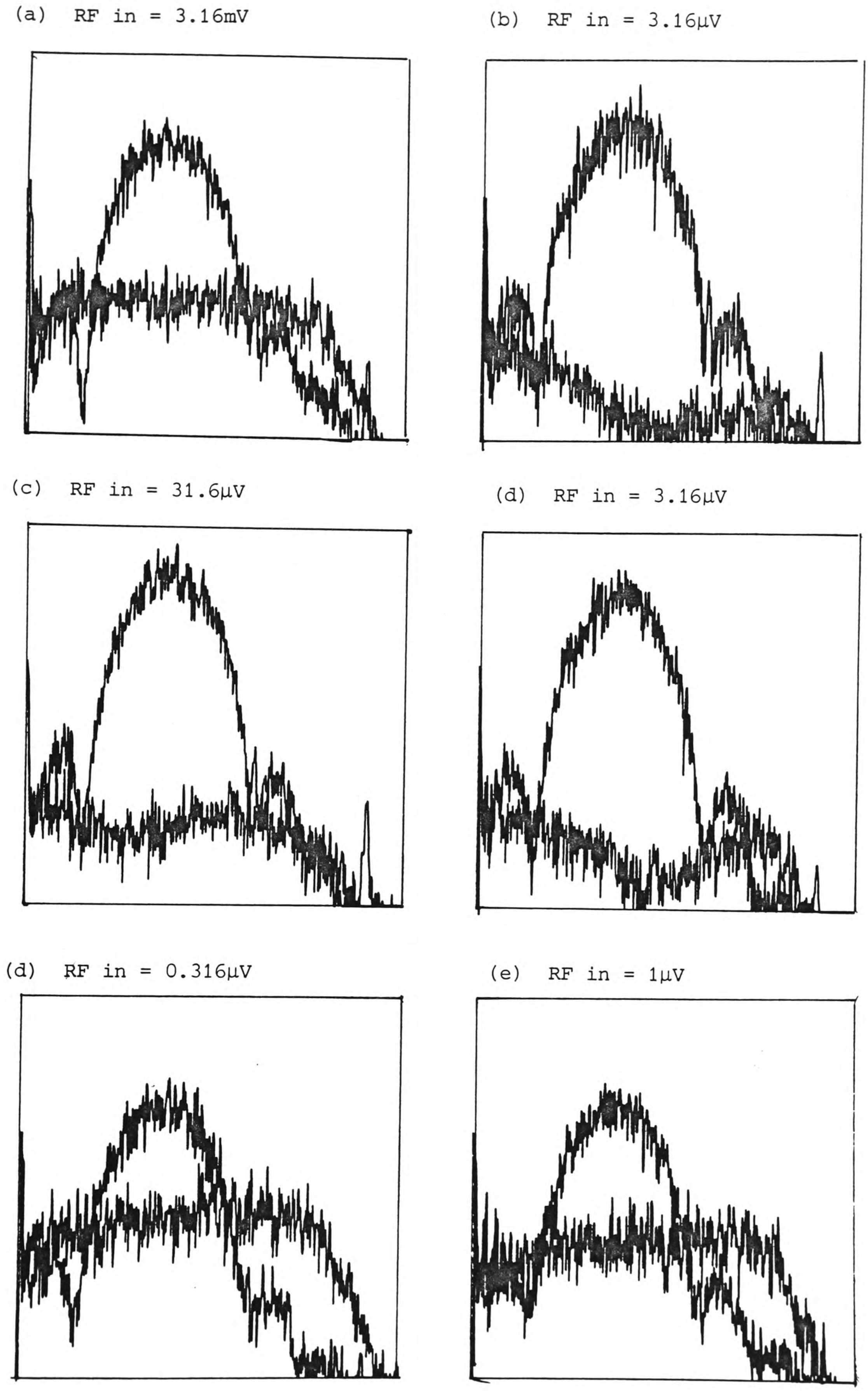


Figure 6.9 : Data Channel Performance - Error Rate Profiles



Spectrum Analyser Settings : (i) Reference Level = 5dBm
(ii) Amplitude Range = 50dB
(iii) Frequency Span = 4KHz

Figure 6.10 : Data Channel Performance

RF input signal level microvolt	Signal/ cross talk ratio(dB)	Speech Quality	Perceived audible effects of cross-talk	Listening effort req´d to understand message
3000	15	3	3	C
1000	26	4	4	A
300	28	4	4	A
100	27	4	4	A
30	24	4	4	A
10	25	4	4	A
3	24	4	4	A
1	15	3	2	C
0.3	9	2	1	D

Table 6.1 : Subjective Assessment of Speech Quality

CHAPTER 7

PERFORMANCE OF THE QNBFAM SYSTEM UNDER FADING CONDITIONS

7.1 The Mobile Radio Signal Environment.

In general,a mobile radio receiver does not have a direct line of sight to its base station transmitter.The received signal is normally the net resultant of many signals that reach the mobile via multiple paths,largely by way of scattering or reflections from and diffraction around buildings and obstructions within the locality of the mobile.Therefore,the RF signal received by a mobile at any point in the coverage area consists of a number of waves whose amplitudes,phases and angles of arrival are random.These multipath radio waves combine to set up a spatially fluctuating standing wave field.It is the movement of the mobile through such a field that causes the receiver to sense a time related fading signal.Fades of a depth less than 20dB are frequent,however,deeper fades in excess of 30dB although less frequent are not uncommon.Such rapid fading,experienced in an urban environment due to multipath propagation,is usually observed over short distances of half a wavelength or so.Figure 7.1 represents a typical plot of the received signal strength for a VHF carrier received in an urban environment.

Several multipath propagation models have been suggested in order to explain the statistical characteristics of the received fields and the associated fast fading signal envelope phenomena of the radio signals received by a mobile receiver (32,33,34,35).Evidence suggests that the scattering model,proposed by Clark (33),is the more appropriate and general model to explain the mechanisms leading to the signal fading characteristics.(A summary of important results is presented in Appendix G).This model,based on the assumption that the mobile received signal is of the scattered type with each component being independent,randomly phased and of random arrival angle,leads to the conclusion that the probability density function of the envelope is Rayleigh distributed as illustrated in Figure 7.2.

Also,whenever the receiver or transmitter is in motion,the received RF signal experiences a "Doppler" shift,the frequency shift being related to the cosine of the spatial angle between the direction of arrival of the wave and the direction of motion of the vehicle as represented by Figure 7.3 and quantified in Equation 7.1.

$$f_d = \frac{v}{\lambda} \cos \alpha_\eta \qquad 7.1$$

Where: λ is the carrier wavelength (m)

v is the vehicle velocity (m/s)

α_η is the angle of arrival for the Nth wavefront

It can be seen that the waves arriving from points situated ahead of the vehicle experience a positive Doppler shift,with a maximum value of v/λ ,and waves from those behind the vehicle do so with a negative Doppler shift.The net effect of the Doppler shift is to broaden the power spectrum of the received RF signal,the actual received spectrum being dependent on the antenna pattern being used (33).However,for an omnidirectional antenna,which is the most common type,the theoretical spectral density of the complex envelope of the received fading RF signal is given by Equation 7.2 and shown schematically in Figure 7.5a.

$$s(f) = \frac{E^2}{2\pi f d} \{1 - (f/_{fd})^2\}^{-\frac{1}{2}} \qquad f \leq fd$$

$$= 0 \qquad f > fd \qquad 7.2$$

Where: E is the rms value of the signal envelope (V).

$fd = v/_{\lambda}$ is the Doppler shift (Hz).

v is the vehicle velocity (m/s).

λ is the carrier wavelength (m)

The characteristics of a multipath radio environment will inevitably change as the mobile moves from one location to another,with different terrain and overall surroundings in terms of building features around the mobile influencing the local phenomenon.As a result,the rapid fades previously described are superimposed on a signal of slowly variable level.It has been suggested that the local mean of the received signal exhibits an approximate lognormal distribution.Measurements by various researchers appear to support this suggestion (36).However,it is the so called Rayleigh fades that places the most severe limits on the quality of both voice and digital transmission at VHF.

Although significant departures from Rayleigh are not unusual,the departures have a much smaller fading range than Rayleigh distributed fades.The corresponding time delays associated with multipath propagation have been directly measured and these have shown that signals due to nearby scatterers,with excess delays 0 to 1μs are clearly Rayleigh distributed while longer echoes with as much as 7 to 9μs delay are non-Rayleigh.The coherence bandwidth which is related to the spread in these delays,and is defined as the bandwidth within which fading has 0.9 or greater correlation,is greater than 40 kHz in urban areas and as large as 250 kHz in suburban environments (37).In either case,a modulated RF signal occupying a bandwidth of 40 kHz or less,is nonselective,that is the fading does not vary with frequency over a typical narrowband VHF channel.

7.2 Effects of Fading on the Transmission of Speech and Data.

Clearly,the presence of fast Rayleigh fading will degrade the operational performance of a mobile radio system.The statistics of the fades are such that 10% of the time the signal will be 10dB below its local mean,1% of the time 20dB below the mean (etc),where "local mean" in this context refers to the mean signal power received (37).The damage that fades can inflict on voice quality received in a 10dB signal to noise environment can be quite severe.However,the amplitude of the fades is not the only statistic of interest,the rate of fades and their duration below a given level is also significant (Appendix G).Since fades are caused by the vehicle motion through the multipath interference pattern both fading rate and fade duration depend on vehicle velocity.As a vehicle travels through a fading signal pattern,interruptions of the voice modulation will be heard as bursts of noise during the interruptions.These interruptions have a different subjective effect which is related to the vehicle speed and the received wavelength.For a mobile system operating with a VHF carrier frequency and travelling at speeds less than 20 miles per hour,the rate of speech interruptions is not as important as their duration.The effect is much like the swishing sound heard when driving slowly with the window open close to parked cars or closely spaced trees.At much higher speeds,the interruptions are heard as a series of "pops" or "clicks".

The consequences of these interruptions caused by fading on the transmission of a stream of digital data can be equally severe since actual blocks of data can be wiped out as the signal fades into the noise.For example,a VHF mobile operating at 200 MHz and travelling at 30 mph can produce an average fade duration below -10dB of the order of 15ms which corresponds to the time taken to transmit 18 bits of data at 1200 bits/sec.Consequently,both bit and word error rates are dominated by burst errors that occur when fades carry the signal down into the noise.For example,with a 15dB input signal to noise ratio,the signal is effectively "in the noise" 3% of the time and whilst in this condition,the error rate approaches one error in every two bits transmitted.Therefore,a bit error rate of 0.015 at 15dB signal to noise ratio can be anticipated solely on the basis of burst errors during fades,with the frequency and length of these bursts dependent on the rate and duration of fades.The bit error rate is also adversely affected by ignition noise interference which is a further channel impairment that cannot be ignored when considering modulated data transmission over mobile radio systems.

The virtual certainty that bits will be lost in any practical mobile radio environment imposes constraints on the types of coding strategies that must be employed to ensure that messages are received successfully and that false messages can be detected and rejected.One coding strategy specifically developed to overcome the loss of information bits during error bursts is the interlace technique (38).Bit interleaving can be used to improve performance when more than a single code word is transmitted.It is a method of effectively dispersing the errors that occur in clusters when the received signal fades,and which are likely to exceed the correcting capability of a code.Before a message is transmitted the bits from several code words are interleaved.Then,when an error burst occurs,the errors will be shared among the interleaved code words and a less powerful code is required to correct them.

7.3 A Real-Time Fading Simulator.

The previous discussion highlights some of the operational problems associated with the propagation of radio signals over a mobile system whose channel characteristics can be considered to be rather severe and hostile to the transmission of speech and data signals.In the light of these operational difficulties it would be rather naive to assume that a QNBFAM multiplexed transmission system would not be seriously affected by the unfavourable channel conditions.In particular,the presence of fading will not only degrade the quality of the recovered speech and data baseband signals but will inevitably adversely affect the demultiplexing process which is employed to separate the two independent channels during a simultaneous transmission.The reciprocal enhancement/cancellation of wanted/unwanted baseband channels which the demultiplexer relies upon to separate the message channels will be degraded leading to an increase in the interaction between the same which will be manifest as "bursts of crosstalk" during the more severe fades.Although the presence of a fast acting AGC system will respond in a way to minimise this interaction, nevertheless,the Rayleigh fading phenomena will cause a deterioration in the performance of the QNBFAM system and could render its application unsuitable to mobile radio systems operating in urban areas.

In order to assess the performance of the QNBFAM scheme under fading channels conditions and obtain meaningful results,with regard to the potential worsening of the crosstalk problem, a "real-time" fading simulator was designed and built. The simulator designed generates Rayleigh distributed fading at variable Doppler rates up to 21 Hz, which is equivalent to a 200 MHz transmission being received by a mobile travelling at a speed of 70 mph. The fading channel simulator employs the well established technique of splitting the RF signal into two quadrature components which are then independently modulated by low frequency noise having a Gaussian amplitude distribution (39). When the two quadrature components are combined by addition the resulting signal will have a Rayleigh amplitude distribution and a uniformly distributed random phase (Appendix G). A schematic of the simulator identifying all relevant functional blocks is illustrated in Figure 7.4.

As with other real time fading simulator designs (40), pseudo random binary sequence (PRBS) generators are employed as the noise sources. The two PRBS generators used are conventional in design employing three 4015 dual 4 bit CMOS shift registers with Exclusive-OR feedback. In order to avoid the all zero code circulating indefinitely a monostable was also connected in the feedback loop via an OR gate such that both registers can be reset at any time. The individual generators are of lengths $K1=2^{19}-1$ and $K2=2^{24}-1$, with feedback connections taken from the following register outputs (19,18,17,12) for K1 and (24,23,21,20) for K2 to ensure maximal bit sequences before repetition occurs (41,42). This means that with a clock frequency of 16kHz the repeat times are 33 and 1048 seconds respectively. Therefore, the output spectrum generated consists of noise extending from the repeat frequency of the sequences up to the clock frequency and beyond with a "sinc(x)" amplitude distribution and can be considered flat to within +/- 0.1 dB up to approximately 12% of the clock frequency.This means that a low pass filter with an upper cut-off frequency less than 2 kHz can convert the unfiltered PRBS digital output into a bandlimited analogue noise voltage which is a good approximation to a source of "White" or Gaussian noise (43).

The simulation of the fading spectrum appropriate to mobile radio is obtained by properly shaping the spectrum of the noise sources by low pass filters.As previously discussed,the resultant spectrum depends on the antenna pattern being used and for an omnidirectional antenna,the spectral density of the complex envelope of the received signal is given in Equation 7.4.Low pass filters having the desired frequency response were designed using five pole filters with a normalised transfer function given by Equation 7.5.

$$H(s) = \frac{1}{(0.897s^2 + 0.31s + 1)(1.543s^2 + 0.841s + 1)(1.944s + 1)} \qquad 7.5$$

The theoretical and practical filter responses are compared in Figure 7.5.

In order to accomodate variable fading rates corresponding to different vehicle speeds the filter responses have to be modified accordingly by some form of tuning.The solution adopted for each noise filter,was to employ two MF10 second order switched capacitor filters (44) in cascade with a first order active low pass section. A switched capacitor filter has the distinct advantage for this particular application since a range of fading rates can be obtained by simply varying the filter clock frequency. In this way the switched capacitor filter characteristics can be tuned without changing the shape of the frequency response which is defined by correct choice of external fixed components. On the other hand, the required cut-off frequency of the first order filter section has to be achieved by setting a preset variable resistor to the correct value in order to ensure the overall response is appropriate to the fading rate to be simulated.

In order to make the fading simulator more adaptable and suitable for a wider range of RF applications in the future, the simulator was designed to operate at the standard IF frequency of 10.7 MHz. This also simplified the design of the RF section significantly since the centre frequency of operation is constant irrespective of the input RF carrier frequency. The incoming signal to be subjected to fading is therefore derived from the IF section of the QNBFAM receiver prior to the second mixer stage and the linear gain controlled IF section which drives the AM detector. This signal is then split into two phase quadrature components by a similar network to that employed in the QNBFAM transmitter and described in section 5.5.2. These phase quadrature signals are then multiplied by the independent Gaussian noise sources by two SL640 integrated circuit double balanced modulators (45).

The required Rayleigh fading signal is finally obtained by adding together the modulated quadrature phased components in a summing network comprising a Common Base amplifier configuration also described in Chapter 5 (Section 5.2.4).This resultant signal is then returned to the QNBFAM receiver after suitable amplitude scaling and buffering.

7.4 Evaluation of the Simulator Performance.

The functionality and operation of the configuration adopted for the real time fading simulator has been evaluated by several investigators (39,40).It has been confirmed that its performance accurately simulates the Rayleigh distributed fading which is characteristic of the mobile radio environment.However,measurements were still undertaken to confirm the statistical characteristics of the important waveforms generated by the simulator (46).

The amplitude distributions of the filtered noise waveforms and the envelope of the simulator output signal were evaluated with the aid of an IBM personal computer running a package known as ILS-PC1 (47).ILS is an "Interactive Laboratory System" which is based on user level software that enables interactive digital signal processing to be achieved in a very straightforward manner.Any waveform to be analysed is initially converted to a digital representation via a 12 bit Analogue to Digital converter (ADC).Such digital signals are then stored in ILS "sampled data" files as 16 bit binary values (-32768 to +32767).An ILS sampled data file is made up of a series of these values and maybe very large limited only by the memory available in the computer.

ILS offers many signal processing functions such as Signal Display and Editing,Data Manipulation,Spectral Analysis and Digital Filtering through a wide choice of options available with its internal set of commands.However,another useful feature of the ILS package is that the sampled data files can be accessed,independent of ILS by a user defined BASIC or other high level program available in the system.This means that data from the sampled data file can be transferred into an array,within a BASIC program,and processed or analysed by a routine written specifically for this purpose,such as analysing the first order statistics of a random signal.Also,if required,processed or modified data can be written back to an ILS data file.This useful facility extends the application area of ILS beyond the range determined by its already extensive command set and also allows the user to establish a personalised version of ILS (48).

The results of the analysis performed on the filtered noise sources and the envelope of the resultant output signal using ILS are shown in Figures 7.6a and 7.6b.As indicated by Figure 7.6a the noise sources were found to approximate a Gaussian distribution over a range of 3 standard deviations.The "closeness of fit" of the output envelope to a Rayleigh distribution is represented by Figure 7.6b.The Rayleigh variate R (Rayleigh cumulative distribution) and the measured variate $\hat{R}$ (measured cumulative disribution) are plotted with both variables normalised to the rms value of the measured variate.If the simulator was exactly Rayleigh distributed,all points would fall on the straight line.As can be seen, the output envelope is within 3dB over a range extending to 30dB below the rms point.

7.5 Effects of Fading on the QNBFAM System

As anticipated,the effects of Rayleigh fading inevitably degrades the performance available from a QNBFAM multiplex radio system.Presenting a QNBFAM receiver with a radio signal which has been subjected to fast fading affects the capability of the demultiplexing operation to separate the speech and data signals without mutual interference.Although the increase in interchannel crosstalk caused by fading reduces the quality of the speech channel it is quite surprising how intelligible the speech signal can still remain even though subjected to rapid bursts of crosstalk from the data channel during fades.The audible effect of this crosstalk interference from the data channel is,as expected,dependent upon the rate at which the fades actually occur,which in turn is related to the velocity of the mobile receiver.At simulated speeds less than 30 mph the audible interference produces a "fluttering" sound which is itself modulated by a higher frequency hissing noise due to the presence of 1200 baud random FFSK data in the digital channel.At higher simulated vehicle velocities,in excess of 40 mph,the audible disturbance is not dissimilar to the sound produced by galloping horses augmented,as was found at the lower fading rates,by the presence of a higher frequency interfering noise type signal.

The subjective assessment of the quality of the voice channel performance when exposed to bursts of data channel crosstalk during fades is summarised in Table 7.1.This assessment was obtained with an effective RF input signal of 10 μV (50 dB input attenuator setting) and the speech channel modulated with extracts involving passages of voice only pre-recorded material whilst the data channel was modulated by random FFSK data.Both baseband channels being modulated to the maximum frequency deviation of 1.25 kHz.The results confirm that the speech channel performance is quite robust and can be considered adequate in terms of message intelligibilty up to the simulated legal speed limit of 70 mph.

As with the speech channel,the data channel error performance is also degraded by the disturbing effects of bursts of interference from the speech channel when the received signal is subjected to rapid fades.To assess the extent of the combined effects of fast fading and increased interchannel interference on the data channel error rate performance relevant measurements were carried out.With an input signal strength set to the same level as with the speech channel assessment,the error rate performance was determined for a range of simulated vehicle speeds from 10 mph up to 70 mph.As with the voice channel assessment,the frequency deviation of both channels was restricted to 1.25 kHz.

The results of these measurements are presented in graphical form in Figure 7.7.The graphs of Figure 7.7 contrast the error rate performance obtained due to the fading phenomena acting alone,i.e.,with the speech channel modulation suppressed,to that due to the combined effects of fading and the resultant increased interchannel interference when the speech channel is modulated by bandlimited random noise.It can be seen that at low vehicle velocity simulations the error rate performance is worse when the crosstalk from the speech channel is present which should of course be anticipated.However,in both cases the errors increase as the simulated vehicle speed is increased with the difference in errors between the separate cases reducing to such an extent that eventually the errors due to the fading effects would appear to dominate the data channel error rate performance.The measurements were repeated on several occasions to authenticate the results obtained.These results would appear to suggest that standard techniques such as diversity,developed to combat the undesirable effects of fading,would also be of significant benefit to a QNBFAM implementation.Any reduction in the extent of received signal fading would obviously contribute to reducing the unwanted interaction between the baseband message channels which would otherwise be exaggerated by the extreme channel impairments associated with a mobile radio signal environment.

