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COURSE ON BASIC TELECOMMUNICATIONS SCIENCE

9 January - 3 February 1989

Digital Radio: Theory and Measurements

Hewlett-Packard, Queensferry Telecom Division

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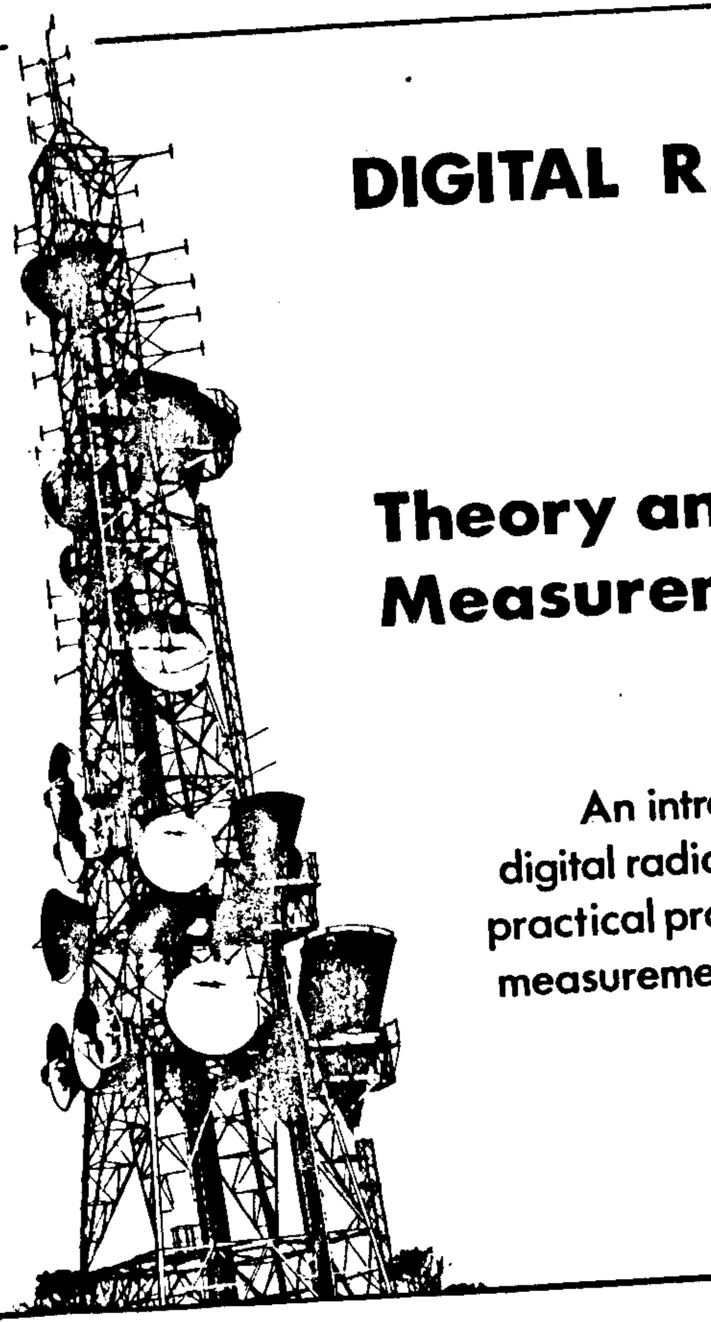

DIGITAL RADIO

Theory and Measurements

An introduction to digital radio principles, practical problems and measurement solutions

HEWLETT PACKARD
QUEENSFERRY TELECOM DIVISION

Application Note 355

SECOND-EDITION

This booklet replaces the two earlier seminar booklets (5954-2048 and 5954-2049). The main differences are in the second section on impairments and measurements. The changes include: updates to the error-performance standards, more comprehensive treatment of transmission impairments and the inclusion of the latest Hewlett-Packard measurement solutions and publications for Digital Radio. There has also been some editing to clarify certain points.