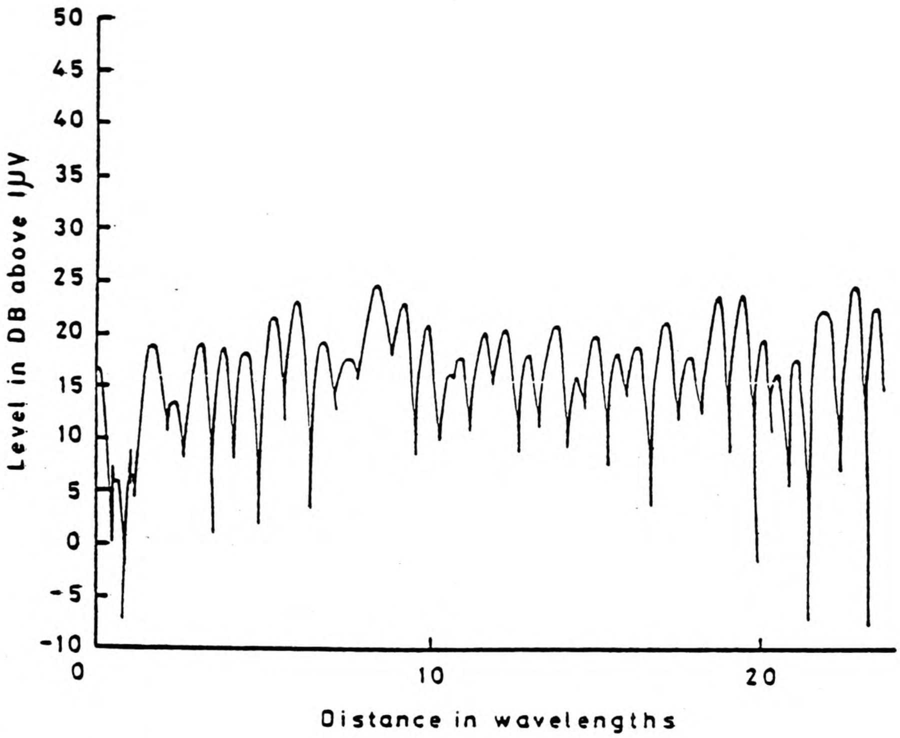


Figure 7.1 : A Typical Received Signal
(London, 168 MHz)

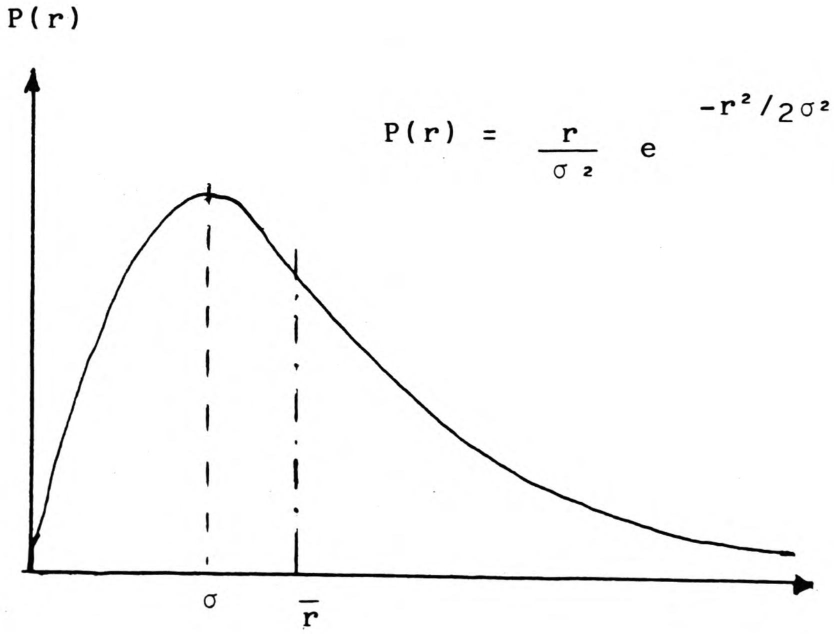
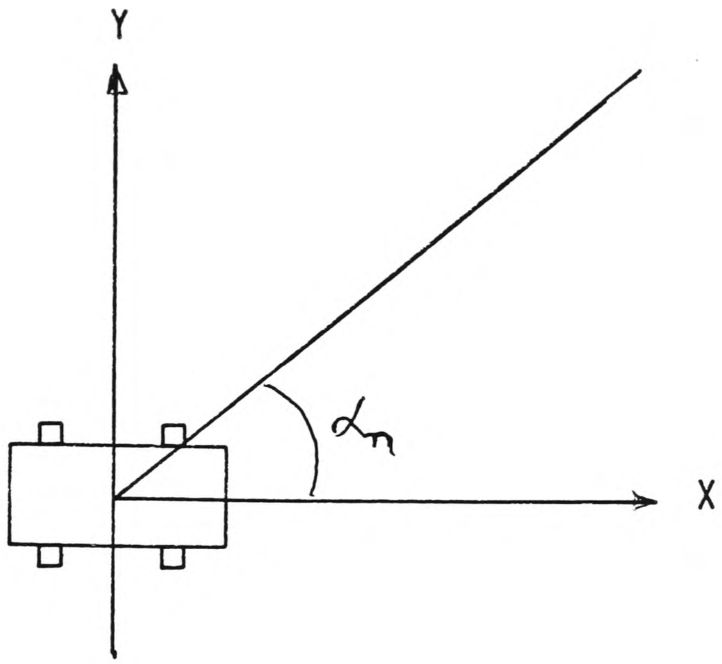


Figure 7.2 : Probability Distribution of the
Fading Envelope



DOPPLER FREQUENCY SHIFT

Figure 7.3

THE RAYLEIGH FADING SIMULATOR

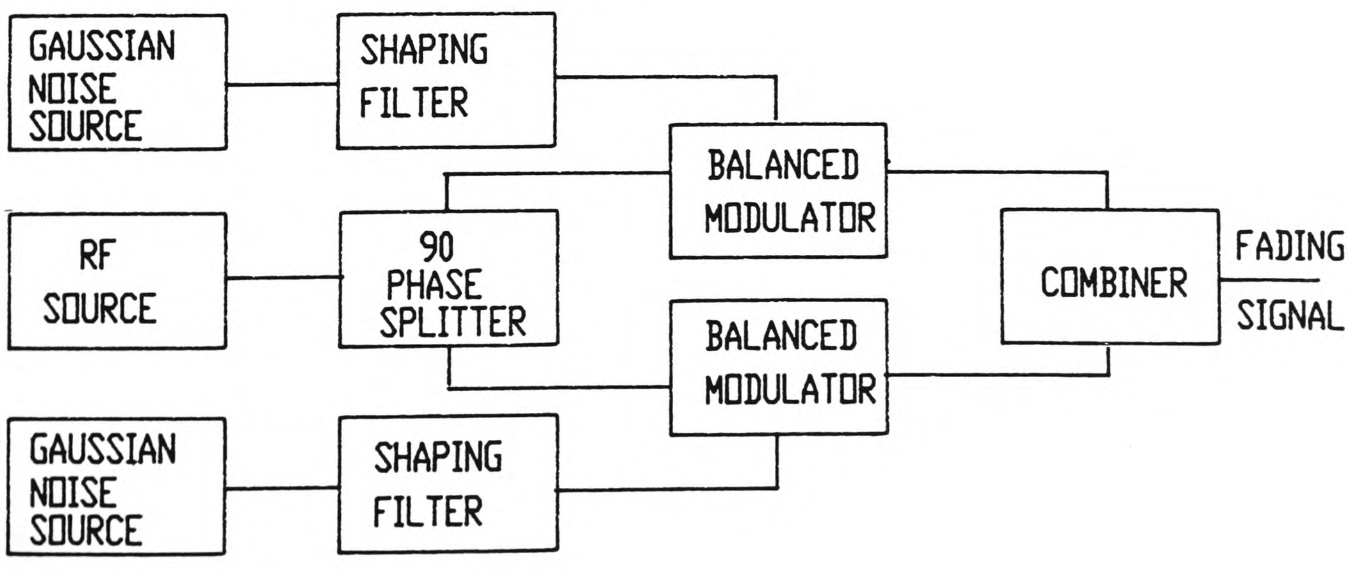
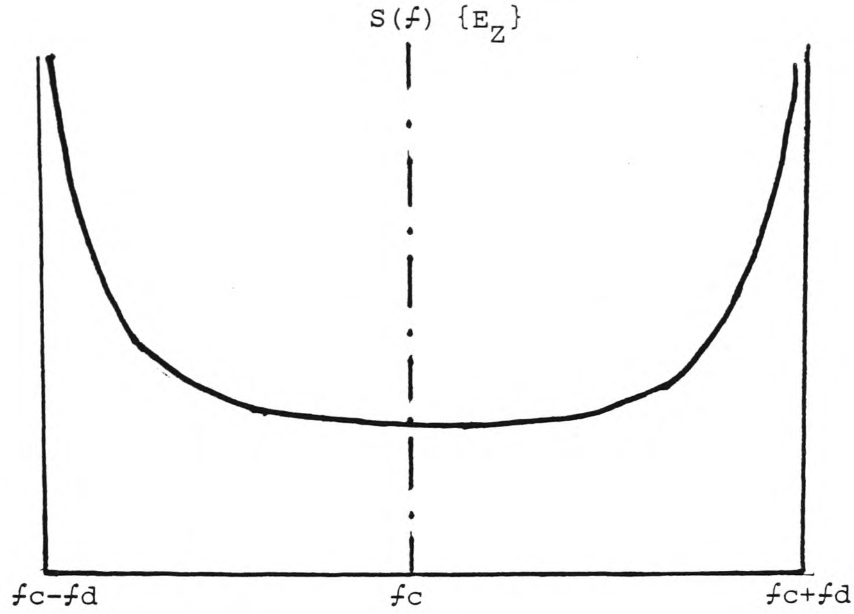


Figure 7.4

(a) RF Signal



(b) Simulator Low Pass Filter Response

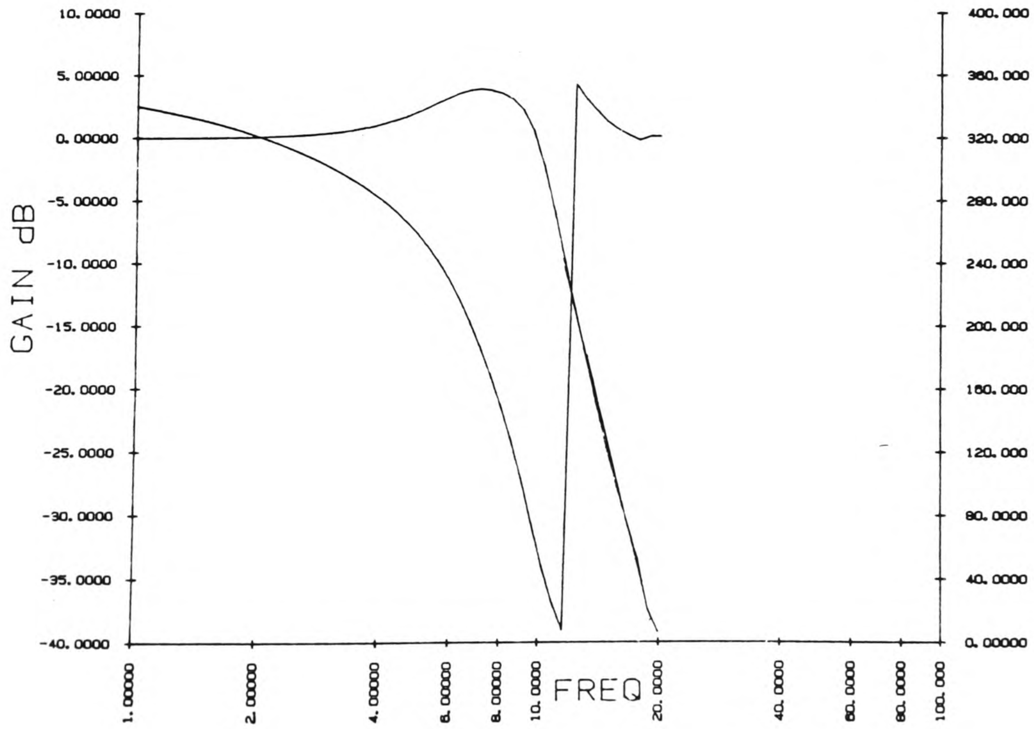


Figure 7.5 : Spectrum of the Fading RF Signal

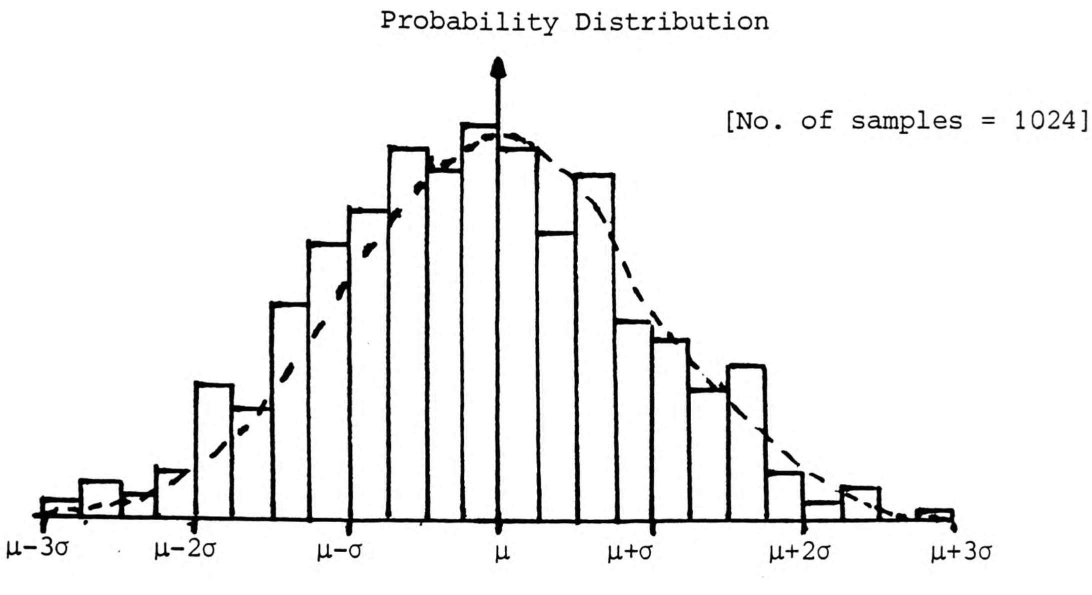


Figure 7.6(a) : Amplitude Distribution of the Filtered Noise

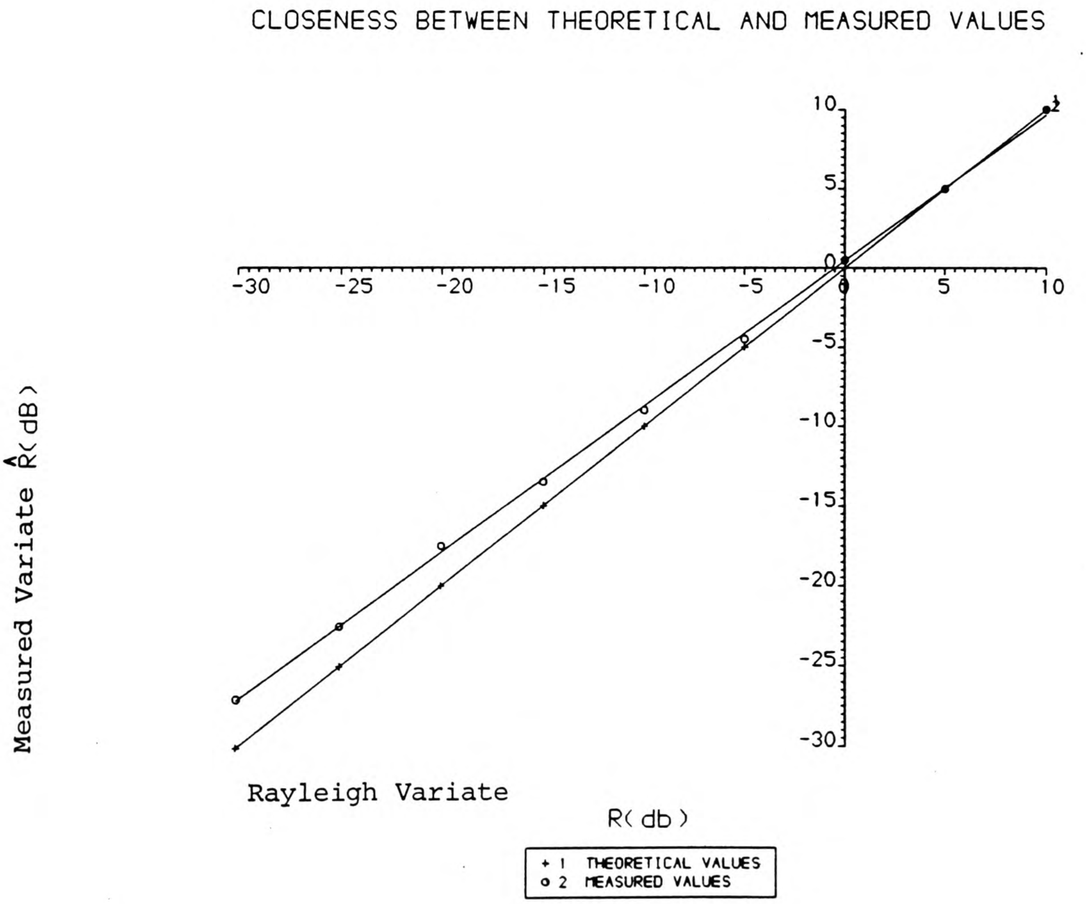


Figure 7.6(b) : Closeness of Fit to Rayleigh

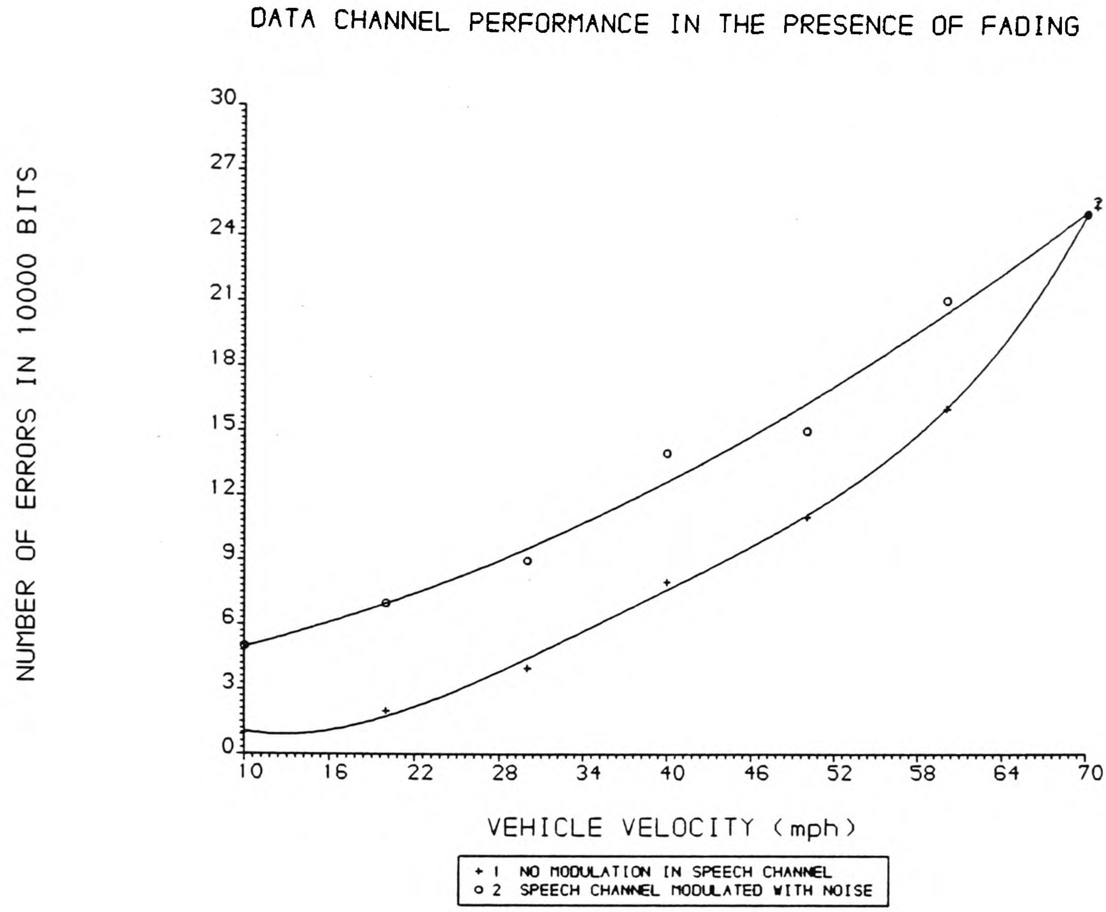


Figure 7.7 : Error Rate Performance at Different Vehicle Speeds

SIMULATED VELOCITY (mph)	SPEECH QUALITY	IMPAIRMENT	LISTENING EFFORT
10	4	4	A
20	4	4	A
30	4	4	A
40	4	3	B
50	4	3	B
60	3	2	C
70	3	1	C

Table 7.1 : Subjective Assessment of Speech Channel when subjected to Fading.

CHAPTER 8

POTENTIAL IMPROVEMENTS FOR THE QNBFAM SYSTEM

8.1 General Comments.

Although every effort was made to ensure that the results obtained,concerning the performance of the prototype QNBFAM multiplex system,were representative of an operational system,it must be emphasised that the results were derived under laboratory conditions.Such conditions must be considered controlled if not necessarily contrived and, arguably,can only approximate the conditions under which an operational QNBFAM system is expected to function. Consequently,the only way to truly assess the feasibilty of the proposed multiplex system is to undertake a series of extensive field trials.This would require a QNBFAM signal being transmitted from a VHF base station and being received by a suitably configured mobile receiver in both urban and rural environments.Such field trials,as described,are the only way to allow the speech and data channel performances to be realistically assessed and evaluated under actual channel conditions.However,it is fair to say that the results obtained lead one to have a reasonable level of confidence about the performance of an operational QNBFAM system.The results indicate that taking the research out of the controlled environment of the laboratory and exposing the system to "real" operational conditions would indeed be a worthwhile exercise.In this context real conditions do not only refer to the fast fading phenomena associated with a mobile radio signal environment but also other unfavourable channel conditions such as ignition noise and co-channel interference.

Another aspect to encourage further work to be undertaken is that for the majority of time the transmitted signal will be a conventional narrowband angle modulated signal.A wealth of experience has been gained with this type of modulation technique which has led to its adoption as the modulation standard for all new mobile radio systems for the recently released Band III spectra.It is envisaged that for the majority of the time a single mode of transmission will be employed.The integrated mode of operation which is very desirable and convenient,would only be required and used for a smaller percentage of the time.

In order to make the QNBFAM transmitter-receiver prototype equipment,developed to date, suitable for field trials to be undertaken a number of refinements must be made.In fact,nearly all the signal processing functions could be improved or enhanced and in certain cases due consideration must be given to ensuring the stability of operation over the temperature range the mobile equipment is to be used.For example,the ambient temperature range of a passenger compartment of a vehicle can range from as low as -20° C overnight,to as high as $+ 60^{\circ}$ C in an unventilated vehicle suffering from the "Greenhouse Effect".

8.2 Potential Modifications to Improve the QNBFAM Transmitter Prototype.

Although no particular problems have been identified or experienced with the transmitter configuration,in its laboratory prototype form,it must be remembered that it is currently operating with a very low level output. Therefore,in order to provide the level of power output required of a base station transmitter (typically 20 Watts) the final RF section of the QNBFAM transmitter will need to be completely re-designed.This re-design must also take into account the MPT 1326 specifications with regard to its operational performance in terms of frequency stability, maximum permissible frequency deviation and spurious outputs (etc).

The final RF section must also enable the composite angle modulated signal to be amplitude modulated by the low frequency information signal derived from the envelope of the QNBFAM signal at the output of the IF section of the transmitter.Traditional designs of high level AM modulators for mobile equipment have suffered from the disadvantages of high cost of implementation and poor efficiency.The ratio of RF power output to DC power input in previous designs has tended to be low because most of the power supplied to the final RF amplifier is AF power from the modulating amplifier. Consequently,the losses of DC to AF conversion are compounded with the losses involved in the final conversion to RF power.Low efficiency is undesirable particularly with battery operated equipment,however,it also increases equipment costs because of the need for larger heatsinks and active devices with higher power ratings.

With the advent of VMOS technology FET power devices are now available which can be used in the cost effective design of VHF mobile transmitter equipment.The particular features of VMOS power FET devices which make them an attractive solution for the application in question are high power gain,better non-linearity performance than bipolar devices and freedom from secondary breakdown and allied phenomena (49).Work is currently underway developing a high level VHF AM modulator for the QNBFAM transmitter prototype based on an original design published by Petrovic and Gosling (50).Very briefly,the proposed AM modulator is based on an N channel VMOS power FET (Siliconix DV1220S) and operates with "gate" modulation.In this mode of operation the final RF output is controlled by varying the bias applied to the gate terminal.Since the gate provides a high input impedance to the low frequency modulating signal, little AF power is needed for gate modulation.Also,the non-linearity problem associated with previous gate (or grid) modulator designs is effectively overcome by using feedback to control the modulators performance.The final output of the transmitter is demodulated using a precision AM detector and compared with the original modulating waveform.The difference between these two signals is then amplified and applied to the gate of the VMOS device in such a way as to form a negative feedback loop.Provided that the amplifier gain is sufficiently large,the RF envelope can be made to follow the modulating voltage very closely.

Another useful feature that could be incorporated into the transmitter configuration and enhance the overall operation of the QNBFAM multiplex system would be a "CTCSS" encoder.A "continuous tone controlled squelch system" would be particularly suited to the QNBFAM system in order to effectively disable the AM section of the receiver remotely when a single channel mode of operation is being used.With a single baseband channel modulating the transmitted carrier the AM section of the receiver is effectively redundant since the transmitted carrier is angle modulated and does not convey any envelope modulation (Section 5.2,Chapter 5).CTCSS encoders are now readily available in integrated circuit form and could be easily incorporated into the transmitter prototype to effect a sub-audio tone squelch system for remotely muting the AM section of the QNBFAM receiver section.Such a device is the FX315 CTCSS encoder available from Consumer Microcircuits (51).Thirty eight separate tone frequencies which range from 67 Hz to 250.3 Hz (52) can be derived from an input reference frequency.An on-chip inverter is also provided to drive an external crystal circuit.Tone selection is achieved by a six bit logic code and permits a low distortion sinusoidal tone to be generated at one of 38 possible sub-audio frequencies.

8.3 Potential Modifications to Improve the QNBFAM Receiver Prototype.

Although the receiver prototype was designed and constructed using circuitry taken directly from existing commercial equipment,as well as dedicated integrated circuits to achieve other essential receiver functions,it nevertheless can be significantly improved.In particular,the receiver sensitivity would need to be enhanced in order to permit realistic field trials to be undertaken.This required modification is not considered a serious problem since the RF "front-end" from the commercial VHF receiver used successfully in implementing important sections of the QNBFAM receiver prototype (the PYE Europa) could be readily employed to achieve the desired level of sensitivity. However,a more appropriate design would involve an RF amplifier configuration to which a form of delayed AGC could be applied in order to extend the dynamic range of the system by preventing front end or first mixer saturation when in receipt of strong signals.With the availablity of dual gate JFET RF transistors such an improvement to the prototype receiver is not seen as an insurmountable problem.

On the subject of automatic gain control,it is relevant to point out that the configuration adopted for the gain controlled IF strip,which drives the AM detector,was chosen because it offered a very convenient solution to the AGC problem.As described in Section 6.3 of Chapter 6,the AGC control introduced is of the traditional feedback approach (FBAGC) and based on a readily available chip-set.Although the results obtained have indicated that the presence of a straightforward FBAGC control undoubtedly improves the system performance,other research workers have confirmed that alternative AGC implementations are better suited to coping with a fast fading radio signal.The published work of McGeehan and Burrows (53,54) has shown that the presence of delay in a FBAGC network makes it virtually impossible to suppress adequately the fast fading phenomena associated with a mobile radio environment.For optimum receiver performance,the faster fluctuations in signal strength which occur because of severe multipath propagation effects,can be more effectively suppressed by means of "feedforward" rather than feedback AGC.However,as yet no consideration has been given to modifying the mode of AGC introduced because of a fundamental limitation of the QNBFAM receiver which will be discussed in the next section.

8.4 An Alternative Receiver Architecture.

The major problem associated with the QNBFAM double superhet receiver prototype developed is undoubtedly the inferior adjacent channel selectivity available.As discussed in Section 6.2 of Chapter 6,a linear phase filter has to be employed,following the first mixer,in order to avoid any corruption of the received multiplexed RF signal envelope which would otherwise result due to the group delay distortion normally associated with narrowband crystal filters.This means that in order to obtain the required group delay performance the amplitude response specifications of the filter have to be relaxed resulting in a significant reduction in selectivity compared with other commercial VHF angle modulated mobile receivers.However,the problem of poor adjacent channel selctivity,due to the QNBFAM systems dependence on a linear phase crystal block filter,can be obviated if an alternative receiver architecture is considered.

An alternative solution can be achieved if a direct conversion or "zero IF" approach is adopted (55).With this technique there is no "image" frequency involved and so radio frequency filtering problems are eased considerably.The radio frequency input signal is mixed with a local oscillator of the same nominal frequency which generates products at "zero" and twice the original frequency respectively.The channel bandwidth and hence receiver selectivity can therefore be defined by a low pass filter at baseband.Obviously,the process of recovering the original modulating signals and obtaining the necessary channel separation of the speech and data baseband signals requires further signal processing.

To achieve this important requirement,results of recent research into a "Universal Radio" architecture (56) conducted at Standard Telecommunications Laboratories (Harlow,Essex) has prompted an alternative demultiplexing strategy to be considered.This alternative demultiplexing approach can also be conveniently used in conjunction with the zero IF receiver structure previously described.A particular advantage of this different technique is that it is not constrained by the need for compatibilty with existing radio equipment and is also well suited to the application of VLSI techniques as most of the circuit functions required can be achieved by exploiting digital signal processing methods.The following discussion provides a brief overview of this alternative QNBFAM receiver architecture.

It has already been shown in Section 2.2.3 of Chapter 2 that the multiplexed QNBFAM signal is a composite angle and amplitude modulated signal which can be represented in the form of Equation 8.1.

$$f(t) = A(t).\sin(\omega_c t + \theta(t)) \qquad 8.1$$

where

$$A(t) = \sqrt{2}\ \{1 - \beta/_2\ (\phi_1(t) - \phi_2(t))\} \qquad 8.2$$

and

$$\theta(t) = \pi/_4 + \beta/_2(\phi_1(t) + \phi_2(t)) \qquad 8.3$$

The terms $\phi_1(t)$ and $\phi_2(t)$ are directly related to the two independent baseband channels as defined in Equations 2.12(a) to 2.12(d).

Consider now this signal being applied to a direct conversion receiver as represented by Figure 8.1.As shown,the QNBFAM signal is mixed with a local oscillator of angular frequency ω_0 ,having both an in-phase and quadrature output.After subsequent low pass filtering this process yields in-phase (Vi) and quadrature (Vq) channel signals of the form defined by Equations 8.4 and 8.5.

$$Vi(t) = \tfrac{1}{2} A(t) \cos [(\omega_c - \omega_o)t + \theta(t)] \qquad 8.4$$

$$Vq(t) = \tfrac{1}{2} A(t) \sin [(\omega_c - \omega_o)t + \theta(t)] \qquad 8.5$$

These channel output voltages correspond to the orthogonal components of a vector $R\underline{|\phi}$,as shown in Figure 8.2,which fully describe the incoming RF signal.The channel output voltages,Vi and Vq,can be digitised and converted from rectangular to polar co-ordinates by a digital signal processor.The speech and data sum and difference signals can in turn be derived from the vector $R\underline{|\phi}$ as shown by the following simple analysis.

Obtaining the R component:

$$R = \sqrt{Vi^2(t) + Vq^2(t)} \qquad 8.6$$

$$= \tfrac{1}{2} A(t) \qquad 8.7$$

Equation 8.7 corresponds to the envelope of the QNBFAM signal which is directly related to the "difference" between the speech and data modulating signals.

Obtaining the ϕ component:

$$\phi = \mathrm{Tan}^{-1} \left[\frac{Vq(t)}{Vi(t)}\right] \qquad 8.8$$

$$\phi = (\omega_c - \omega_o)t + \theta(t) \qquad 8.9$$

The term $\theta(t)$ corresponds to the original angle modulation of the QNBFAM signal which in turn is related to the "sum" of the speech and data channels.This sum signal can be obtained by differentiating Equation 8.9.In a sampled data system this simply corresponds to the difference between successive samples.

$$\phi' = (\omega_c - \omega_o)ts + [s(t) + d(t)] \qquad 8.10$$

As indicated by Equation 8.10,the output signal derived from the ϕ' component consists of the speech and data channel sum signal and a dc offset due to the difference between the carrier and local oscillator frequencies.The required channel separation of the independent speech and data message signals can therefore be derived by a demultiplexer based on simple sum and difference amplifier networks as achieved in the existing implementation.The signal processing functions required for the message signal recovery and channel separation is conveniently summarised in Figure 8.3

Although no investigative work has been undertaken to evaluate this alternative proposition,nevertheless,it does represent a new direction and initiative for the research to be continued in the near future.In fact,besides being more suited to integrated circuit fabrication,the alternative receiver configuration would appear to provide a more elegant solution to the successful implementation of a QNBFAM multiplexing system.Further research effort is obviously needed to test the validity of this statement.

DIRECT CONVERSION RX ARCHITECTURE

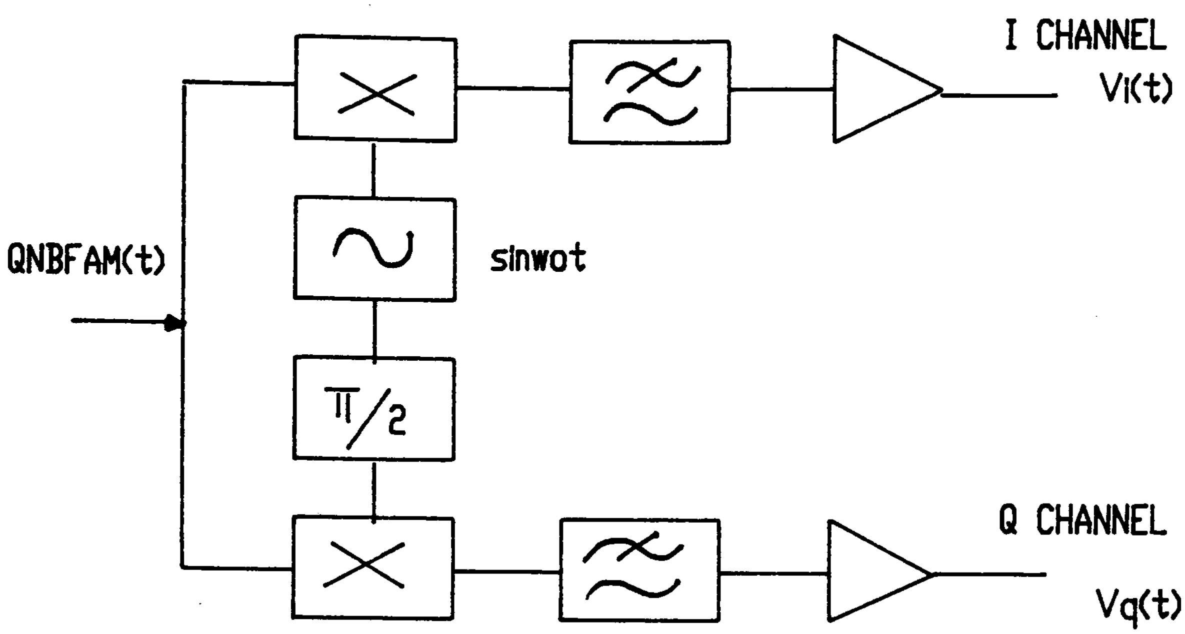


Figure 8.1

PHASOR REPRESENTATION OF RF SIGNAL

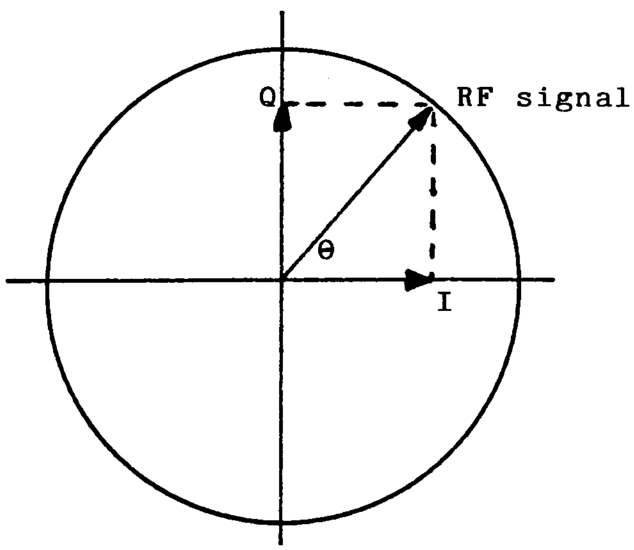


Figure 8.2

CO-ORDINATE CONVERSION & SIGNAL RECOVERY

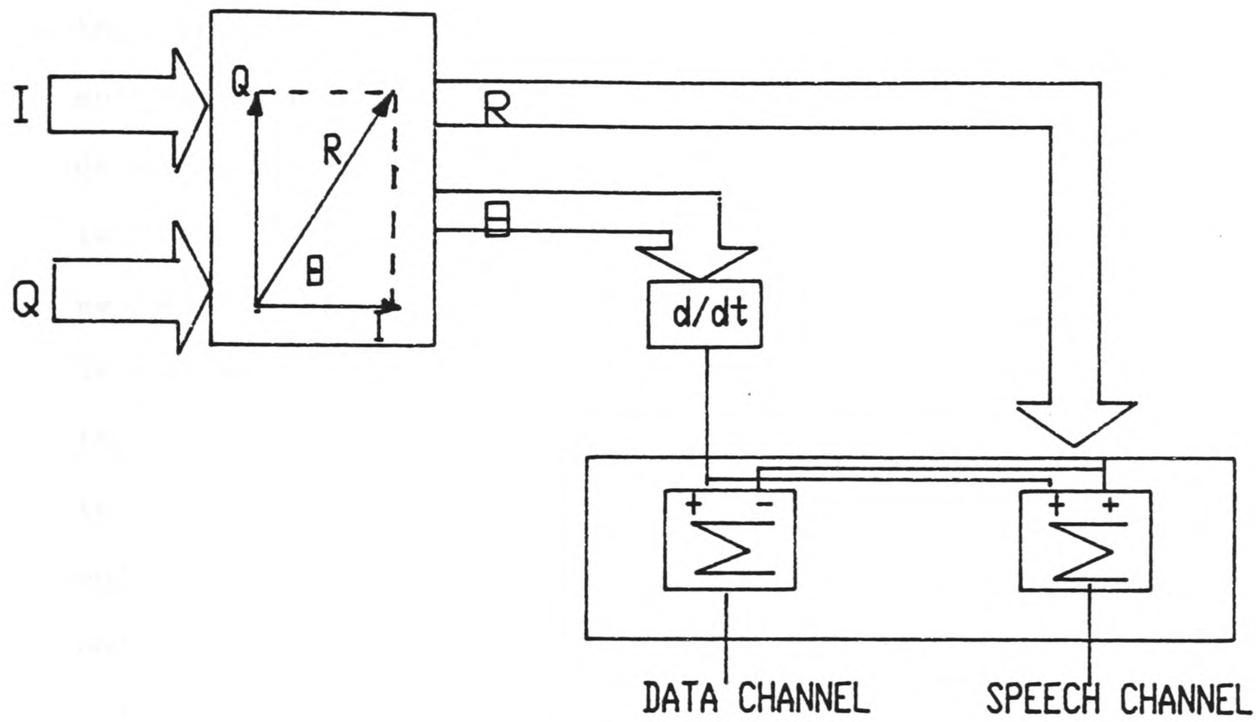


Figure 8.3

CHAPTER 9

CONSIDERATIONS FOR THE FUTURE.

9.1 Reflections on the QNBFAM Research Effort.

The main objective of this section is to review the successes achieved during the execution of the research and to identify the areas still to be resolved.The initial progress made and success achieved was undoubtedly due to the initiative involved in using the Astec 3 simulation software package as an investigative tool.The techniques developed,during the early part of the research effort,led to the simulation software being customised and applied in a way which was particularly suited to the investigations that it was eventually used for.Indeed,the results obtained from these investigations pointed the direction for the research to progress and eventually led to the concept of the QNBFAM. multiplexing strategy.It is relevant at this stage to point out that the functional modelling approach and associated interface software,developed by the author,have also been adopted to support other current research projects.For example,a project being undertaken by research staff at the Polytechnic of Wales,in conjunction with INMOS Ltd.,of Newport,is now using Astec 3 in the functional modelling mode as a vehicle for investigating thermal effects in VLSI semiconductor devices.

Arguably the most important contribution of the research has been the evolution and development of the QNBFAM multiplexing scheme as a means of integrating simultaneous speech and data transmissions over a mobile radio channel.The QNBFAM system,based on narrowband angle modulation,has been thoroughly investigated in a true engineering sense.That is,the need for such a scheme was identified,various potential solutions considered and finally the most appropriate implementation evaluated.In the course of this evaluation transmitter and receiver equipment has been designed,built and tested allowing the operational performance of a laboratory system to be determined.

A comprehensive set of practical results,obtained from the prototype equipment,have shown that the QNBFAM system is indeed capable of supporting a simultaneous transmission of independent speech and data baseband channels. Moreover,because of the unique method employed to multiplex the speech and data channels,the system can offer a significant measure of compatibilty which means it can be used alongside existing VHF mobile transceiver equipment.For example,the receiver configuration of the QNBFAM system will actually permit the reception of conventional NBFM signals without any unecessary deterioration in quality.Also,with only one baseband channel active,speech or data,the transmitter can be easily configured to transmit a narrowband angle modulated signal that can be readily received by conventional mobile radio equipment.However,having the ability to support a multiplexed voice and data transmission the QNBFAM system clearly offers considerably more scope than a single baseband channel mode of operation.

Practical results have also been obtained to assess the degrading effects of Rayleigh fading on the performance of a QNBFAM multiplexed radio transmission.The results have confirmed that,as expected,the interaction of the independent speech and data channels increases during fading conditions.However,by exploiting well established techniques for combating fading,such as diversity reception and an appropriate AGC strategy,an acceptable operational performance can be confidently predicted.As with all data transmissions over a mobile radio channel an effective error correcting code,together with an automatic request repeat transmission protocol,are also recommended to ensure an acceptable BER performance in terms of "error-free" seconds.Further investigations are of course needed to support these claims.

As discussed in Chapter 8,a number of improvements are also needed in terms of the transmitter and receiver equipment in order to make them suitable for field trials.In particular,the selectivity characteristic of the receiver needs to be resolved,which because of the nature of the QNBFAM signal structure,would appear to suggest a complete re-think in terms of the receiver architecture.However,with the recent advances made in cellular and mobile radio communications and its associated technology the implementation of integrated speech and data facilities and other value added services may well be achieved without having to use the QNBFAM system which is still in its infancy.

Inevitably,since the start of the project in the summer of 1984 mobile radio techniques have advanced considerably.In fact,many of the original ideas concerning "value added services" have now become commonplace in certain applications.The following sections discuss some of the more relevant developments that have taken place over the last few years and identify potential techniques that will no doubt lead to far superior mobile radio communications in the future.

9.2 Current Developments in Mobile Radio.

9.2.1 Gascord 1200

As described in section 1.3,British Gas are now conducting field trials with their Gascord 1200 system which uses data transmission to enhance the capacity of each radio channel.It also provides automatic management of large single frequency radio networks and the facility to transmit and receive data in text form.The data transmission rate is 1200 bits/sec and the system can accomodate both speech and data communications between the base station and mobiles as well as between mobiles.Each vehicle is fitted with a hand held portable computer and radio modem such that messages received,or to be transmitted,can be stored in memory.Special error correction techniques have been employed to ensure a high degree of reliability.The system also automatically chooses the best base station to communicate with a particular mobile.The overall system is actually controlled from a central computer which employs a front end processor to provide the error correction.

When a vehicle is selected for contact there is a visible and audible indication to prompt the mobile operator to call base.Calls from the vehicle are queued in order of arrival,though there is a queue jumping system in cases of emergency.The operator can call the base by pressing an emergency key or can indicate the current situation by one of ten other status keys,one of which is a request for speech contact.A protocol built into the system automatically acknowledges the receipt of calls.Should a call not be acknowledged,it is automatically retransmitted,but will be aborted if the second attempt is unsuccessful.The trial area chosen includes some of the busiest parts of central London.The rationale is that if it can operate effectively in one of the most heavily congested areas for radio communications in the world it should be able to work anywhere else!

9.2.2 Trunked Mobile Radio in Band III.

Following advice from the Director General of OFTEL,the DTI have decided to allocate frequencies in Band III to establish private mobile radio on a nationwide basis.In response,new VHF trunked mobile radio systems have been developed and launched in the UK (57).These trunked systems provide an integrated network with a full range of facilities particularly suited to the business user.The national networks,recently launched by "GEC National One" and "Band Three Radio Ltd.",provide a flexible carrier for basic voice and data communications and are intended to offer the following services:-

VOICE SERVICES

* Selective,two-party calls
* Fleet calls (broadcast and group)
* Dispatcher facilities (including queuing)
* Priority and emergency calls
* Unattended mode (storage of call data)
* P.A.B.X. incoming and outgoing calls
* P.S.T.N. outgoing calls (mobile to telephone)
* Advice of call transfer
* Value added voice services

DATA SERVICES

* Status reporting
* Circuit-switched data (audio band)
* Store-and-forward short data messages
* Shared data channel
* Bureau data services
* Access to public data services
* Vehicle tracking and security

In a trunked system mobiles communicate through their respective trunked common base station which are interconnected by a voice and data switching network.This allows authorised mobiles to communicate as they roam throughout the combined coverage area of the networked base stations.

9.2.3 Cellular Radio

The phenomenal growth in cellular radio has been one of the outstanding successes in UK communications this decade.In fact,many of the original ideas concerning the introduction of value added services using the QNBFAM system are now being offered on cellular networks.For example,ISTEL and Racal-Vodata have combined together to introduce value added network services (vans) over the Vodaphone cellular radio network.The system uses the Racal-Vodatel developed special protocol,called cellular data link control (C.D.L.C.) which is aimed at overcoming the problem of transmitting data through the hostile environment of a cellular radio path.C.D.L.C. has been designed to be transparent to all user applications.A 2400 bits/sec modem giving a 1200 bits/sec net user rate is available which incorporates this protocol and is compatible with the vast majority of lapheld and portable terminals.

Even though the two UK cellular networks,Vodafone and Cellnet,only started service in 1985 they have already experienced congestion in the London area.However,to help combat this situation work is already underway to introduce a "Pan-European" digital cellular radio network by the early 1990s.Great Britain,West Germany,France and Italy have already agreed on standards and specifications for the proposed narrowband digital cellular radio network.The indications are that the range of services available to UK users will increase with the implementation of the pan-European network and its intended integration with the Integrated Services Digital Network (ISDN).

The UK research effort to support the introduction of the digital cellular network is being co-ordinated by the Cellular Radio Advisory Group (CRAG),which is chaired by the DTI and includes representatives from British Telecom,GEC/Marconi,Phillips,Plessey,Racal and STC.A joint digital cellular radio "test-bed" experiment is currently being developed by BT at Martlesham,GEC/Marconi at Baddow,and Racal at Reading.Jointly funded by the DTI and the research partners,the experiment uses 125 and 312 kbits/sec digital radio links with 16 kbits/sec speech coding.The experiment is specifically aimed at investigating the performance of a narrowband digital system.