This booklet is divided into two parts. The first part provides an introduction to digital radio concepts and implementation. The second part reviews the impairments that may occur on a practical digital radio system and how these degradations affect the performance of a radio link. Measurement solutions to these problems are presented. The booklet is laid out as a slide storyboard so that it can be used in presentations, however the text provides sufficient information for it to be used as a normal application note.

A NOTE TO THE PRESENTER:

In the second part of the seminar, as each measurement type is described, it is useful to demonstrate the appropriate Hewlett-Packard test equipment. This is indicated throughout the booklet by a box of text at the foot of the page similar to this. In some cases a unique piece of dedicated test equipment can be presented, in other cases a range of products is available. Examples of these are: spectrum analyzers, network analyzers and bit-error rate testers. These product families are continually evolving and, in the case of BER testers, different interface standards and bit-rates are used in various parts of the world. With a knowledge of the audience, select some appropriate products based on the applications information in the booklet, and include one or two product slides to support your demonstration; product slides can usually be obtained from the manufacturing division.

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PART 1

THEORY AND IMPLEMENTATION

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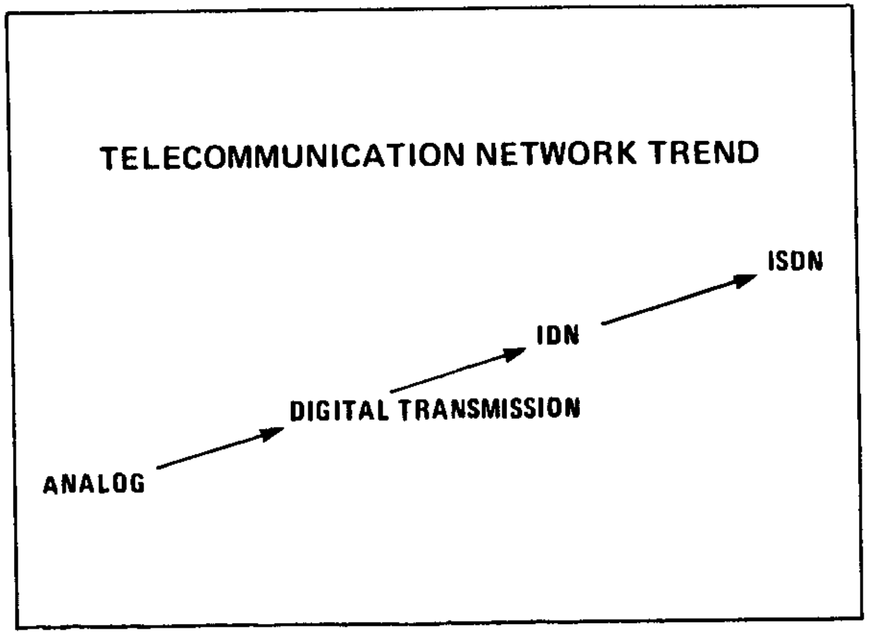
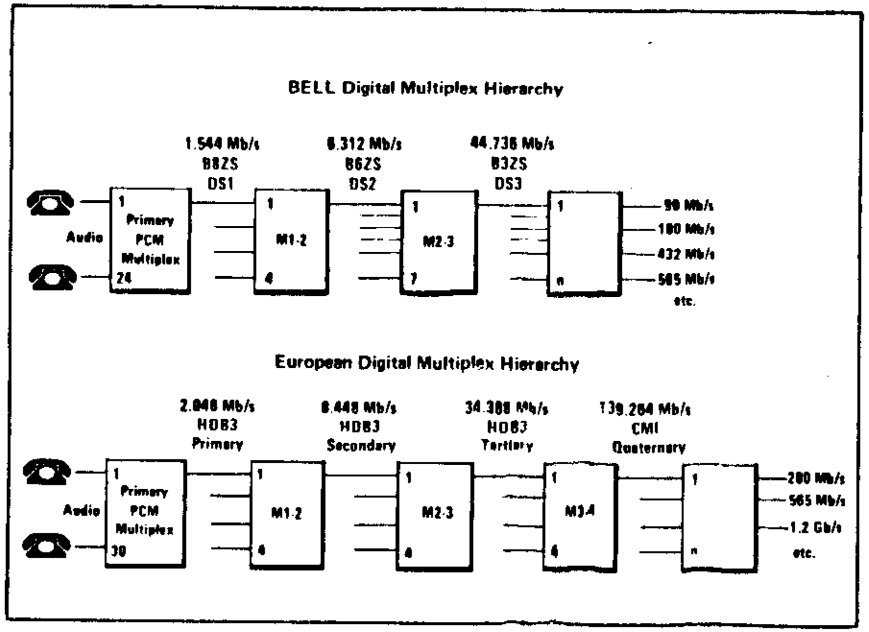
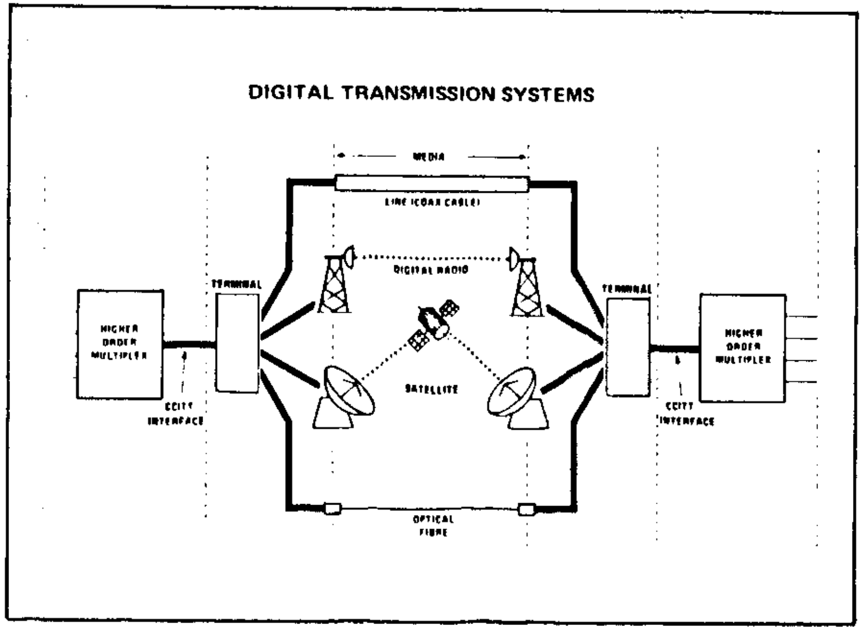
In the first part of this presentation we will consider the concepts of digital radio transmission and consider the main building blocks of the digital radio. We will look in some detail at the items listed on this slide and we will start by reviewing the role of digital radio in the Integrated Digital Network.



Telecommunications networks worldwide are currently converting (at varying rates) from analog to digital systems with the ultimate objective being the creation of Integrated Digital Networks (ie: networks with connections provided via digital switching/transmission) and Integrated Services Digital Networks (ie: multi-service networks with digital access, switching and transmission).



The various digital services, whether they be digitized telephony (64kb/s PCM or 32kb/s ADPCM), data, videotex, facsimile etc. are time division multiplexed (TDM) together to form higher rate bit streams. This is done in stages as shown in this slide of the North American and European digital hierarchies. These are the most commonly used hierarchy rates, different hierarchies are used in Japan and in some military systems. The output of each multiplex stage may form the tributary stream for the next stage of multiplexing in a high-capacity system, or may pass directly to the transmission system in a lower-capacity route.