9.2.4 Mobile Packets.

Besides offering a truly integrated service,an all digital network has the potential to offer an alternative and very efficient form of information transfer based on packet switching which would appear to offer significant benefits to mobile radio communications.As discussed in Chapter 7,the major problem with mobile data transmission is the pronounced multipath fading due to scattering and reflection from buildings and other obstructions.Such fading inevitably leads to bursts of errors during the transmission of data.To overcome these error bursts it is normal practice to use error-correcting codes and repeated transmissions,leading to a significantly reduced throughput,in order to achieve some degree of robustness.However,when digital signals are transmitted in very short bursts or packets most of the packets will be accepted by the receiver on its first transmission,since the majority of transmission packets will actually occur between fades.In only a limited number of cases will automatic re-transmission be required.

A T & T Bell laboratories have been investigating the transmission of packetised voice and data on cellular mobile networks operating between 800 and 900 MHz (58).This study is based on high level data link control (H.D.L.C.) procedures as defined by the International Standards Organisation,even though this protocol has not been designed specifically for use on fading channels.This H.D.L.C. uses a 16-bit frame check sequence to detect transmission errors.Unlike more conventional data transmission systems,as long as the system is in a connected mode,there is virtual assurance that a packet receiver will not accept for display a packet that has acquired errors during transmission.The rationale being that if a packet is accepted it will be received without errors.Another benefit of a packet mode of transmission is that by using short bursts of specifically addressed packets,it is possible for many stations to share a single frequency channel.

The Bell Laboratories study has shown that a good throughput can be achieved at UHF with vehicle speeds up to 70 mph.While the average delay increases significantly with increasing packet length,the choice of a very small packet increases the percentage of "overheads" and housekeeping signals and therefore tends to reduce the effective throughput.It is clear that,at least theoretically,there is a useful range of values for the packet size over which the delay is small and efficiency high.This enables the possibility of transmitting packetised speech and data on mobile radio channels with high spectrum utilisation and efficiency.

9.2.5 A Micro-cellular Structure.

Another significant development,worthy of consideration,is the proposed move toward a micro-cellular structure using much higher frequencies than are currently employed.The present cellular mobile radio system has an allocation of the order of 1000 duplex radio channels,however,future systems will require a substantial increase in capacity.A possible solution lies in the use of the 60 GHz band where radio signals attenuate rapidly in air,allowing smaller micro-cells to exist and operational bandwidths of a few gigahertz (59).

The propagation characteristics of electromagnetic signals at 60 GHz allow cell boundaries to be defined accurately by use of appropriate antennas and,if necessary,assembling suitable obstructions to prevent interference.In this way an office block could form a microcellular network,with each room forming its own cell .Out of doors each cell could be made to cover a small portion of a roadway with antennas attached to lamposts or buildings.It is envisaged that,under the new micro-cellular system,an individual telephone number will no longer refer to a subscribers home or office but to a portable transceiver that will be carried at all times.This will then make trying to contact someone by telephone a much less frustrating activity.The large bandwidth available should also make data services and even video services available to the user.

The emerging picture is one of short distance low power communications where packetised ISDN digital data can flow via the PSTN and mobile radio networks.A long term goal is to render the mobile radio network completely transparent so that PSTN signals will not require re-formatting for mobile radio transmission.Indeed,all mobile radio systems will be truly integrated,including communications via satellites to mobiles in remote areas,and to people travelling in aeroplanes and in ships around the world.Eventually a new network will evolve that regards all terminals to be mobile,even if they never move!With a personalised communication number we will be able to communicate with anyone who wishes to do so,at any speed,and from anywhere.Whether or not the QNBFAM system can contribute to the future of mobile communications,with such rapid developments taking place,remains to be seen.

REFERENCES

1."Independent Review of the Radio Spectrum"(30-960 MHz), The Merriman Review,HMSO publication No.9000., 1983.

2.MPT 1303 "Angle-modulated VHF and UHF radio equipments,incorporating integral antennae,for use in the PMR service", June 1983.

MPT 1317 "Code of practice for transmission of digital information over land mobile radio systems",April 1981.

MPT 1323 "Angle-modulated radio equipment for use at base and mobile stations in the PMR service operating in the frequency band 174-225 MHz",July 1983.

(Department of Trade and Industry,Radio Regulatory Division Performance Specifications.)

3.M.C.Pinches and R.D.King, "The potential for further reduction in channel spacing in the VHF band" IEE Conf. Pub No.139.

4.W.Gosling et al,"Sideband diversity,a new application of diversity particularly suited to land mobile radio", The Radio and Electronics Engineer,Vol 48.No.3, March 1978.

5.A.Fleet,"Adjacent channel performance of FM,AM and SSB at VHF" IEE Conf., Comm 80, Pub No.162.

6.R.C.French,"Mobile radio data transmission-error performance", Proc IERE Conf.,Nov 1975.

7.R.C.French,"Mobile radio data transmission in the urban environment" IEEE Int. Conf., ICC 76, June 1976.

8.R.C.French,"Error rate predictions and measurements in the mobile radio data channel", IEEE Trans., VT 27, 1978.

9.J.D.Parsons et al,"Error rate reduction in VHF mobile radio data systems using specific diversity reception techniques", IEE Proc., Vol 127, Pt F., No. 6., Dec 1982.

10.J.R.Edwards,"High reliabilty data transmission to mobile vehicles", The Radio and Electronic Engineer, 1978.

11.J.R.Edwards,"Transceiver design for a mobile radio data transmission system",IERE International Conference on Land Mobile Radio,Sept 1979.

12.W.Gosling and R.J.Holbeche,"A feasibilty study for a voice plus data mobile radio system of the future", IEE Conf., Comm 76., Pub No. 139.

13.J.P.McGeehan and A.J.Bateman, "Phase locked transparent tone-in-band (TTIB) : A new spectrum configuration particularly suited to the transmission of data over SSB mobile radio networks" ,IEE Trans., Com 32., Jan 1984.

14.J.P.McGeehan and A.J.Bateman, "Theoretical and experimental investigations of feedforward signal regeneration as a means of combating multipath propagation effects in pilot based SSB mobile radio systems", IEEE Trans., VT 32., Feb 1983.

15.G.Benelli and R.Fantacci, "An integrated voice-data communication system for VHF links", IEEE Trans., Com 31., No. 12., Dec 1983

16."Astec 3 user manual", SIA Computer Services Ltd.,Edbury Gate,Lower Belgrave,London.

17.P.A.Witting,"Astec 3 introductory manual" SIA Computer Services Ltd.,Edbury Gate,Lower Belgrave,London.

18.M.Charnley,"A cost-effective baseband signal generator" Electronics and Wireless World, Dec 1985.

19.M.R.Scroeder,"Synthesis of low peak factor signals" IEEE Trans. on Information Theory, Jan 1970.

20.G.R.Jessop,"VHF/UHF Manual" 4th Edition , The Radio Society of Gt.Britain., 1983.

21.C.Myers,"RF links replace wire lines in LMR systems" MRT International, Vol 2, Issue 3, 1985.

22."Getting to know Balanced Mixers" Hatfield RF Components, W & G Instruments Ltd., Plymouth, Devon.

23."The Type 1763 Double Balanced Mixer", Hatfield RF Components, W & G Instruments Ltd., Plymouth, Devon, Publ No. 0203-0285.

24."The SL623C AM detector and AGC generator", Radio Linear Circuits Applications Handbook, Plessey Semiconductors Ltd.

25."Designing with the SA/NE604" Mullard Application Note AN199, Mullard Ltd., June 1985.

26."The SAD 1024 bucket brigade delay line", Reticon Corporation Ltd., Sunnyvale California.

27.G.Daryanani,"Principles of Active Network Synthesis and Design" John Wiley $ Sons, 1976.

28."The FX 409 FFSK Modem", Consumer Microcircuits Ltd., Publ No. D/409/3 Feb 1985.

29."Subjective assessment of programme sound" CCIR, Vol XII, Transmission of Sound broadcasts and TV signals over long distances, Report 623, Geneva 1982.

30.G.A.Arredondo and J.I.Smith,"Voice and data transmission in a mobile radio channel", IEEE Trans, VT 26, No. 1, Feb 1977.

31."Broadband Amplifier Applications", Plessey Semiconductors Ltd., Sept 1984.

32.J.F.Ossana,"A model for mobile radio fading due to building reflections : Theoretical and experimental fading waveform spectra" BSTJ 43, Nov 1964.

33.R.H.Clarke,"A statistical theory of mobile radio reception" BSTJ 47, July 1968.

34.W.C.Jakes,"Microwave Mobile Communications" Wiley Interscience, 1974.

35.M.J.Gans.,"A power spectral theory of propagation in the mobile radio environment" IEEE Trans., VT 21.

36.J.D.Parsons et al.,"Median signal strength predictions for mobile radio propagation in London" Elec Letters, 16, No. 5, Feb 1980.

37.W.C.Y.Lee,"Mobile Communication Engineering" McGraw-Hill, 1982.

38.P.J.Mabey,"Mobile radio data transmission - Coding for error control", IEEE Trans., VT 27, No. 3 , Aug 1978.

39.Arredondo et al,"A multipath fading simulator for mobile radio", IEEE Trans, VT 22, No. 4, Nov 1973.

40.J.R.Ball,"A real-time fading simulator for mobile radio", The Radio and Electronic Engineer, Vol 52, No. 10, Oct 1982.

41.W.J.Hurd,"Efficient generation of statistically good pseudo-noise by linearly interconnected shift registers", IEEE Trans, C 23, No. 2, Feb 1974.

42.E.J.Watson,"Primitive polynomials (Mod 2)" Mathematics of Computation, 16, No. 79, July 1962.

43.H.R.Beastall,"A white noise generator", Wireless World, March 1972.

44.D.McGee,"Switched capacitor filters in CMOS", Electronic Engineering, Jan 86.

45."The SL 640 double balanced modulator", Radio Linear Circuits Application Handbook, Plessey Semiconductors Ltd.

46.D.S.Deogan,"A fading simulator for mobile radio applications", BSc(Hons) Project dissertation, Dept of E&EE, The Polytechnic of Wales, May 1987.

47."ILS-PC1 a Users Guide", Signal Technology inc., California, August 1984.

48.M.Charnley,"Signal processing using ILS-PC1 : A student guide to using the Integrated Laboratory System on the IBM personal computer", Dept of E&EE, The Polytechnic of Wales, Sept 1987.

49."MOSPOWER Applications Handbook", Siliconix Inc., Santa Clara, California, 1984.

50.V.Petrovic and W.Gosling,"VHF/AM transmitter using VMOS technology", Electronic Engineering , June 1978.

51."The FX315 CTCSS Encoder", Consumer Microcircuits Ltd., Publ. D/315/3, Feb 1987.

52.MPT 1306,"Continuous tone controlled signalling for use in the Land Mobile services", DTI Radio Regulatory Division, Jan 1978.

53.J.P.McGeehan,"Automatic Gain Control", Chapt. 7, Radio Receivers, IEE Telecomms Series, Peter Peregrines Ltd., 1986.

54.D.F.Burrows and J.P.McGeehan, "The application of feedforward AGC to Mobile receivers as a means of reducing multipath interference", IEE Colloquium on Modern Techniques for Combatting Multipath Interference, Digest No. 1979/62, Nov 1979.

55.I.A.W. Vance,"An integrated cicuit VHF radio receiver", The Radio and Electronic Engineer ,Vol 50, No. 4, Apr 1980.

56.P.A.Masterton et al,"Digital Techniques for Advanced Radio", IEE Int Conf. on Mobile Radio Systems, Sept 1984.

57.P.J.Delow,"Trunked Mobile Radio in Band III",Electronics and Wireless World,Dec 1986.

58.M.R.Karim,"Packetised Voice and Data on Cellular Mobile Networks",A.T. & T Technical Journal,May/June 1986.

59.C.J.Haslett,"Use of 60 GHz band could lead to truly personalised mobile communications",Electrotechnology,Jan 1987.

APPENDICES

APPENDIX A

"Comparison of FM and PM modulated signals"

(i) Introduction.

Both phase and frequency modulation are produced during angle modulation and consequently PM and FM are closely related.

(ii) Relationship between frequency and phase deviation.

The instantaneous phase and frequency deviation of an angle modulated wave are related by the following expressions:-

$$\phi(t) = 2\pi \int_0^t fi(t)\,dt \qquad \text{A1}$$

or

$$fi(t) = \frac{1}{2\pi} \frac{d\phi(t)}{dt} \qquad \text{A2}$$

where: $\phi(t)$ = Phase deviation.

$fi(t)$ = Frequency deviation.

By inspection, the frequency deviation of an angle modulated signal is proportional to the "rate of change" of phase deviation.

(iii) Comparison of FM and PM signals with tonal modulation.

Let message signal = $V_m \cos \omega_m t$

Then:

$$e(t)\Big|_{FM} = A_c \cos(\omega_c t + m_f \sin \omega_m t) \qquad A3$$

$$e(t)\Big|_{PM} = A_c \cos(\omega_c t + m_p \cos \omega_m t) \qquad A4$$

The following observations can be made with regard to the phase deviation and the frequency deviation of both types of angle modulated signal:-

Case 1: Phase Deviation ($\Delta\phi$).

* In both FM and PM " $\Delta\phi$ max" is proportional to the message signal amplitude.
* The phase deviation of the PM wave is independent of the message signal frequency and in phase with the message signal amplitude.
* The phase deviation of the FM wave is inversely proportional to the message signal frequency and in quadrature with the modulating signal amplitude.

Case 2: Frequency Deviation (Δf_c).

* In both FM and PM " $\Delta f_c(max)$" is proportional to the message signal amplitude.

* The frequency deviation of the PM wave is proportional to the message signal frequency and in quadrature with the message signal amplitude.

* The frequency deviation of the FM wave is independent of the message signal frequency and in phase with the message signal amplitude.

(iv) Spectrum of an angle modulated waveform.

For convenience, consider the expansion of Equation A3:

$$e(t)\Big|_{FM} = Ac\ \cos(\ \omega ct + mf\sin\omega\ mt) \qquad A3$$

$$= Ac.Re(exp(j\ \omega ct).exp(jmf\sin\ \omega mt)) \qquad A5$$

The function $exp(jmf\sin\ \omega\ mt)$ is periodic in ωm and can therefore be written as a Fourier series:

$$exp(jmf\sin\omega\ mt) = \qquad Cn exp(\ jn\omega\ mt) \qquad A6$$

The coefficients "Cn" represent the magnitudes of the sidefrequency components which depend in turn on the modulation index mf and the corresponding Bessel function:-

$$Cn = Jn(mf) \qquad A7$$

Therefore,Equation A5 can be re-written:-

$$e(t)_{FM} = Ac.Re(\sum_{n=-\infty}^{+\infty} Jn(mf)exp(j\omega ct + jn\omega mt)) \qquad A8$$

Taking "real" parts:-

$$e(t)_{FM} = Ac(\sum_{n=-\infty}^{+\infty} Jn(mf)cos(\omega ct + n\omega mt)) \qquad A9$$

Observations:

1. An infinite set of sidefrequencies are produced even when modulated by a single tone.However,not all sidefrequencies are significant.Any sidefrequency < 1% of the unmodulated carrier maybe ignored without introducing distortion.

2. The amplitude of each component of an angle modulated wave,including the carrier term,is a function of the modulation index.Therefore,the modulation index determines the number of significant sidefrequencies and hence the required transmission bandwidth.

3. Even order sidefrequencies are colinear with the carrier component,whereas,odd order sidefrequncy components are in quadrature with the carrier.

APPENDIX B

"Spectra of QNBFAM signals and their corresponding Basebands"

Case 1: Tonal Modulation.

(i) Baseband Spectra.

I Modulation = 1 kHz tone.

Q Modulation = 0

Reference Level = -5 dBm

Vertical Scale = 5 dB/Div

Centre Frequency = 3 kHz

Frequency Span = 6 kHz

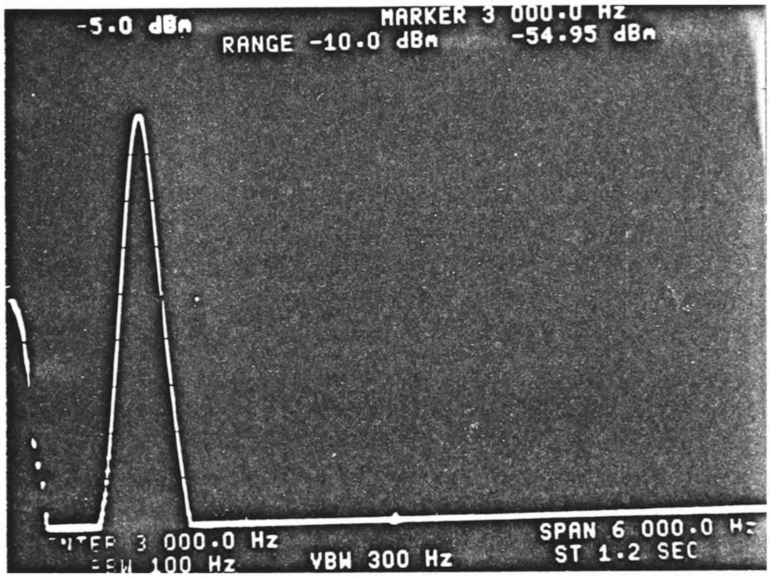


Figure B1(i)

(ii) QNBFAM RF Spectra

Δ fc(max) = 1.25 kHz

Reference Level = -10 dBm

Vertical Scale = 10 dB/Div

Centre Frequency = 170 MHz

Frequency Span = 100 kHz

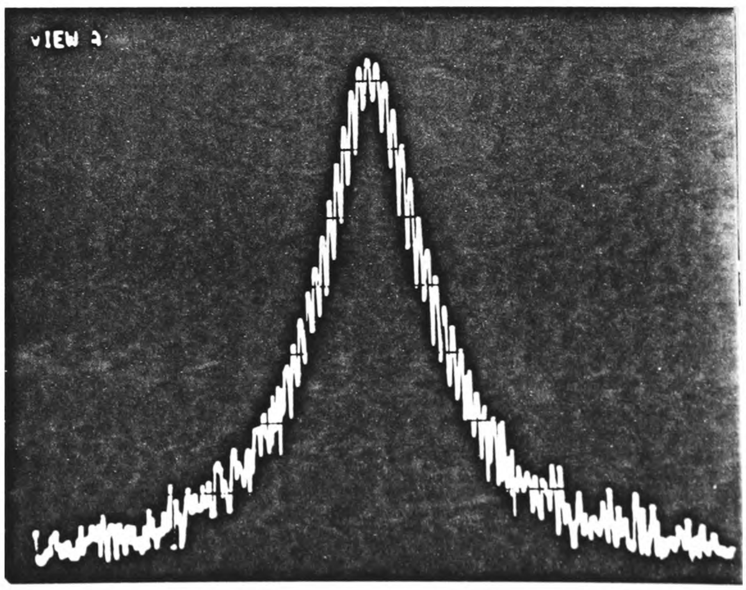


Figure B1(ii)

Case 2: FFSK Modulation

(i) Baseband Spectra.

I Modulation = 0

Q Modulation = 1200 Baud FFSK

Reference Level	=	-20 dBm
Vertical Scale	=	5 dB/Div
Centre Frequency	=	2 kHz
Frequency Span	=	4 kHz

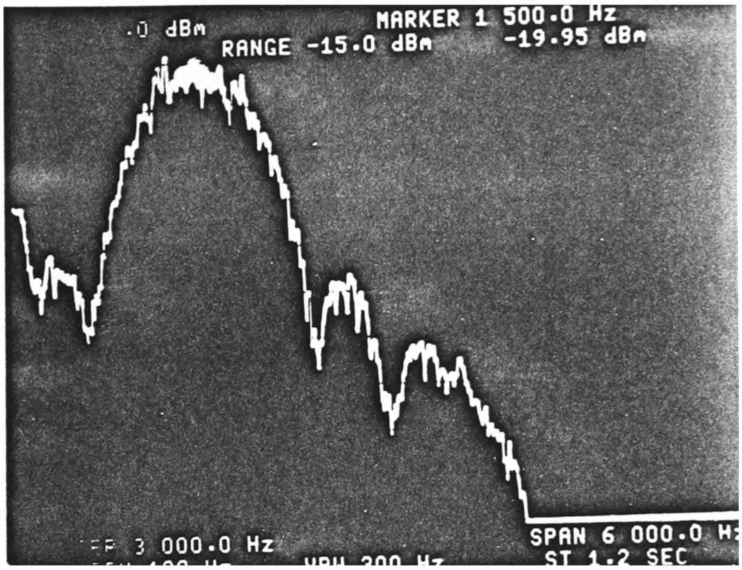


Figure B2(i)

(ii) QNBFAM RF Spectra

$\Delta fc(max)$ = 1.25 kHz

Reference Level	=	-10 dBm
Vertical Scale	=	10 dB/Div
Centre Frequency	=	170 MHz
Frequency Span	=	100 kHz

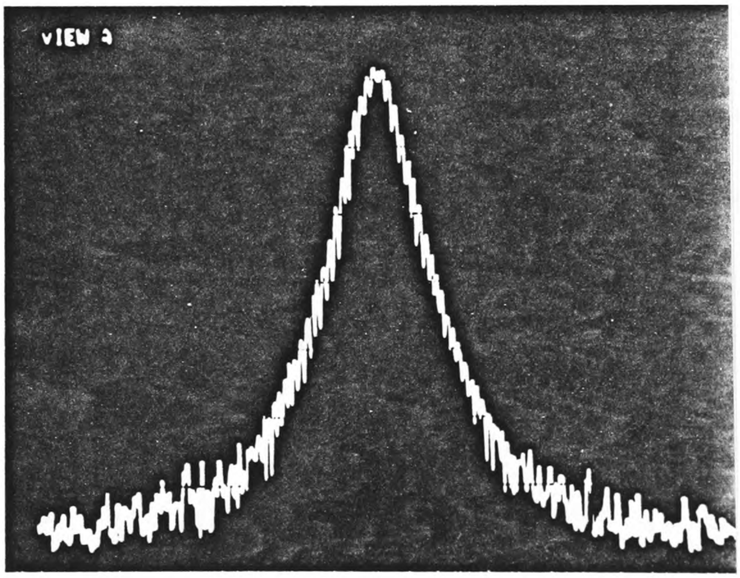
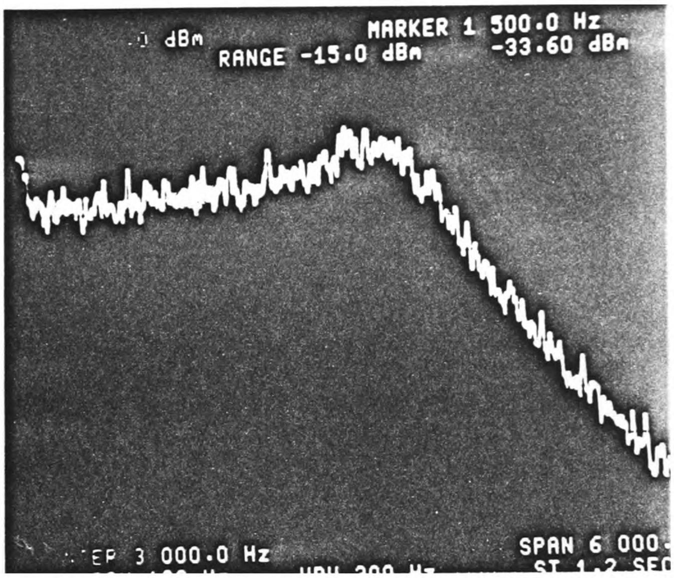


Figure B2(ii)

Case 3: Bandlimited Noise

(i) Baseband Spectra



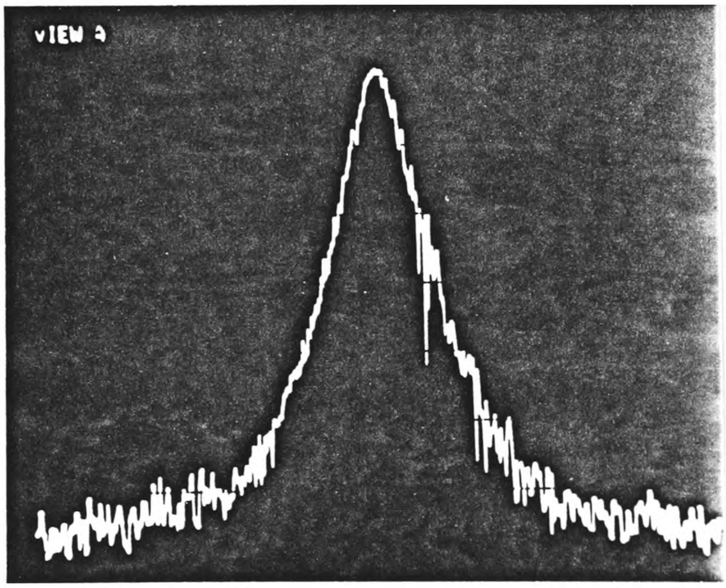
I Modulation = Bandlimited Noise

Q Modulation = 0

Reference Level	=	-20 dBm
Vertical Scale	=	5 dB/Div
Centre Frequency	=	3 kHz
Frequency Span	=	6 kHz

Figure B3(i)

(ii) QNBFAM Spectra



$\Delta fc(max) = 1.25$ kHz

Reference Level	=	-10 dBm
Vertical Scale	=	10 dB/Div
Centre Frequency	=	170 MHz
Frequency Span	=	100 kHz

Figure B3(ii)

Case 4:Simultaneous Modulation of I & Q Channels

(i) Baseband Signals

I Modulation = Bandlimited Noise (Figure B3(i))

Q Modulation = 1200 Baud FFSK (Figure B3(ii))

(ii) QNBFAM RF Spectra

$\Delta fc(max)$ = 2.5 kHz

Reference Level = -10 dBm

Vertical Scale = 10 dB/Div

Centre Frequency = 170 MHz

Frequency Span = 100 kHz

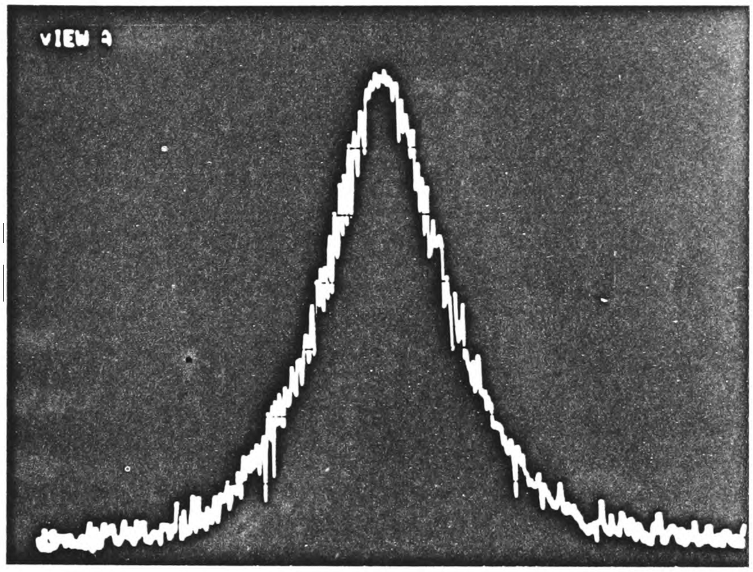


Figure B4

Note:-

(a) Baseband spectra measured by a Hewlett-Packard 3584A 20 Hz-40 MHz Spectrum Analyser.

(b) RF spectra measured by a Hewlett-Packard 8559a 0.01-21 GHz Spectrum Analyser.

APPENDIX C

"The Distortion Factor" (DF rms)

Evaluation of the distortion caused by synchronous detection of a narrowband angle modulated waveform conveying a complex modulating signal.

(i) The Complex Modulating Signal:

$$m(t) = 0.5\sin(3\omega t) + 0.5\cos(5\omega t) + 0.5\sin(7\omega t) + + 0.5\cos(11\omega t) \qquad \text{C1}$$

where $\omega = 2\pi f1$ and $f1 = 200$ Hz

(ii) Bandwidth considered:

From $f1 = 200$ Hz to $f17 = 3.4$ kHz

(iii) The Distortion Factor:

In a conventional distortion factor meter the total voltage of the unwanted frequency components is measured by filtering out the fundamental and measuring the voltage of the residue.This is then compared with the total signal voltage in such a way that the instrument will indicate the distortion factor directly.If the fundamental component is S,harmonic distortion is D,and the noise N,the total signal is therefore (S+D+N).When S is eliminated (D+N) will therefore remain.The distortion term D will comprise a number of harmonic components H1,H2,H3 etc.Consequently,if the distortion meter employs a true r.m.s. indicator,the indicated distortion factor will be given by:-

$$DF\ rms = \frac{\sqrt{H1^2 + H2^2 + \ldots\ldots\ldots + N^2}}{\sqrt{S^2 + H1^2 + H2^2 + \ldots + N^2}} \qquad C2$$

This represents the distortion factor as commonly defined.

Therefore,for the evaluation to be considered which involves a complex periodic signal define:-

$$DF\ rms = \frac{\sqrt{H + N}}{\sqrt{S + H + N}} \qquad C3$$

where: S = Normalised power of the fundamental components of the recovered version of m(t).

H = Normalised power of the harmonic components of the recovered version of m(t) within the baseband.(200 Hz - 3.4 kHz)

N = Normalised power of all other spurious components within the recovered baseband. (200 Hz - 3.4 kHz)

APPENDIX D

"Distortion of the SINXOS baseband spectra"

(i) The SINXOS Baseband Signal:

$$ml(t) = 0.15\sin(\omega t) + 0.32\cos(2\omega t) + 0.45\sin(3\omega t) + \\ + 0.5\cos(4\omega t) + 0.45\sin(5\omega t) + 0.32\cos(6\omega t) + \\ + 0.15\sin(7\omega t)$$

where: $\omega = 2\pi f1$ and $f1 = 200$ Hz

(ii) Distortion caused by synchronous detection

The distortion of the recovered SINXOS baseband spectra after synchronous detection of a QNBØM multiplexed signal is evaluated in the following way:-

Let P = Total normalised power of the signal recovered in Channel 1 (over the SINXOS baseband).

where: P = S + D

S = True signal power

D = Mean square error

also: $$D = \frac{1}{N}\sum_{n=1}^{N} en^2$$

where: (a) en = The error between the rms value of each spectral component, as obtained by simulation, and what the actual value of each component should be to maintain the sinc(x) spectral distribution.

(b) N = Number of frequency components of the SINXOS baseband spectra.

Therefore:-

$$\% \text{ Distortion} = \frac{D}{S} \times 100\% = \frac{D}{P-D} \times 100\%$$

Note: A separate computer program was prepared to perform the relevant analysis on the results obtained by computer simulation.

APPENDIX E

"Group Delay Distortion"

(i) Important Aspects of Group Delay.

Group delay was first related to electrical networks by Nyquist in 1928 and defined as in Figure E.1.

Two aspects of Group Delay:-

A:Signal Propagation.

$$tg = -\frac{d\phi}{d\omega} \qquad \text{E1}$$

The group delay "tg" determines the propagation delay of signal energy or information (Envelope delay).The phase delay "tp = $-\phi/\omega$ " determines the steady state phase relationship between input and output (Carrier delay).For an ideal network tg=tp.

B:Signal Distortion.

$$\frac{d\phi(\omega)}{d\omega} = \underbrace{to}_{\text{Constant:Ideal}} + \underbrace{\frac{d[\Delta\phi(\omega)]}{d\omega}}_{\text{Deviation:Non-Ideal}} \qquad \text{E2}$$

The derivative of the phase characteristic can be used to specify and measure the deviation from the ideal (Linear) phase characteristic.Only the deviation from the constant (Group delay) is important in this case.

Figure E.2 shows the propagation delay and the electrical length of a transmission line having an ideal phase characteristic.In a practical network,the phase characteristic is not strictly a linear function of frequency.Therefore, the signal delay varies for different spectral components and other delay terms become meaningful as explained in the following paragraphs.

Phase Delay:

$$tp(\omega) = -\frac{\theta(\omega)}{\omega} \qquad E3$$

The phase delay of a network at any frequency is given by the slope of the vector starting at the origin (ω =0) and ending at the point ω , $-\theta$ .The phase delay does not indicate a propagation delay,it describes the steady state phase relationship between input and output for a specific frequency component ω .

Group Delay:

$$tg(\omega) = -\frac{d\theta(\omega)}{d\omega} \qquad E4$$

The group delay of a network at any given frequency ω is given by the derivative or the slope of the phase characteristic at that point ω .In general,the group delay describes the propagation delay of a narrow frequency group with centre frequency ω with the assumption that the amplitude characteristic is constant and the phase is linear over the same interval $\Delta\omega$ .However, a certain group delay time tg does not mean that no energy arrives at the output of the network before that time.Half the energy of the frequency group $\Delta\omega$ arrives before the time tg.The group delay can also be interpreted as the time it takes for a sinusoidal signal of constant frequency ω applied to the network at t = 0 to build up to 50% of its final value.

Envelope Delay:

$$te(\omega) = \frac{\Delta\theta e(\omega)}{\omega_m} \qquad \text{E5}$$

The envelope delay is the steady state delay of the signal envelope of a modulated signal and approaches the definition of group delay for a narrow signal spectrum ($\omega_m \longrightarrow 0$) relative to the slope changes of the phase characteristic.Envelope delay is often used interchangeably with group delay for this reason.(Figure E.3)

Delay of the Centre of Gravity:

$$te(\omega) = - \frac{d\theta(o)}{d\omega} = tg(0) \qquad \text{E6}$$

In a dispersive network,the signal envelope gets distorted and the signal delay cannot be well defined at the end points.(Figure E.4).In this case,it is more convenient to reference the delay time to the "centre of gravity" of the signal.The propagation time of the centre of gravity is the group delay time for $\omega=0$ or $\omega=\omega_0$ in the bandpass case.

Signal Front Delay:

$$tsf = - \lim_{\omega \to \infty} \frac{\theta(\omega)}{\omega} \qquad E7$$

As the signal disperses due to non-ideal network characteristics, the front of the signal obviously propagates faster than the centre of gravity and, therefore, the front is less delayed.

(ii) Signal Distortion.

The derivative $d\phi/d\omega$ also indicates a deviation from an ideal (distortion free) linear phase and becomes a figure of merit for the phase linearity of a network (Figure E.5). An example of signal distortion due to a non-constant group delay is shown in Figure E.6 which indicates the effect of a wide deviation angle modulated signal being passed through a linear system with a non-ideal phase characteristic. Now each component of the modulated signal can be considered as a narrow band signal centered around the instantaneous frequency ω_i. Consequently, after passing through the network, the signal components are shifted according to the group delay characteristic and the output signal indicates the presence of harmonic distortion.

(iii) Distortionless Transmission.

The criteria for distortionless transmission depends on two important criteria. First, the amplitude response must be flat over the bandwidth of interest.This means all frequencies within the bandwidth will be attenuated identically. Second, the phase response must be linear over the bandwidth of interest.This means that the relative phase relationships between frequencies within the bandwidth will be preserved after transmission.

A realisable network or system will only approximate the ideal transfer characteristic and the residuals $\Delta A(\omega)$ and $\Delta\phi(\omega)$ will cause signal distortion. $\Delta A(\omega)$ is the deviation from the ideal (constant) amplitude characteristic and $\Delta\phi(\omega)$ is the deviation from the ideal (linear) phase as indicated in Figure E.7.

For most practical networks, the deviations $\Delta A(\omega)$ and $\Delta\phi(\omega)$ can sufficiently and accurately be approximated by representing them as power series expansions of the form:-

$$\Delta A(\omega) = a1\omega + a2\omega^2 + \ldots .$$
$$\Delta\phi(\omega) = b2\omega^2 + b3\omega^3 + \ldots . \qquad \text{E8}$$

...around the centre of the passband (ω_c).(Figure E.8.)

By substituting for $\Delta A(\omega)$, $\Delta\phi(\omega)$ in the equation describing the input/output signal relationship the signal modifications due to the amplitude and phase coefficients ai and bi can be calculated and tabulated for standard signals:-

$$g(t) = F_T^{-1}[F(j\omega).(1 + \Delta A(\omega) - j\Delta\phi(\omega) + \ldots)] \qquad \text{E9}$$

(iv) Distortion of Modulated Signals.

For an amplitude modulated signal of the form :-

$$f(t) = \mathrm{Re}(1 + \alpha(t))\exp(-j\omega_c t) \qquad \text{E10}$$

Table 1 of Figure E.9 shows the signal or distortion components which will be generated as a function of the coefficients ai,bi.The resulting signal contributions are grouped in the direction of the carrier as P_α and perpendicular to it as Q_α .In the case of AM signals,the P component adds to or modifies the original information α (t) whereas the Q component introduces parasitic phase modulation.For example,a linear component a_1 in the amplitude characteristic will introduce no change in the index of modulation of the AM signal,however,additional phase modulation is introduced which will distort the signal envelope.This form of distortion is called AM to PM conversion as illustrated in Figure E.10.

A similar table can be devised for angle modulated signals as shown in Figure E.11.In this case the Q components are the ones which directly interfere with the original modulation information,whereas,the P components only affect the amplitude of the signal.The residual AM due to the P components can often be eliminated by a subsequent limiter without affecting the angle modulation.However,as discussed in the thesis,the presence of a residual AM is a serious problem which can lead to unwanted interchannel crosstalk.In contrast,a parabolic phase component "b2", as indicated in Figure E.12 can generate second harmonic distortion due to the $b2\dot{\psi}^2$ which cannot be separated from the original information content.

The presence of another ψ_2 in addition to ψ_1 ,leads to a form of intermodulation which is often referred to as differential delay (Figure E13).In this case the magnitude or the phase relationship of the original signal $\psi_1(t)$ will change as a function of $\psi_2(t)$ in the presence of a deviation in the transfer characteristic (e.g. $a_3\omega^3$).The magnitude change due to ψ_2 is usually referred to as differential gain and the phase change as differential phase.

(v) Summary and Conclusions.

Distortion effects on Modulated signals:

* Change of depth/index of modulation.

* Non-linear envelope distortion.

* Modulation conversion (AM to PM) and (PM to AM)

* Intermodulation when more than one modulating signal is present.

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* Non-linear envelope distortion.

* Modulation conversion (AM to PM) and (PM to AM)

* Intermodulation when more than one modulating signal is present.