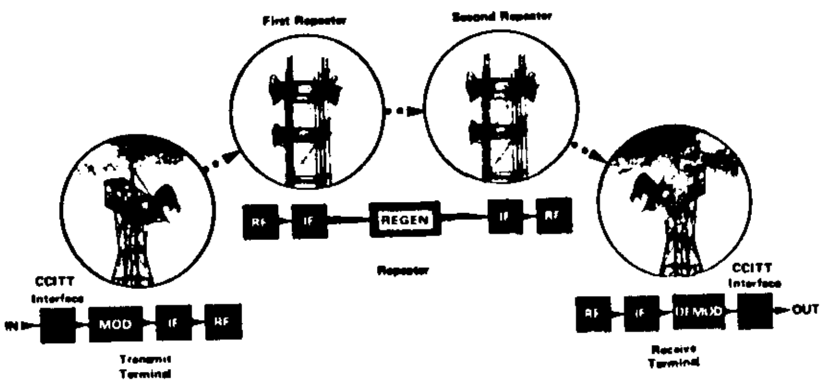


The bit rates and interface codes etc. shown in the previous slide are all standardized by CCITT (International Telephone and Telegraph Consultative Committee) and are independent of the particular transmission medium used. The transmission system may carry traffic at any of the bit-rates in the hierarchy, depending on the capacity through-put requirements of the system. The testing and the performance of the system at these interfaces relates to network performance in the IDN and again is specified by CCITT (Rec. G821, G703 etc) independent of the transmission media. These standards have been adopted by CEPT in Europe and by the ANSI/ECSA* T1 Committee in North America.

This slide shows the four commonly used methods of transmission. Optical fiber is the most popular for high-capacity routes in Network Operators (PTT's, Telcos and Common Carriers) where existing routes or "way-leaves" exist. However microwave radio and satellite have many applications in lower capacity routes, in difficult terrain and in private and military communication networks where the advantages of flexibility, security and speed of installation offered by radio are particularly valuable.

* *Exchange Carriers Standards Association T1 Committee of the American National Standards Institute.*

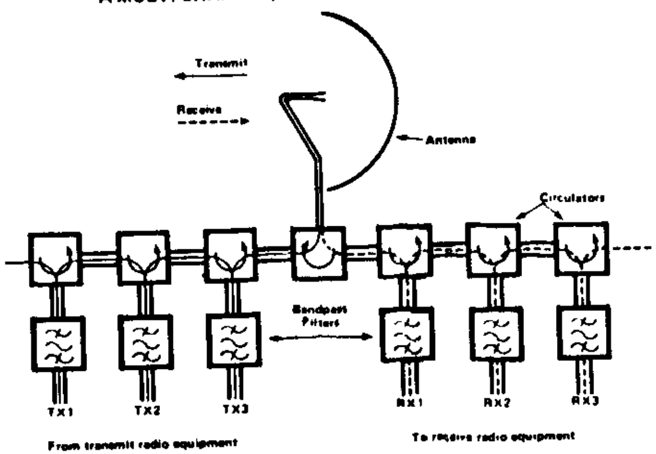
A MULTI HOP, ONE WAY MICROWAVE LINK



A practical radio relay system often consists of several hops as the maximum distance between transmit and receive antennas or "hop length" is normally 30-60km (20-40 miles) in a line-of-sight system. The intermediate stations are called **repeater stations** and the traffic data stream may not necessarily be brought down to the CCITT interface at these points, but simply regenerated at the binary level. Some radios use a direct IF repeater without regeneration. This saves cost, but some of the benefits of digital transmission are lost because of the build up of noise and distortion in a similar way to analog radio systems.

The microwave frequency bands and the radio channel spacing in these bands have all been standardized by CCIR (International Radio Consulative Committee) and FCC in North America. Some typical frequency bands are 2Ghz (used for lower capacity), 4GHz, 6GHz, 7GHz, 8GHz, 11GHz and 13GHz. Above 11GHz rain attenuation becomes a greater problem necessitating a shorter hop length for a given system availability. There is a new generation of radios becoming available, operating in the range 15-50GHz which provides low and high capacity short-haul links in cities for interconnecting business centres with main transmission centres. The small physical size of antennas at these frequencies, makes this type of link very easy to install.

A MULTI CHANNEL, TWO WAY ANTENNA SYSTEM



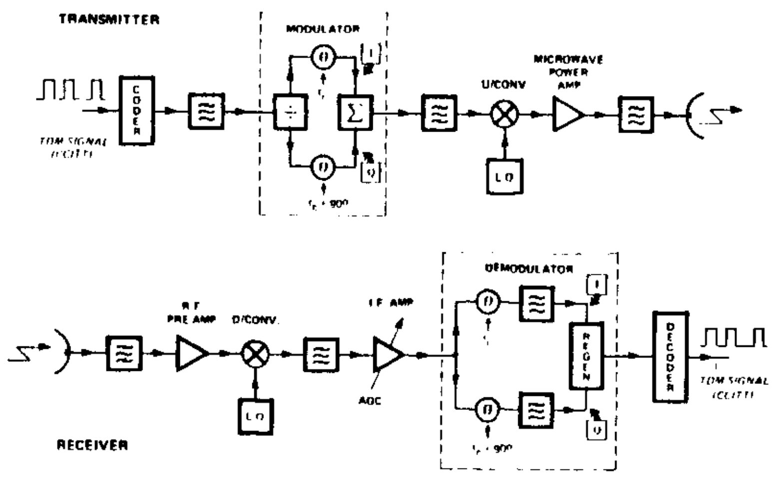
Several radio channels operate simultaneously in the same microwave band and through the same antenna. The combination and separation of these radio channels is carried out by waveguide filters and circulators. The filters must be carefully designed to avoid interference between adjacent channels. In addition, adjacent channels may be placed alternately on horizontal and vertical polarizations of the antenna to further improve the isolation.* There is also increasing interest in "frequency reuse", whereby two independent radio channels operate simultaneously on the same frequency but on opposite antenna polarizations.

Sometimes one channel in a multichannel system is assigned as a protection channel. If the error-rate on a particular traffic channel exceeds a certain threshold, the protection switch will automatically switch the traffic to the protection channel (called frequency diversity). The protection channel can also be used for scheduled out-of-service maintenance on a traffic channel.

* *Lightly loaded multi-channel systems may use one polarization for all receivers, and the other polarisation for all transmitters (for improved isolation, Tx-Rx). More heavily loaded systems will place adjacent channels on alternate polarizations, using separate Tx and Rx antennas for maximum adjacent-channel isolation as well as Tx-Rx isolation.*

DIGITAL RADIO BLOCK DIAGRAM

DIGITAL RADIO BLOCK DIAGRAM



Having seen, in simple terms, how the digital radio network is constructed we will now consider the block diagram of a digital radio transmitter and receiver in a particular channel.

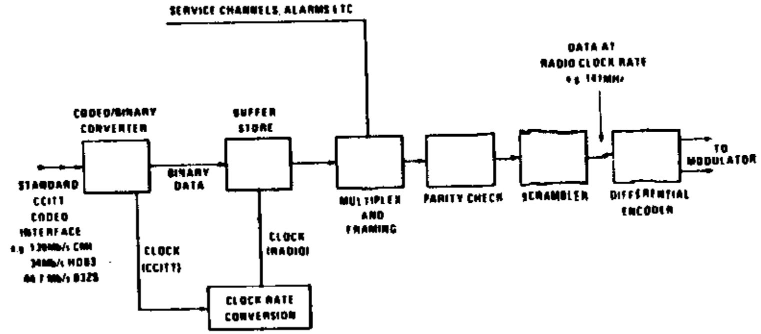
Although we are discussing digital line of sight radio, many of the concepts also apply to digital satellite systems.