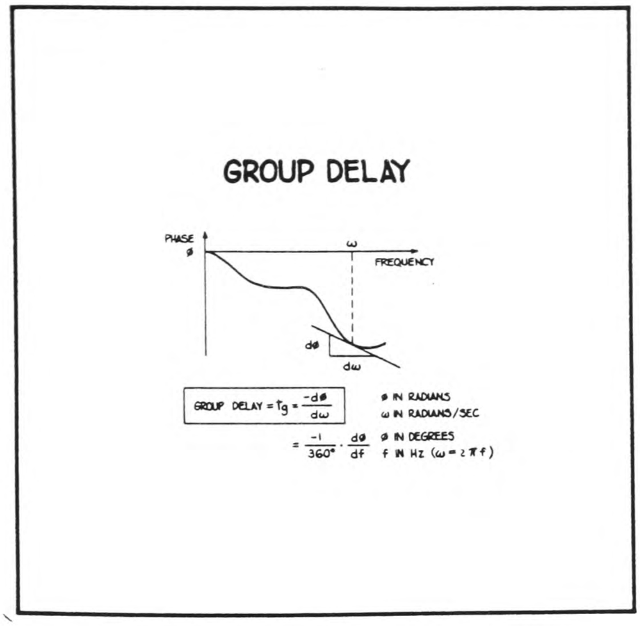


FIGURE E.1

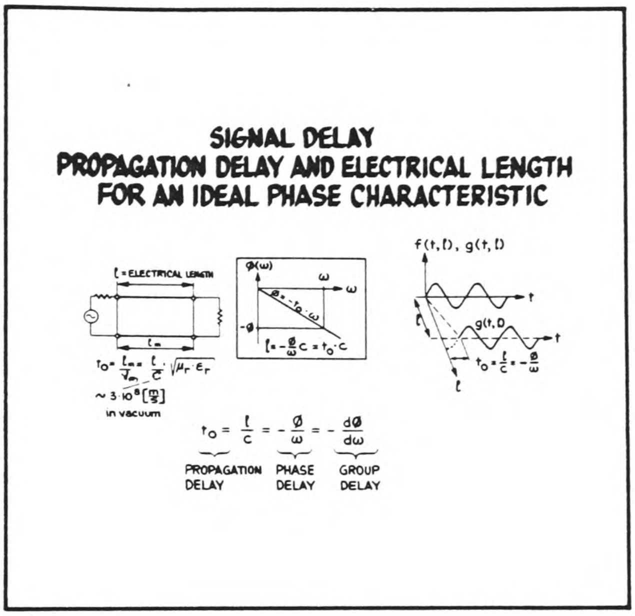


FIGURE E.2

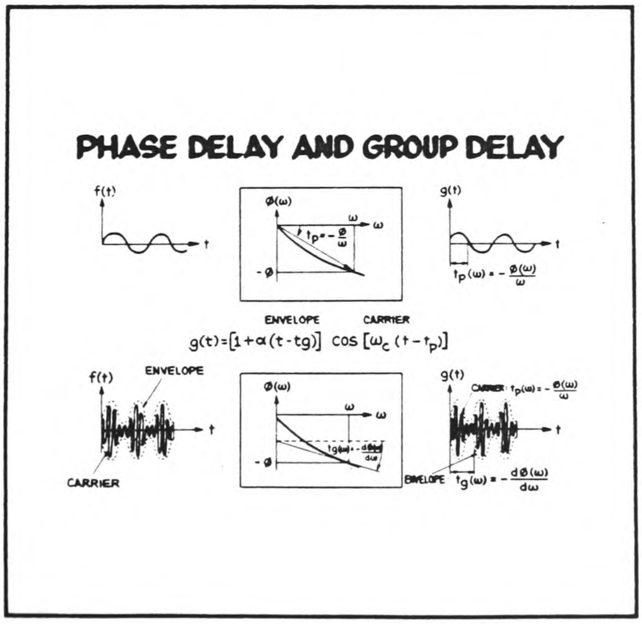


FIGURE E.3

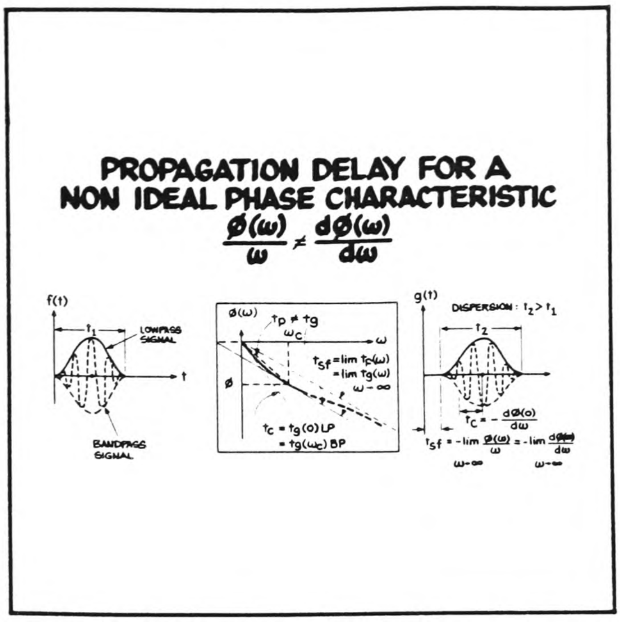


FIGURE E.4

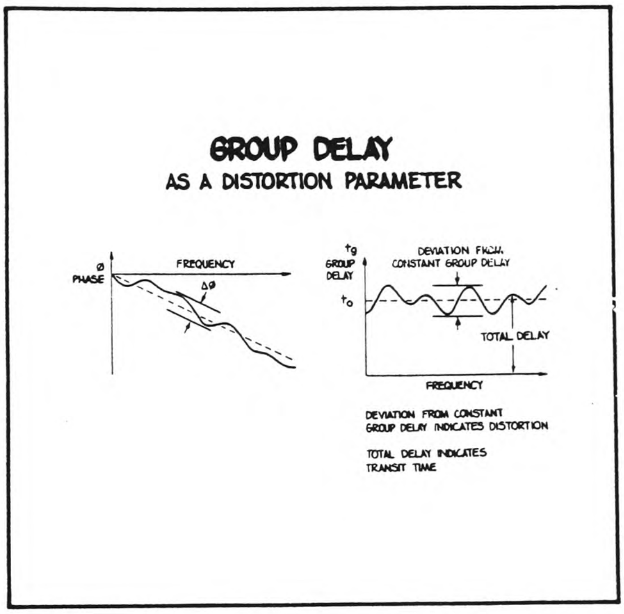


FIGURE E.5

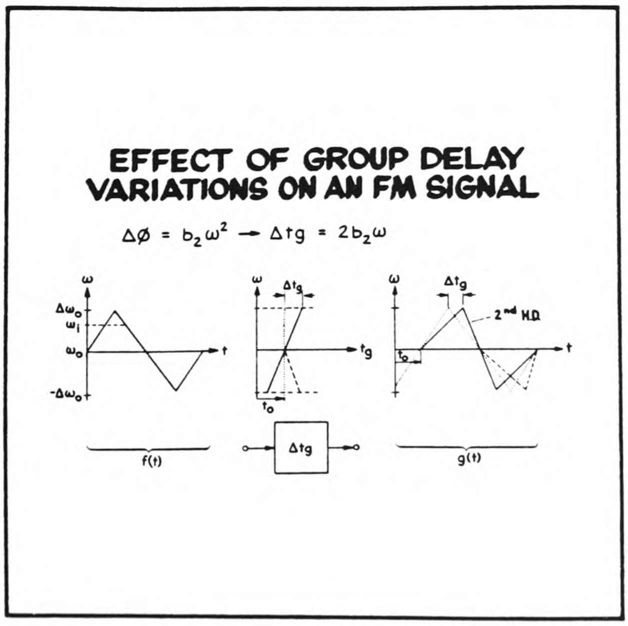


FIGURE E.6

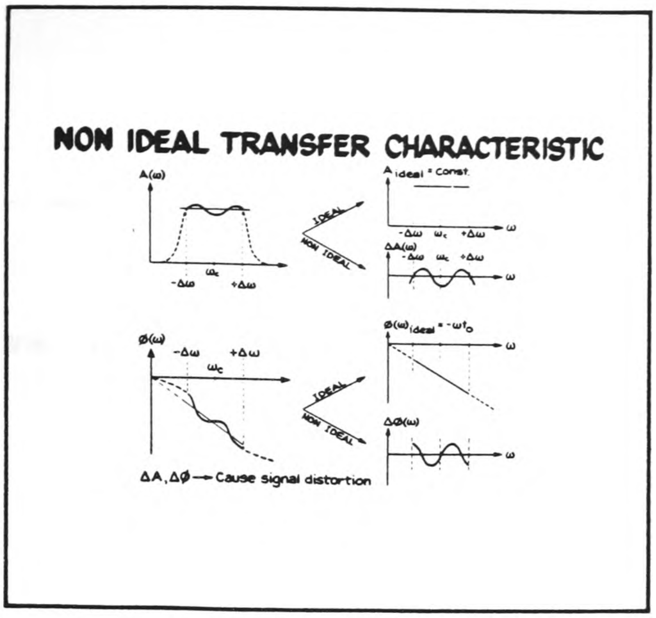


FIGURE E.7

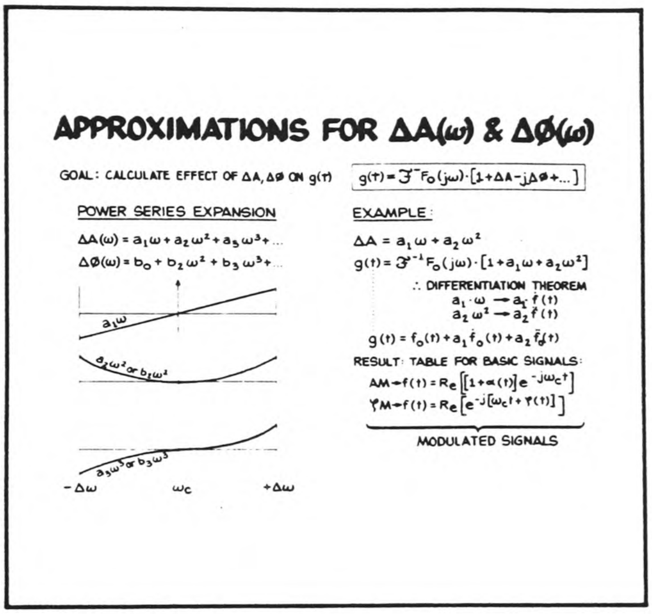


FIGURE E.8

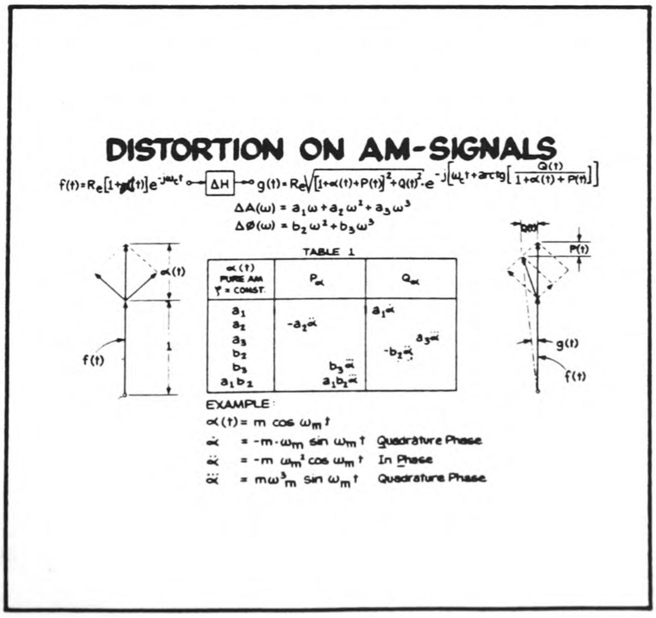


FIGURE E.9

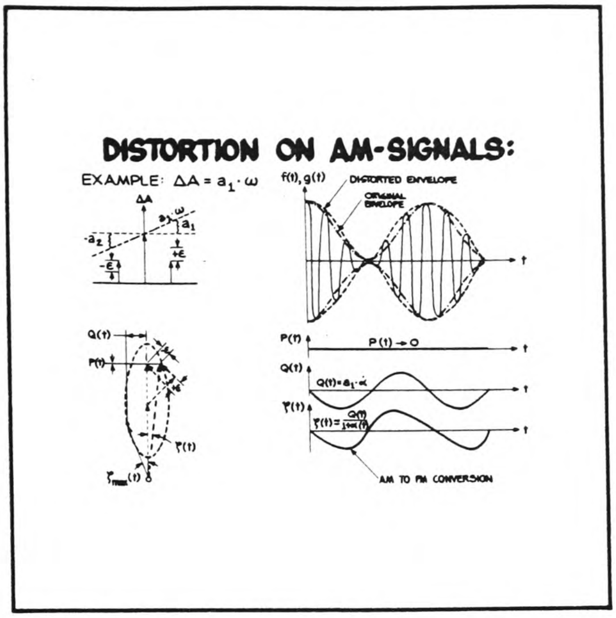


FIGURE E.10

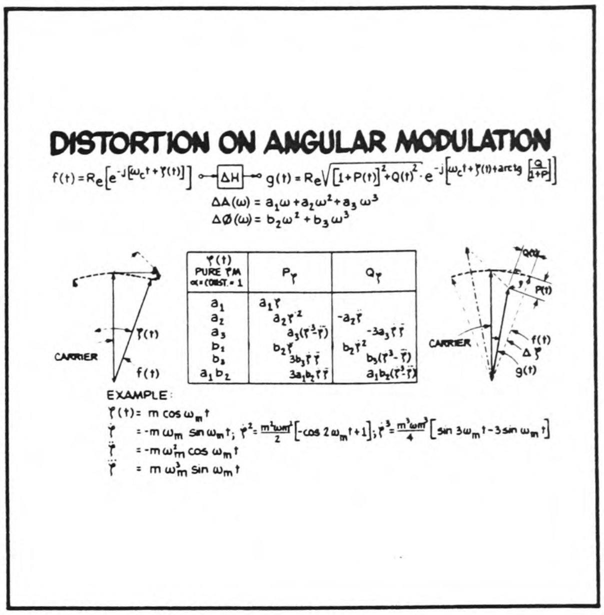


FIGURE E.11

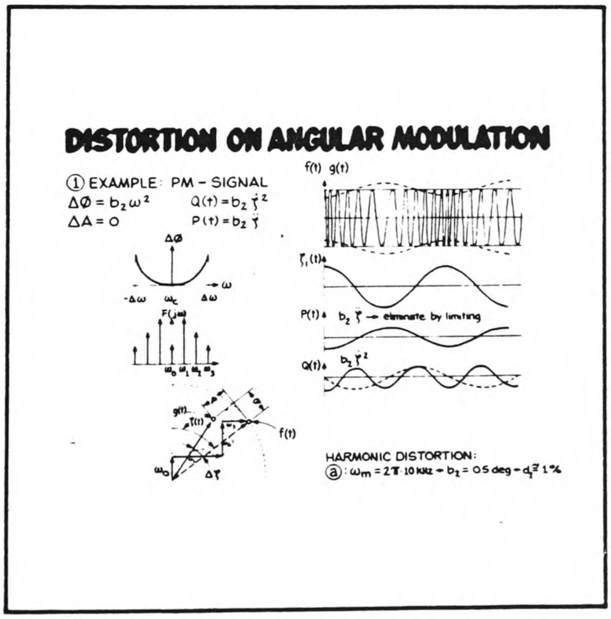


FIGURE E.12

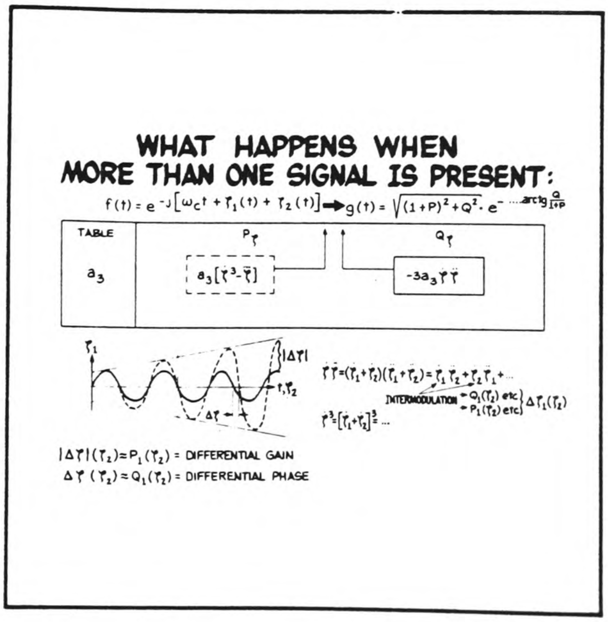


FIGURE E.13

APPENDIX F

Fast Frequency Shift Keying" (FFSK)

(i) Introduction.

Frequency Shift Keying (FSK) is one of the oldest and more popular methods of signalling binary data over a bandlimited transmission channel.Very simply,the two logic states "1" and "0" are represented by two distinct frequencies:-

$$s_1 = A\cos\omega_1 t = A\cos 2\pi (fo + \delta)t \qquad \text{F1}$$

$$s_2 = A\cos\omega_2 t = A\cos 2\pi (fo - \delta)t \qquad \text{F2}$$

where:

fo = Centre Frequency

δ = Deviation or frequency shift

β = Modulation index = $\Delta fc/fm = (fs-fm)Tb = 2\delta\ Tb$

Tb = Bit duration

Fast Frequency Shift Keying (FFSK) is a special case of FSK.It is characterised by a peak to peak frequency deviation of one half the maximium transmitted bit rate and a phase continuous waveform.

Useful Properties of FFSK:-

(a) Bandwidth only 1.5xBaud Rate

(b) FFSK spectrum decays more rapidly (f^{-4}) in contrast to PSK spectra (f^{-2}).

(c) Self synchronisation capabilities.

(ii) Maintaining Phase Continuity.

FFSK is a form of "continuous-phase" frequency shift keying (CPFSK).Essentially,FFSK is binary FSK except that the mark and space frequencies are synchronised with the input binary rate.With FFSK the mark and space frequencies are selected such that they are separated from the centre frequency by an exact odd multiple of one half of the data rate (i.e: fm and fs =n(fb/2),where n = any odd whole integer).This ensures that there is a smooth phase transition in the analogue output signal when it changes from a mark to a space frequency,or vice versa.This situation is shown in Figure F1 which also compares the FFSK output waveform with that of a non-continuous FSK waveform.As indicated,each transition of the FFSK waveform occurs at a zero crossing,consequently, there are no phase discontinuities as there are in the case of the FSK signal.

(iii) The FFSK Spectrum.

As previously described,with FFSK abrupt changes at the bit transition instants, characteristic of other FSK implementations, are avoided.This results in the frequency spectrum of an FFSK signal being narrowly concentrated around the centre frequency making this modulation system very attractive for transmission over restricted bandwidths (Mobile radio!).

Consider the spectrum of a practical 1200 baud FFSK signal:-

$f_c = 1500$ Hz

$f_s = 1800$ Hz

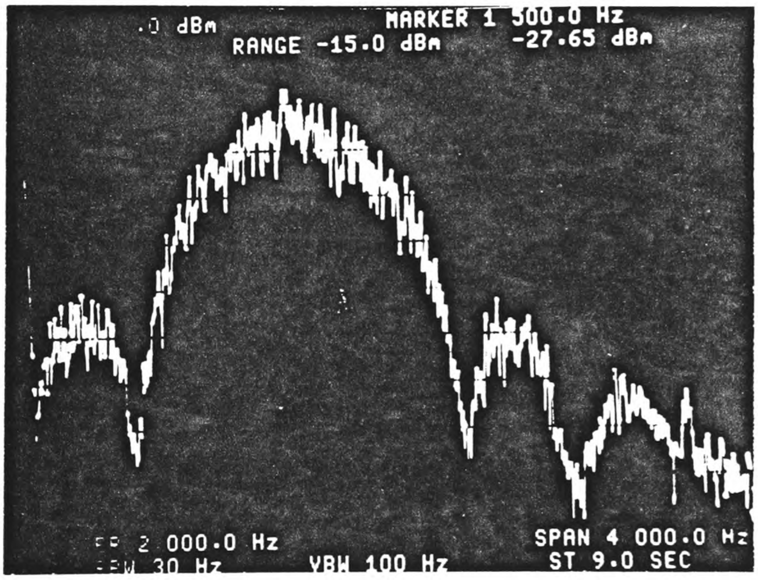
$f_m = 1200$ Hz

$\Delta f = 600$ Hz

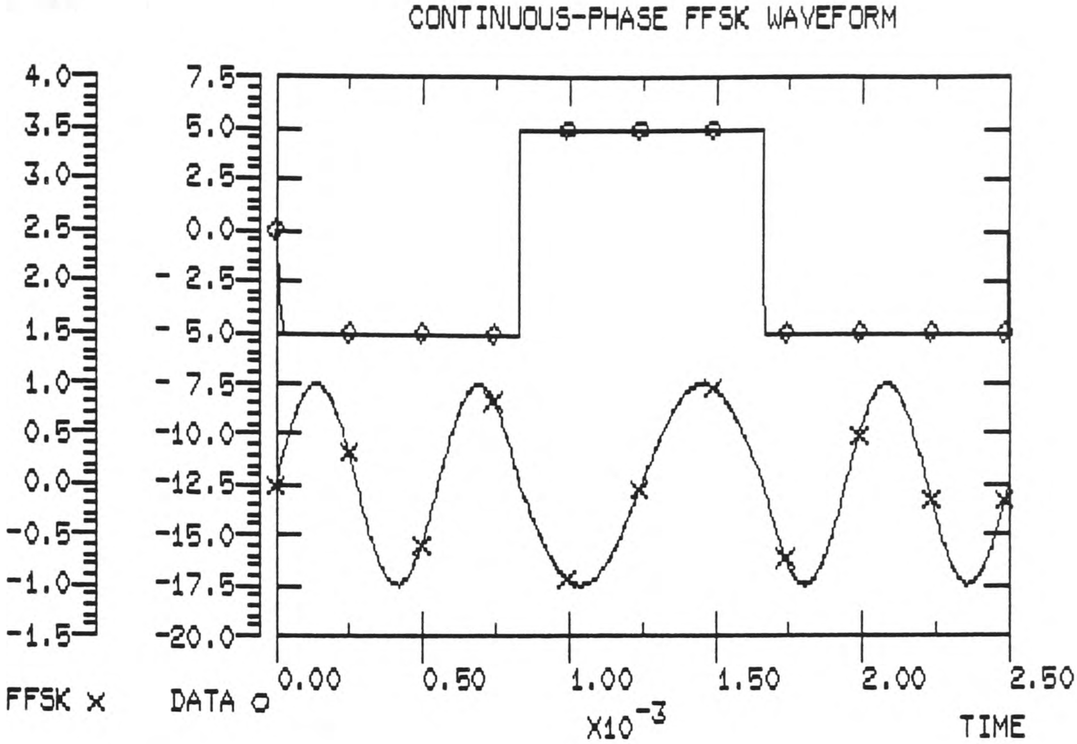
For 1200 baud:-

$T_b = 1/1200$, therefore $f_b = 1200$ Hz & $f_b/2 = 600$ Hz

Spectrum of a 1200 baud FFSK Data Transmission



(a) FFSK



(b) FSK

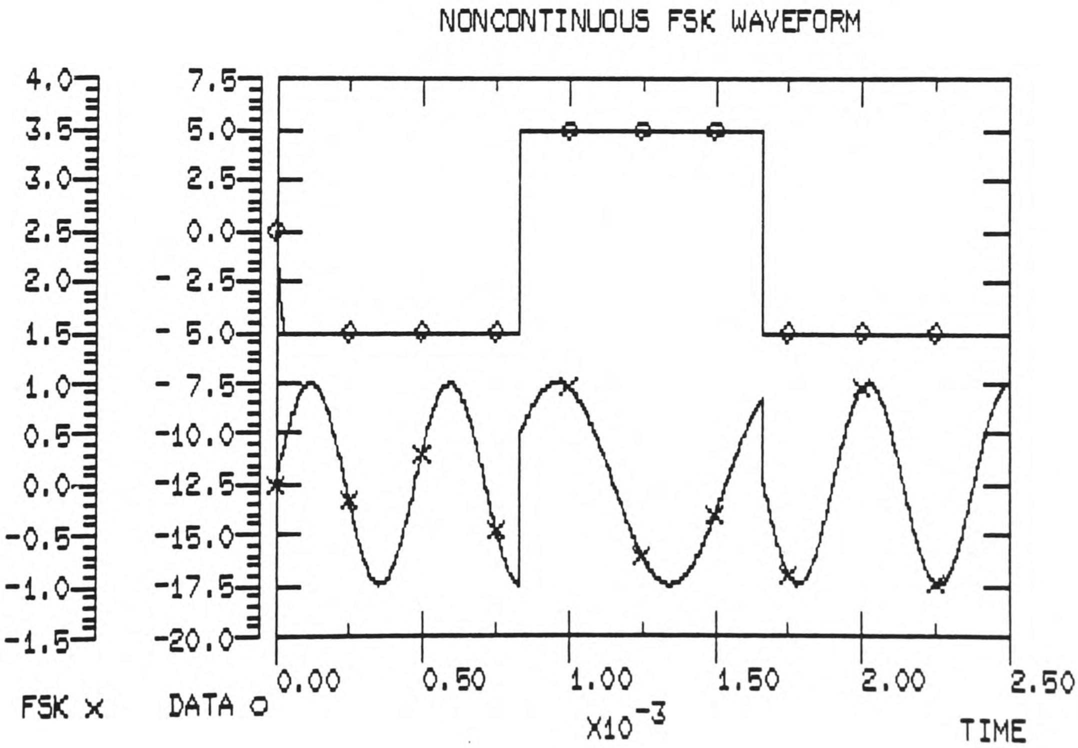


Figure F1 : Comparison of FSK and FFSK

(iv) Error Rate Performance of FFSK.

Consider the performance of a coherent FSK receiver:-

$$P(e) = 1/2\mathrm{erfc}\,[Po^2\,(Tb)/8\sigma_o^{\,2}]^{\frac{1}{2}} = \mathrm{Erfc}\,[Po^2\,(Tb)/4\,\sigma_o^{\,2}]^{\frac{1}{2}} \qquad F3$$

where: $$Po^2\,(Tb)/\sigma_o^{\,2} = 2/\eta\ \ (E_1 + Eo - 2E_{10}) \qquad F4$$

$$E_1 = Eo = A^2\,Tb/2 \quad \text{(Power x Bit period)} \qquad F5$$

$$E_{10} = \int_0^{Tb} fs(t).fm(t)dt \qquad F6$$

Which eventually leads to:-

$$\frac{Po^2\,(Tb)}{\sigma\ o^2} = \frac{2A^2\,Tb}{\eta}\left[1 - \frac{\sin 2\pi(2^\delta)Tb}{2\pi\,(2^\delta)Tb}\right]^{\frac{1}{2}} \qquad F7$$

Consequently, the probability of error is a function of the frequency shift between the two signalling states and hence the modulation index. However, for FFSK $\sin(2\pi(2\delta)Tb) = 0$, and therefore the term represented by Equation F6 is also zero. Under these conditions the two signalling waveforms form an "Orthogonal" set and the probability of error can be represented by Equation F8:-

$$P(e) = 1/2\mathrm{erfc}(E/2\eta)^{\frac{1}{2}} = \mathrm{Erfc}(E/\eta)^{\frac{1}{2}} \qquad F8$$

APPENDIX G

"A Summary of Important Results Derived from the Scattering Model"

(i) Introduction.

The scattering model,proposed by Clarke (33),assumes that the field incident on the mobile antenna is a "scattered field" composed of a number of randomly-phased plane waves.Clarke assumed that the plane waves are vertically polarised with azimuthal angles of arrival and the RF phase angles as random and statistically independent. Furthermore,the RF phase angles are assumed to have a uniform probability density function over 0 to 2π radians.This is plausible at VHF and above where the wavelength is small enough to ensure that small changes in path length result in appreciable change of the RF phase.

A statistical model,based on scattered waves,allows important relationships to be established for the received signals in the form of first and second order statistics of the signal envelope.

(ii) Behaviour of the received E field.

Consider a vertically polarised wave incident on a mobile antenna as illustrated in Figure G1.

The field components at the mobile are defined as:-

$$Ez = \sum_{i=1}^{N} ai\exp(-j\beta Vt\cos(\phi i - \alpha)) \qquad G1$$

$$Hx = \sum_{i=1}^{N} ai/\eta \sin\phi i\exp(-j\beta Vt\cos(\phi i - \alpha)) \qquad G2$$

$$Hy = -\sum_{i=1}^{N} ai/\eta \cos\phi i\exp(-j\beta Vt\cos(\phi i - \alpha)) \qquad G3$$

where: ϕi = The angle between the positive x axis and the direction of propagation.

α = The angle between the x axis and the direction of the actual vehicle.

ai = A complex variable representing the amplitude and phase of the received RF signal

η = Intrinsic wave impedance (120π)

Since ai is a complex variable it can be separated into real and imaginary parts:-

$$ai = Ri + jSi \qquad G4$$

Inserting Equation G4 into Equation Gl and separating into real and imaginary parts yields:-

$$Ez = X1 + jY1 \qquad G5$$

The real part of Ez is:-

$$X1 = \sum_{i=1}^{N} (Ri\cos\psi i + Si\sin\psi i) \qquad G6$$

The imaginary part of Ez is:-

$$Y1 = \sum_{i=1}^{N} (Si\cos\Psi i - Ri\sin\Psi i) \qquad G7$$

where: $\Psi i = \beta Vt\cos(\phi i - \alpha)$ G8

Since all N values of Ri,Si and ϕi are time independent,then from the "Central Limit Theorem" it follows that X1 and Y1 are independent random variables which are normally distributed as long as the value of N is greater than 10.

(iii) First order statistics of the received E field.

Equation G5 can be readily converted to polar co-ordinates:-

$$Ez = X1 + jY1 = r\exp(j\theta) \qquad G9$$

where: $r = \sqrt{X1^2 + Y1^2}$ G10

which corresponds to the amplitude of the received electrical field component.

and: $\theta = \mathrm{Tan}^{-1}(Y1/X1)$ G11

which corresponds to the phase of the received electrical field component.

Also,the real and imaginary parts of the received E field can be considered independent variables each with a Gaussian distribution, a variance σ^2, and zero mean. Therefore, the joint probability function $p(x,y)dxdy$ can be transformed to $p(r,\theta)drd\theta$:-

$$p(x,y)dxdy = p(r,\theta)drd\theta \qquad \text{G12}$$

From Figure G2:-

$$dxdy = rdrd\theta \qquad \text{G13}$$

$$p(r,\theta) = rp(x,y) \qquad \text{G14}$$

Now since x and y are independent then:-

$$p(x,y) = p(x).p(y) = \frac{1}{2\pi\sigma^2} \exp[-(x^2 + y^2)/2\sigma^2] \qquad \text{G15}$$

Therefore:-

$$p(r,\theta) = \frac{r}{2\pi\sigma^2} \exp(-r^2/2\sigma^2) \qquad \text{G16}$$

with: $p(r,\theta) = p(r).p(\theta)$ G17

since r and θ are independent variables with ($0 < r < \infty$) and ($0 < \theta < 2\pi$). Then:-

$$p(r) = r/\sigma^2 \exp(-r^2/2\sigma^2) \qquad \text{G18}$$

and: $p(\theta) = 1/2\pi$ G19

Equation G18 corresponds to a Rayleigh distribution as shown in Figure G3

The cumulative probability distribution function of the fading envelope $P(r \leqslant R)$ is given by:-

$$P(r \leqslant r_1) = \int_0^{r_1} p(r)dr = 1 - \exp(-r_1^2/2\sigma^2) \qquad G20$$

Table G1 Rayleigh-Fading-Signal Properties.

Parameters αi	Symbols	Values	dB over $2\sigma^2$	Percentage $p(re \leqslant \alpha i)$
Average power	$E(re^2)$	$2\sigma^2$	0	63
Mean value	$E(re)$	$\frac{\sqrt{\pi}}{2}\sqrt{2\sigma^2}$	-1.05	53
Median value	Mre	$0.832\sqrt{2\sigma^2}$	-1.59	50
Standard deviation	σre	$\frac{\sqrt{4-\pi}}{2}\sqrt{2\sigma^2}$	-6.68	-

(iv) Second order statistics of the received fading envelope

The "level-crossing rate" and the "average duration of a fade" are both second order statistics of the received fading envelope.For a specified signal level,say R,the level crossing rate for a crossing in the positive direction is given by:-

$$N(R) = \int_0^\infty \dot{r}.p(R,\dot{r})d\dot{r} \qquad \text{G21}$$

where: $p(R,\dot{r})$ is the joint p.d.f of R and $\dot{r}$

(The dot indicates the time derivative)

For a vertical monopole the level crossing rate becomes:-

$$N(R) = \sqrt{2\pi}\, fd\rho \exp(-\rho^2) \qquad \text{G22}$$

where: $\rho = R/Rrms$ and corresponds to the specified level relative to the rms amplitude of the fading envelope.

The normalised level crossing rates for a Rayleigh fading signal received by a vertical monopole is shown in Figure G4.The maximum level crossings occur at ρ = -3 dB and decreases as the level is lowered or increased.

The average duration of fades below a specified level R is defined as :-

$$\text{Average duration of fade} = \frac{\text{Probability } r \leqslant R \text{ in time interval } \tau}{\text{Level crossing rate } N(R)}$$

For a Rayleigh fading envelope,the probability $r \leqslant R$ is given by :-

$$P(r \leqslant R) = \int_0^R p(r)dr = 1 - \exp(-\rho^2) \qquad \text{G23}$$

Substituting for N(R) from Equation G22, the average duration of fade for a vertical monopole is given by:-

$$\bar{\tau}(R) = \frac{\exp(\rho^2) - 1}{\rho \, fd \sqrt{2\pi}}$$

This expression is plotted in Figure G5 which shows that as the ratio ρ is decreased the average duration of the fades decreases since the signal remains above the level R. However, as the level is increased, a point is reached where the signal lies below the level R most of the time and $\bar{\tau}(R)$ can be very long.

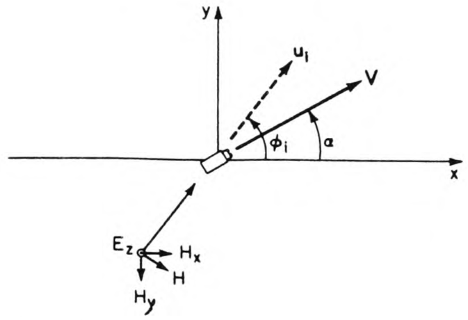


Figure G1 : Short term fading model

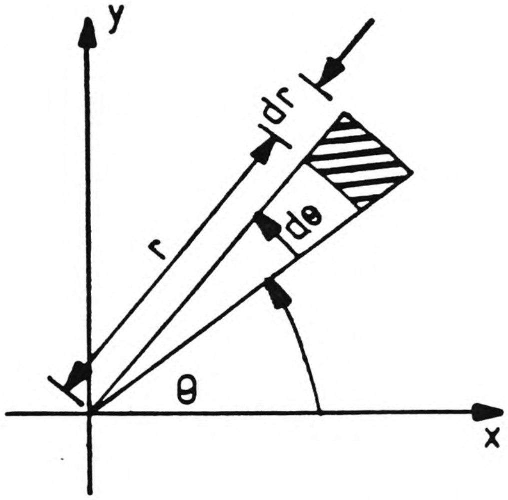


Figure G2 : Gaussian to Rayleigh transformation

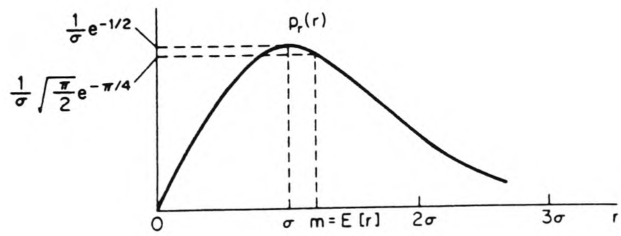


Figure G3 : Rayleigh distribution of the variate r

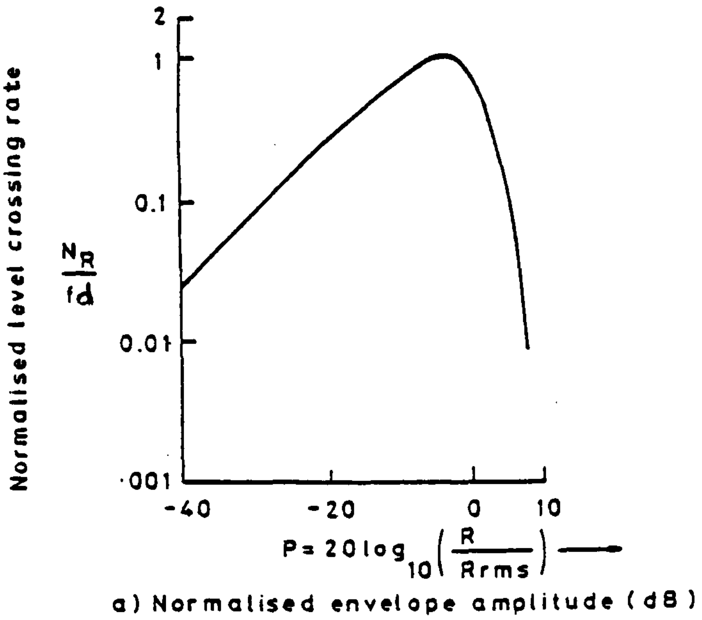


Figure G4 : Level crossing rate

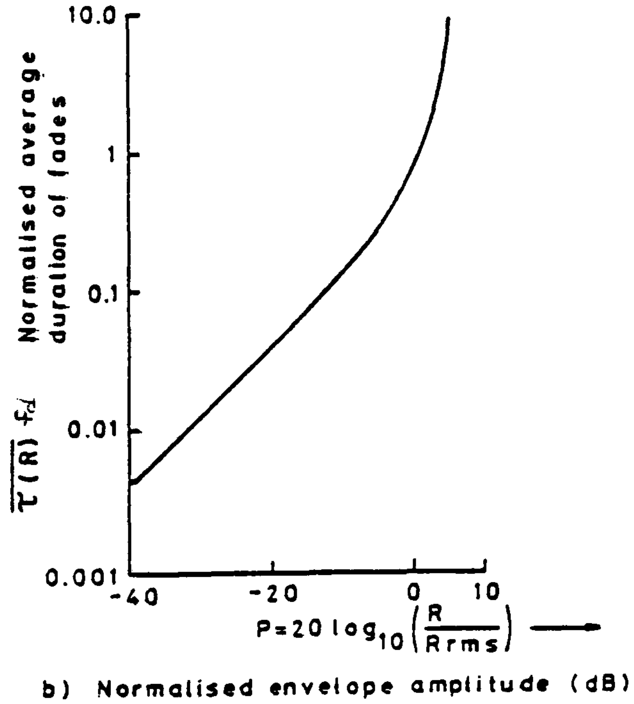
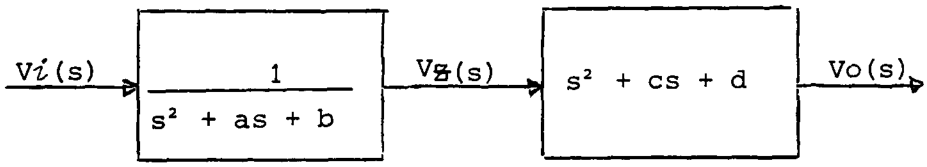


Figure G5 : Duration of fades

(i) Transfer Function Modelling Using Astec 3.

Many electronic configurations, such as filters, are conveniently described in terms of a transfer function. In order to perform a computer simulation of such networks with Astec 3 it is necessary to decompose the transfer function into a set of simultaneous first order differential equations. This technique is best illustrated with the aid of a suitable example. Consider a generalised second order transfer function:-

$$H(s) = \frac{Vo(s)}{Vi(s)} = \frac{s^2 + cs + d}{s^2 + as + b}$$



where $$V_z(s) = \frac{Vi(s)}{s^2 + as + b}$$

Therefore

$$S^2 V_z(s) + as\, V_z(s) + bV_z(s) = Vi(s)$$

This expression has the following equivalent time domain form:-

$$V_z'' = Vi - a\, V_z' - b\, V_z$$

Let $V_Z = X_1$

And $V_Z' = X_2 = X_1'$

Hence $V_Z'' = X_2' = Vi - aX_2 - bX_1$ H1

Now $Vo(s) = s^2 \ V_Z(s) + csV_Z(s) + dV_Z(s)$

Converting to its equivalent time domain form:-

$$Vo = V_Z'' + cV_Z' + dV_Z \qquad \text{H2}$$

Substituting Equation H.1 into Equation H.2:-

$$Vo = X_2' + cX_2 + dX_1$$

$$= Vi + (c - a)X_2 + (d - b)X_1$$

The relationship between the output and input voltages can now be expressed in a form suitable for Astec 3 simulation:-

```
PVOUT=UIN+(PC-PA)*PX2+(PD-PB)*PX1;
PX1' =PX2;
PX2' =UIN-PA*PX2-PB*PX1;
```

Where the coefficients PA,PB,PC and PD could be supplied as formal arguments to a model containing this set of first order differential equations.In this way,any type of second order filter could be implemented.

The following Astec 3 model corresponds to a second order lowpass filter which has the transfer function:-

$$H(s)_{LPF} = \frac{Vo(s)}{Vi(s)} = \frac{1}{\left(\frac{s}{\omega o}\right)^2 + K\left(\frac{s}{\omega o}\right) + 1}$$

Where ωo is the angular cut-off frequency of the filter:-

```
!MODEL LPF2(IN-OUT-COM):
PF2;
PK;
PI=4*FARCTAN(1.0);
POMEGA=2*PI*PF2;
PA=PK*POMEGA;
PB=POMEGA*POMEGA;
UIN(IN-COM);
PX1´ =PX2;
PX2´ =UIN-PA*PX2-PB*PX1;
PVOUT=PB*PX1;
EOUT(OUT-COM)PVOUT;
```

The formal argument PF2 corresponds to the cut-off frequency of the lowpass filter,whereas PK is related to the filter "Q" factor.

(ii) User Defined Fortran Functions.

Another particular aspect of Astec 3 worthy of reference is the facility of supplying up to ten user defined Fortran function sub-routines.When these supplied routines are made known to Astec 3 at the start of a run they can be used as though they were members of the built in set of Fortran functions.A Fortran function can be allocated to branches,local and global parameters,or on the right hand side of differential equations in a !GLOBAL:,!CIRCUIT: or a !MODEL: sequence.The format for user defined functions is as follows:-

Function Name(argument,....,argument)

The following restrictions are placed on the entry of Fortran functions:-

- For each argument which is directly or indirectly a function of an electrical variable or a parameter defined by a differential equation,the user must provide,within the function,statements which calculate the partial derivative of the function with respect to the argument.Arguments which are constants do not need this.

- When partial derivatives are to be calculated an array type REAL*8 and DIMENSION(1) must be declared.This array must be added to the arguments as the last one at the right hand end.It is used to convey the partial derivatives to Astec 3.The partial derivative of the nth argument must be assigned to the nth position in the array.

- The partial derivative affects the convergence properties of the numerical routines and not the accuracy.

The following examples illustrate the role of user defined Fortran functions.The first example calculates the sum of four arguments:-

```
FUNCTION FSUM(A1,A2,A3,A4)
IMPLICIT REAL*8(A-Z)
FSUM=A1+A2+A3+A4
RETURN
END
```

Fsum can be used in the $DESC command as follows:-

```
PX2=2;
E1(3-1)FSUM(1.5,PX,E4,&T);
E4(A-B)10;
```

The arguments must not refer to any electrical variable or differential parameter.

This next example shows a way of writing a function to allow the arguments supplied when the function is used to refer to electrical variables or differential parameters:-

```
FUNCTION FMEAN(A1,A2,DF)
IMPLICIT REAL*8(A-Z)
DIMENSION DF(1)
FMEAN=0.5*(A1+A2)
DF(1)=0.5
DF(2)=0.5
RETURN
END
```

Arguments A1 and A2 may now refer to electrical variables or differential parameters.i.e.:-

```
EOUT(OUT-GND)FMEAN(VRY,PY);
PY=FLOG(VR4/VR2);
```

APPENDIX I

"A Selection of User-Defined Astec 3 Functional Models"

(i) Signal Sources

(a) QUADOS

```
!MODEL QUADOS(SINOUT-COSOUT-COM):
PA;
PFREQ;
PERROR;
PI=4*FARCTAN(1.0);
POMEGA=2*PI*PFREQ;
PHI=(PERROR/180)*PI;
ESIN(SINOUT-COM)PA*FSIN(POMEGA*&T);
ECOS(COSOUT-COM)PA*FCOS((POMEGA*&T)+PHI);
```

QUADOS performs the function of a quadrature oscillator providing sinusoidal and cosinusoidal output waveforms.Local arguments PA and PFREQ determine the amplitude and frequency of the output waveforms produced during a transient simulation.The actual quadrature relationship between the output waveforms can be affected by the local argument PERROR.

(b) SINXOS

```
!MODEL SINXOS(OUT-COM):
PA;
PFUND;
PI=4*FARCTAN(1.0);
POMEGA=2*PI*PFUND;
P1=0.3*PA*FSIN(POMEGA*&T);
P2=0.64*PA*FCOS(2*POMEGA*&T);
P3=0.9*PA*FSIN(3*POMEGA*&T);
P4=1*PA*FCOS(4*POMEGA*&T);
P5=0.9*PA*FSIN(5*POMEGA*&T);
P6=0.64*PA*FCOS(6*POMEGA*&T);
P7=0.3*PA*FSIN(7*POMEGA*&T);
EOUT(OUT-COM)P1+P2+P3+P4+P5+P6+P7;
```

SINXOS produces a complex periodic signal which consists of a fundamental and harmonic components up to and including the seventh.The envelope of the discrete spectra of this waveform follows a sinc(x) distribution with zeroes at d.c. and the eighth harmonic.The amplitude and frequency of the fundamental can be defined by the local arguments PA and PFUND.

(c) DISTOS

```
!MODEL DISTOS(OUT-COM):
PA;
PFUND;
PI=4*FARCTAN(1.0);
POMEGA=2*PI*PFUND;
P3=PA*FSIN(3*POMEGA*&T);
P5=PA*FCOS(5*POMEGA*&T);
P7=PA*FSIN(7*POMEGA*&T);
P11=PA*FCOS(11*POMEGA*&T);
EOUT(OUT-COM)P3+P5+P7+P11;
```

Similar to SINXOS,this model generates a complex periodic signal with a well defined discrete spectrum.The actual spectrum consists of four distinct and equal amplitude frequency components,which are odd and prime number multiples of the fundamental,which is itself excluded.Local arguments PA determines the amplitude of each spectral component,whereas,PFUND determines the fundamental frequency.

(ii) Coupling Networks.

(a) RCNET

```
!MODEL RCNET(IN-OUT-COM):
PCAP;
PRES;
UIN(IN-COM);
CIN(IN-OUT)PCAP;
ROUT(OUT-COM)PRES;
```

This model corresponds to a straightforward CR coupling network.The relevant component values can be determined by PCAP and PRES respectively.

(b) IFTDT1

```
!MODEL IFTDT1(PR1-PR2-SEC1-SEC2-COM):
PLP;PCP;PRP;
PLS;PCS;PRS;
PCPS;PCSS;
PK;
PHP;PHS;
PM=PK*FSQRT(PLP*PLS);
CP(PR1-PR2)PCP;
RP(PR1-PR3)PRP;
HP(PR3-PR2:PHP)PLP*IHP+PM*IHS;
HS(SEC3-SEC2:PHS)PLS*IHS+PM*IHP;
CS(SEC1-SEC2)PCS;
RS(SEC1-SEC3)PRS;
CPS(PR2-COM)PCPS;
CSS(SEC2-COM)PCSS;
```

IFTD1 is a double tuned transformer coupled network with all component and other relevant values including Q,coupling coefficient and phasing of the windings determined by a comprehensive set of local arguments.Unfortunately,mutual inductance is not modelled directly by Astec 3.However, one solution is to employ "type H" elements (16).

(iii) Filters.

(a) HPF2

```
!MODEL HPF2(IN-OUT-COM):
PF2;
PK;
PI=4*FARCTAN(1.0);
POMEGA=2*PI*PF2;
PA=PK*POMEGA;
PB=POMEGA*POMEGA;
UIN(IN-COM);
PX1´ =PX2;
PX2´ =UIN-PA*PX2-PB*PX1;
PVOUT=PX2´ ;
EOUT(OUT-COM)PVOUT;
```

HPF2 is a generalised second order high pass filter with the transfer function defined below:-

$$H(s) = \frac{s^2}{s^2 + as + b}$$

The characteristics of the filter,in terms of cut-off frequency and response type,are determined by PF and PK respectively.

(b) BPF

```
!MODEL BPF(IN-OUT-COM):
PF0;
PBW;
PA=PBW;
PI=4*FARCTAN(1.0);
POMEGA=2*PI*PF0;
PB=POMEGA*POMEGA;
UIN(IN-COM);
PX1´ =PX2;
PX2´ =UIN-PA*PX2-PB*PX1;
PVOUT=PA*PX2;
EOUT(OUT-COM)PVOUT;
```

This model corresponds to a band pass filter with a transfer function given by the following equation:-

$$H(s) = \frac{cs}{s^2 + as + b}$$

The centre frequency and passband can be determined by PFO and PBW.

(c) PHLAG

```
!MODEL PHLAG(IN-OUT-COM):
PFREQ;
PHASE;
PI=4*FARCTAN(1.0);
POMEGA=2*PI*PFREQ;
PHRADS=(PHASE/180)*PI;
PHI=PHRADS/2;
PA=(POMEGA)/(FTAN(PHI));
UIN(IN-COM);
PX1´ =UIN-PA*PX1;
PVOUT=PA*PX1-PX1´ ;
EOUT(OUT-COM)PVOUT;
```

PHLAG corresponds to an "All-pass" filter having a transfer function of the form:-

$$H(s) = \frac{a - s}{a + s}$$

Having a flat frequency response this model can be used to introduce a particular value of phase shift at any spot frequency of interest.

(iv) Angle Modulators.

(a) NBPM

```
!MODEL NBPM(IN-OUT-COM):
M1 (IN-2-4-COM)MIXER:PKMIX=1;
M2 (2-3-COM)QUADOS:PA=1,PFREQ=100K,PERROR=0;
M3 (4-5-COM)PREAMP:PAMP=0.1;
M4 (5-3-OUT-COM)ADDER;
```

NBPM corresponds to a functional model of a low deviation phase modulator ($\beta = 0.1$) operating at a carrier frequency of 100 kHz.As indicated,this model relies on other user defined models,namely,MIXER;QUADOS;PREAMP and ADDER which are "nested" within the model NBPM.

(b) QNBPM

```
!MODEL QNBPM(IN1-IN2-OUT-COM):
M1 (2-3-COM)QUADOS:PA=1,PFREQ=100K,PERROR=0;
M2 (IN1-2-4-COM)MIXER:PKMIX=1;
M3 (IN2-3-5-COM)MIXER:PKMIX=1;
M4 (4-6-COM)PREAMP:PAMP=0.1;
M5 (5-7-COM)PREAMP:PAMP=0.1;
M6 (3-6-8-COM)ADDER;
M7 (2-7-9-COM)ADDER;
M8 (8-9-OUT-COM)ADDER;
```

QNBPM is another functional model which employs nesting of other user defined models.Its function is to produce two narrow band angle modulated signals in phase quadrature.As with NBPM,the carrier frequency is 100 kHz with the maximum modulation index normalised to 0.1.

(v) Digital Circuits.

(a) CLOCK

```
!MODEL CLOCK(OUT-COM):
PLOW;PHIGH;
PSTART;PWIDTH;PERIOD;
PRISE;PFALL;
PEOUT=FPULSE(PLOW,PHIGH,PSTART,PRISE,PWIDTH,PFALL,
             PERIOD);
EOUT(OUT-COM)PEOUT;
```

This model is based on a "built in" Fortran function FPULSE which can be used to produce a periodic pulse waveform.The high and low states,mark-space ratio and rise and fall times can all be determined by setting appropriate local arguments.

(b) XORGAT

```
!MODEL XORGAT(INA-INB-OUT-COM):
UINA(INA-COM);
UINB(INB-COM);
IF (((UINA.GT.0).AND.(UINB.LE.0)).OR.((UINA.LE.0)
                                   .AND.(UINB.GT.0)))
THEN PEOUT=1
ELSE PEOUT=-1;
EOUT(OUT-COM)PEOUT;
```

As its name suggests,this model performs the function of an Exclusive-OR gate on digital signals presented to its two inputs.

(c) DIGDEL

```
!MODEL DIGDEL(IN-OUT-COM):
PTOR;
PLOW;
UIN(IN-COM);
PEOUT=FDELAY(UIN,PTOR);
IF (POUT.LE.0.0) THEN PEOUT=PLOW
ELSE PEOUT=POUT;
EOUT(OUT-COM)PEOUT;
```

This functional model also relies on a built in Fortran function FDELAY to achieve a programmable digital delay network. DIGDEL can be used in such applications as implementing a shift register.

(d) PRBSC

```
!MODEL PRBSC(IN-OUT-COM):
M1 (IN-13-OUT-COM)ORGATE;
M2 (OUT-3-COM)DIGDEL:PTOR=1ML,PLOW=-1;
M3 (3-4-COM)DIGDEL:PTOR=1ML,PLOW=-1;
M4 (4-5-COM)DIGDEL:PTOR=1ML,PLOW=-1;
M5 (5-6-COM)DIGDEL:PTOR=1ML,PLOW=-1;
M6 (6-7-COM)DIGDEL:PTOR=1ML,PLOW=-1;
M7 (7-8-COM)DIGDEL:PTOR=1ML,PLOW=-1;
M8 (8-9-COM)DIGDEL:PTOR=1ML,PLOW=-1;
M9 (9-10-COM)DIGDEL:PTOR=1ML,PLOW=-1;
M10 (10-11-COM)DIGDEL:PTOR=1ML,PLOW=-1;
M11 (11-12-COM)DIGDEL:PTOR=1ML,PLOW=-1;
M12 (9-12-13-COM)XORGAT;
```

PRBSC is a pseudo binary sequence generator which is based on the DIGDEL delay network to form a 10 stage shift register. With Exclusive-OR feedback taken from relevant taps in the shift register network a maximal code length sequence of 1023 is achieved.

APPENDIX J

"A Selection of Circuits for the QNBFAM Transmitter"

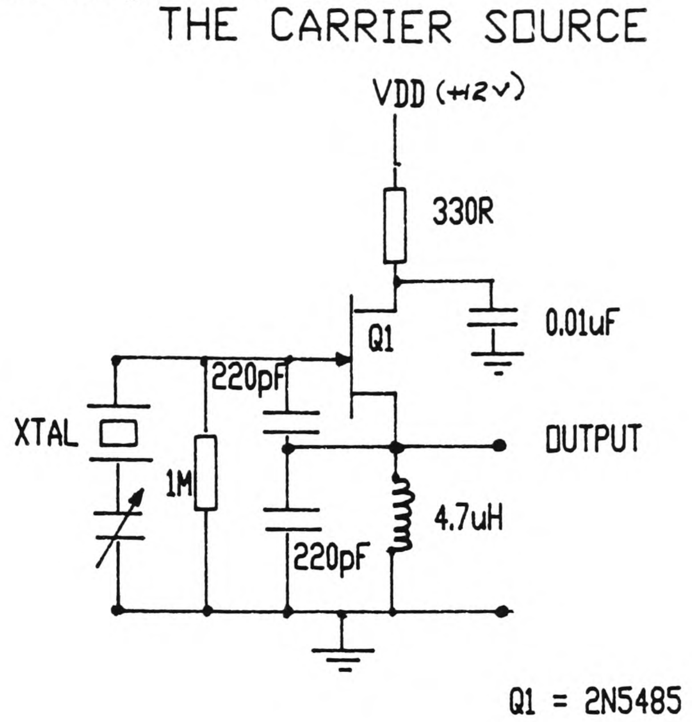


Figure J1

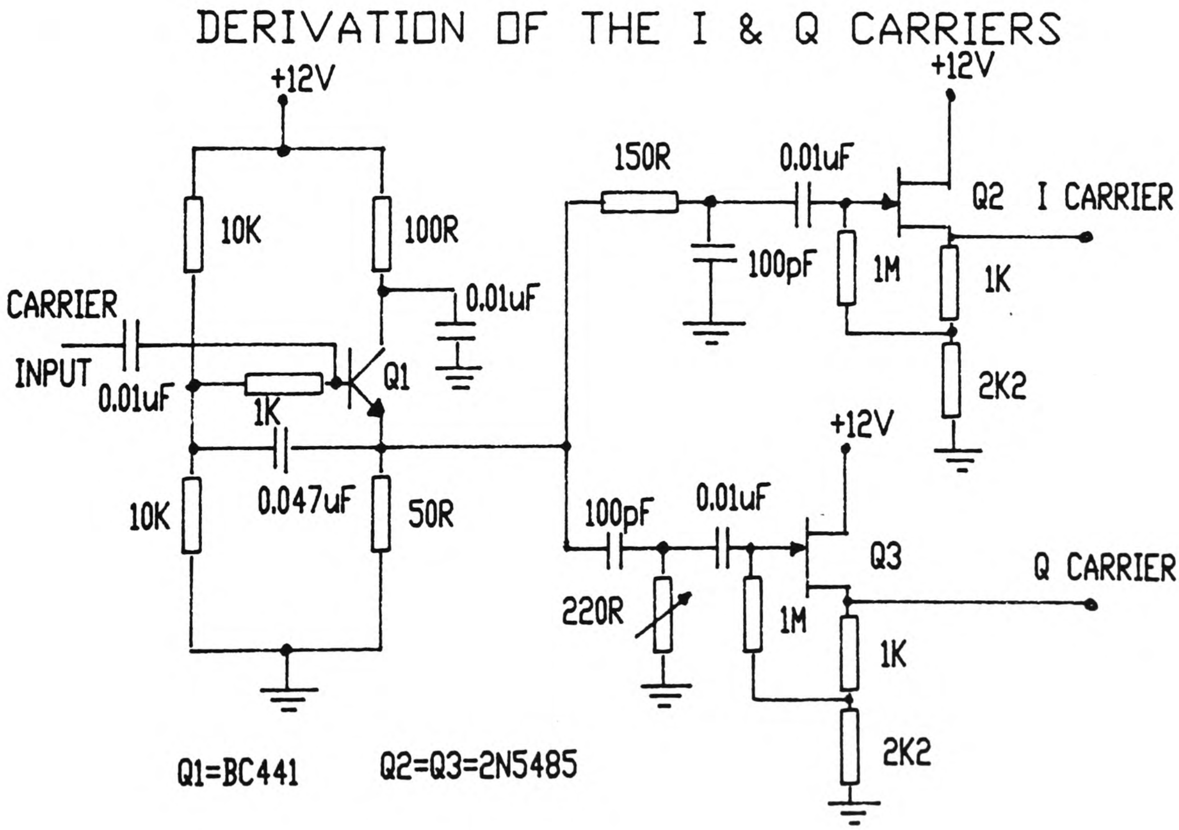


Figure J2

NARROWBAND ANGLE MODULATOR

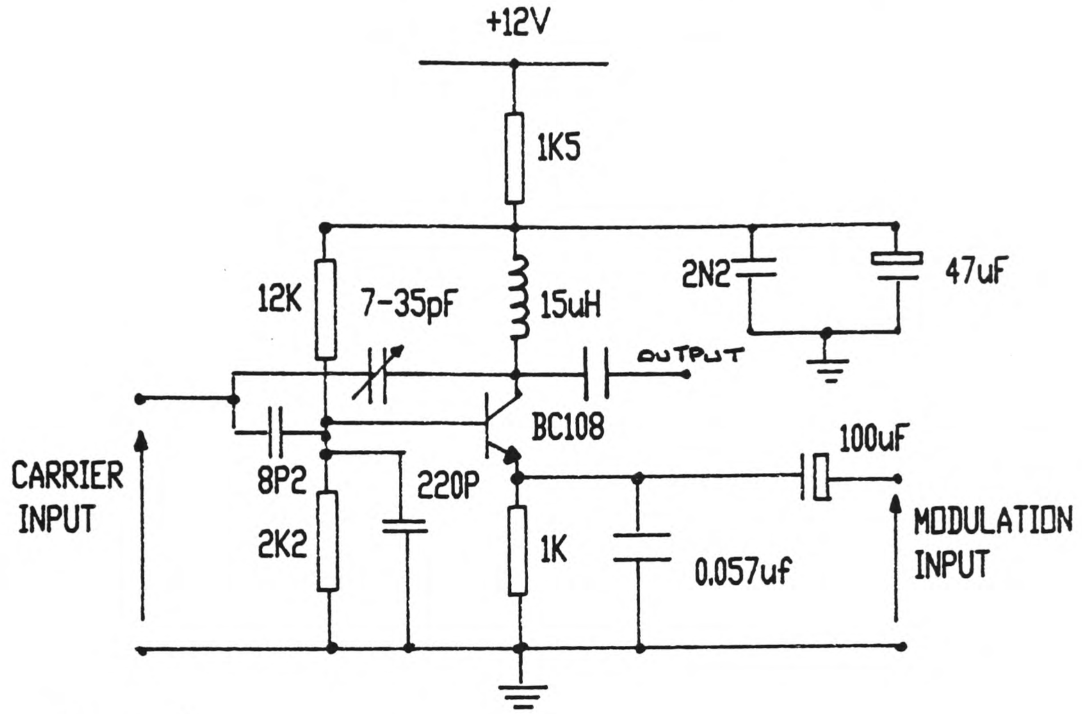


Figure J3

NB ANGLE MODULATOR AMPLITUDE LIMITER

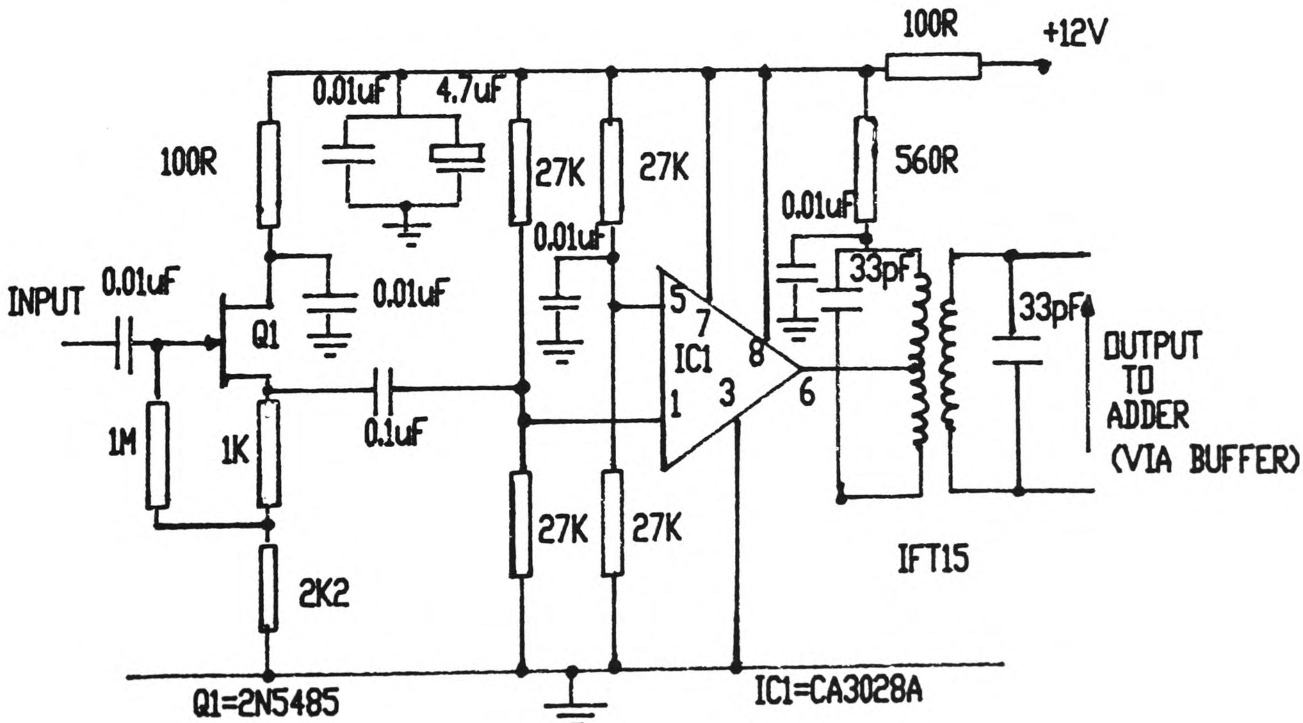


Figure J4

AMPLITUDE LIMITER BUFFER CIRCUIT

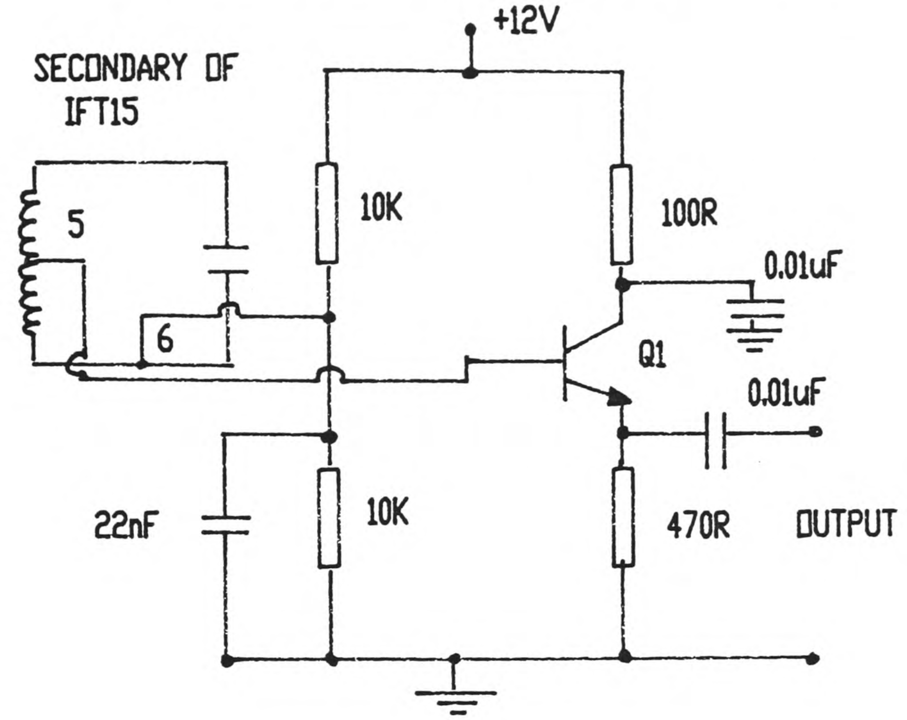


Figure J5

THE COMMON BASE SUMMING AMPLIFIER

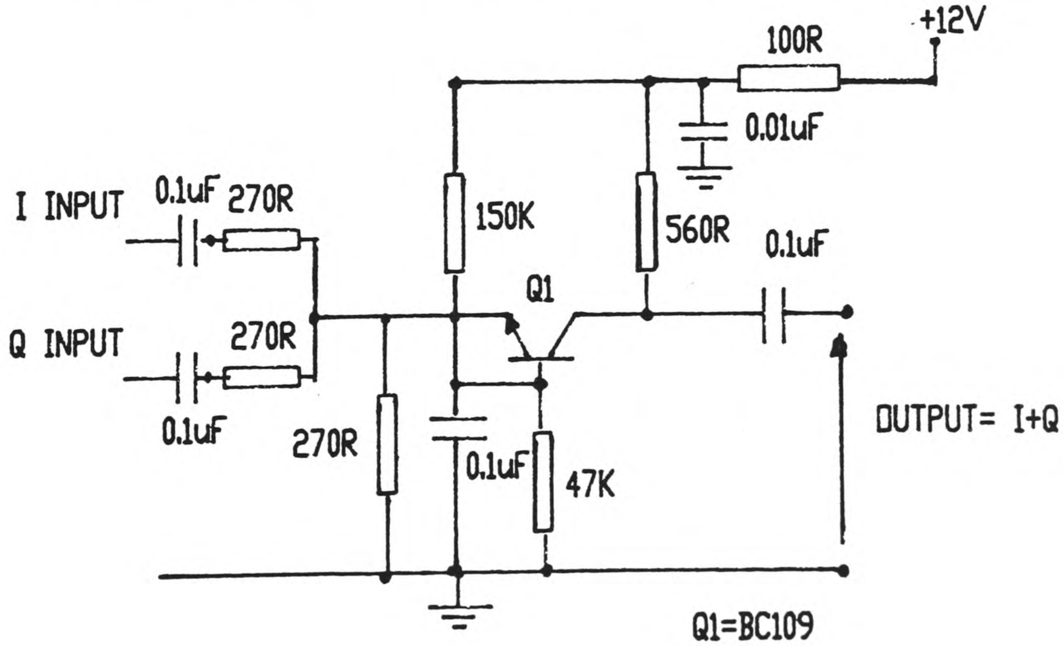
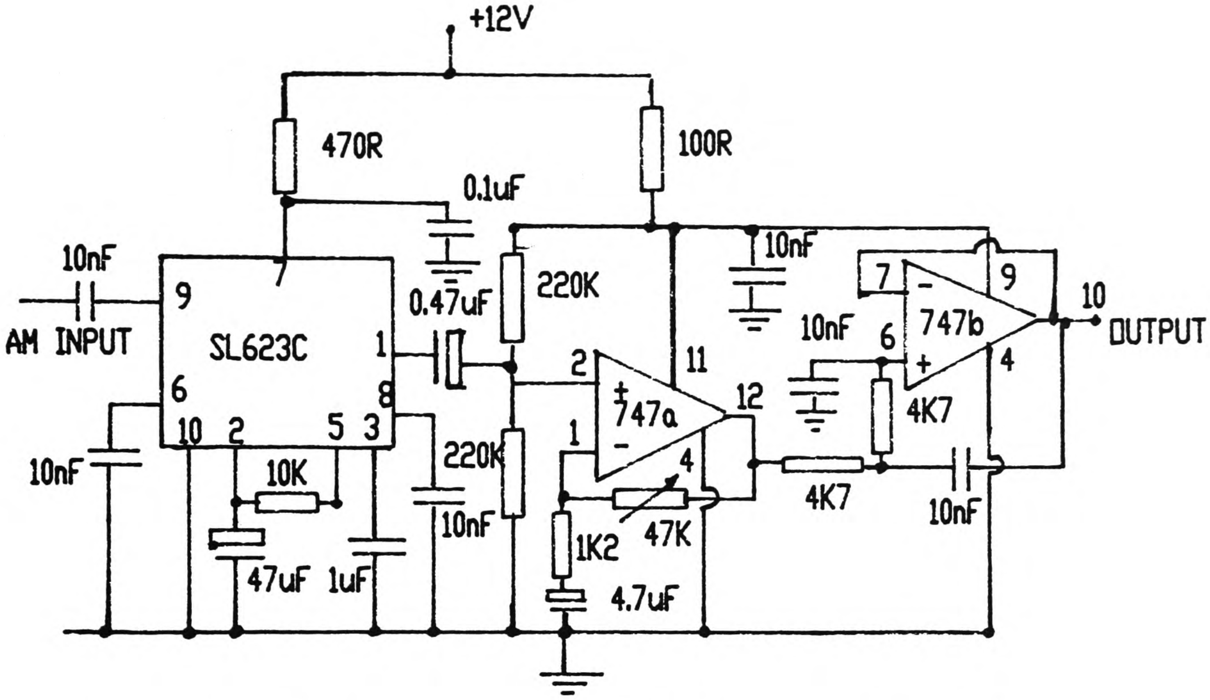


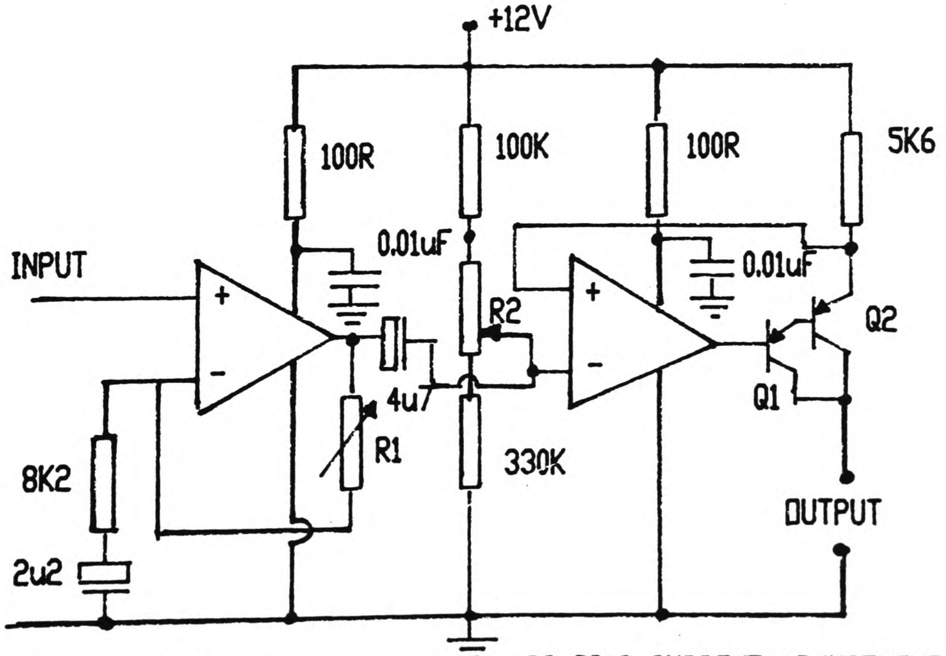
Figure J6



AM DETECTOR : AMPLIFIER : FILTER

Figure J7

AMPLITUDE MODULATOR CONTROL CIRCUIT



R1=MODULATION DEPTH

R2=BIAS CURRENT ADJUSTMENT

Figure J8

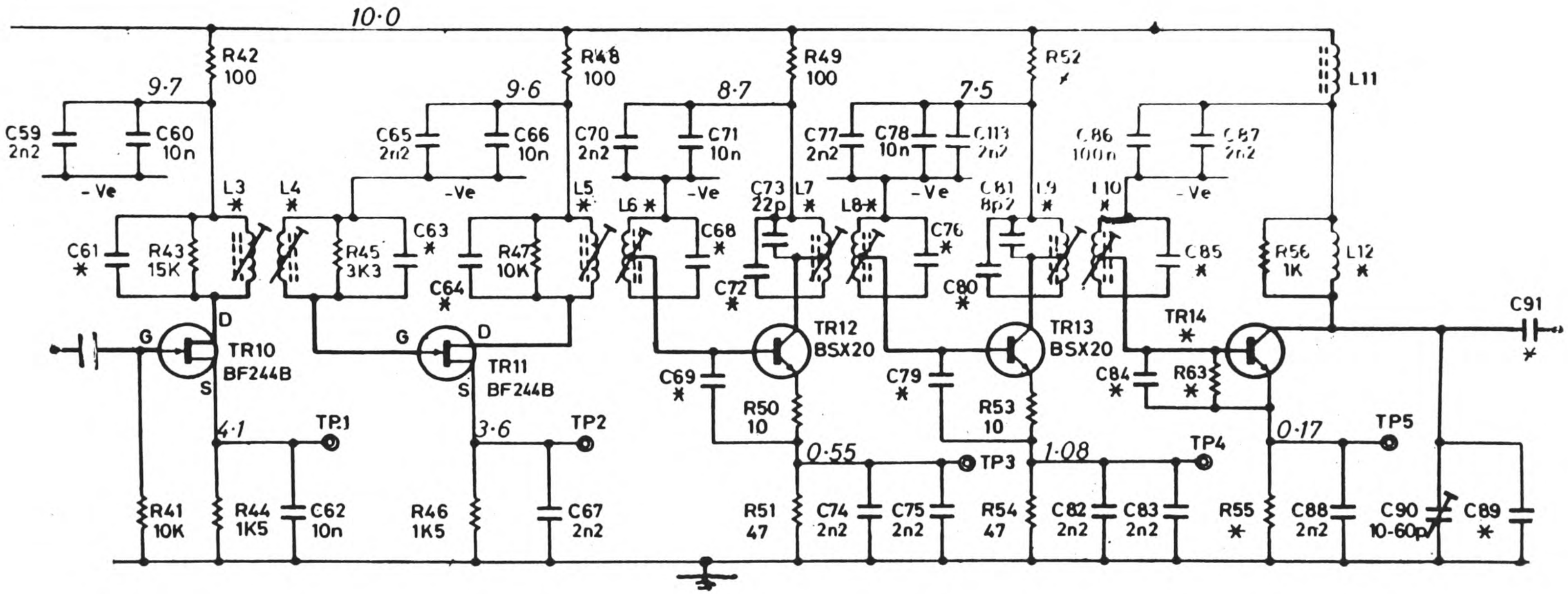


Figure J9 : The x16 Frequency Multiplier (PYE Europe)