Here is a simplified block diagram of a digital radio transmitter and receiver. Those of you familiar with analog radio will recognize a strong similarity in the block diagram, though the modulator and demodulator sections are very different as we shall see later. This block diagram shows I.F. modulation and demodulation (at the familiar 70 MHz or 140 MHz I.F.) with up and down conversion to the microwave transmit frequency. Most high-capacity digital radios use this system but there are quite a number of low-capacity radios with simple modulation schemes which use direct modulation at microwave frequencies. In this case the modulator is connected directly to the microwave power amplifier.

Most radios use the same receiver structure with down-conversion to the I.F. where the automatic gain controlled amplifier (typically 50 - 60 dB range) maintains a constant level to the demodulator during fading.

Notice the various filters through the transmitter and receiver. These are very important in the overall design as we shall see later. First we will look at the coder and decoder sections which provide the interface to the outside world.

TYPICAL DIGITAL RADIO CODER



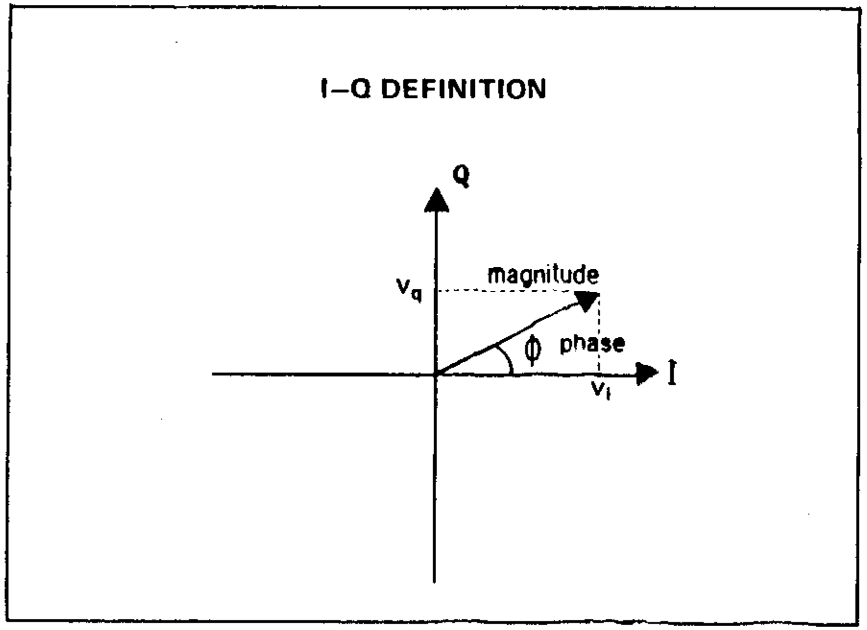
At first this block diagram looks rather complicated, however, its function is simply to provide the standard CCITT interface to the integrated digital network and then adapt the sequential bit stream to add the additional information used by the radio. The result is that the radio operates at a higher bit-rate than the CCITT interface. The additional information such as digital service channels* and alarms are multiplexed into the data stream along with framing signals to allow the receiver to sort out which bit is which. After this, a parity circuit adds a parity bit to produce an even or odd number of ones in a given block of data, and then the signal is passed through a scrambler to randomise the data being transmitted. The parity check is used by the receiver to check for errors in transmission and to initiate protection switching. The differential encoder provides the interface to the digital modulator and decides how the binary data will be encoded on the individual phase states.

In practical radios, two or more of these blocks may be combined into a single function or even one integrated circuit! At the receiver the decoder performs a similiar function in reverse. Note at a repeater station where no CCITT interface is required, some of the blocks may not be required. Generally this digital circuitry is highly reliable and does not require testing in installation or maintenance with the exception perhaps of jitter testing at the CCITT interfaces (G823 CEPT Standards, G824 North American Standards and Bell Technical References 43501 and 43806 and ECSA T1X1.3 Committee).

* *Service Channel and Alarm capabilities are typically short haul, "party-line" communication channels used for maintenance of the radio system. Some radios do not use digital service channels but instead frequency modulate the audio channel directly onto the carrier signal independently of the digital transmission.*

MODULATION SCHEMES

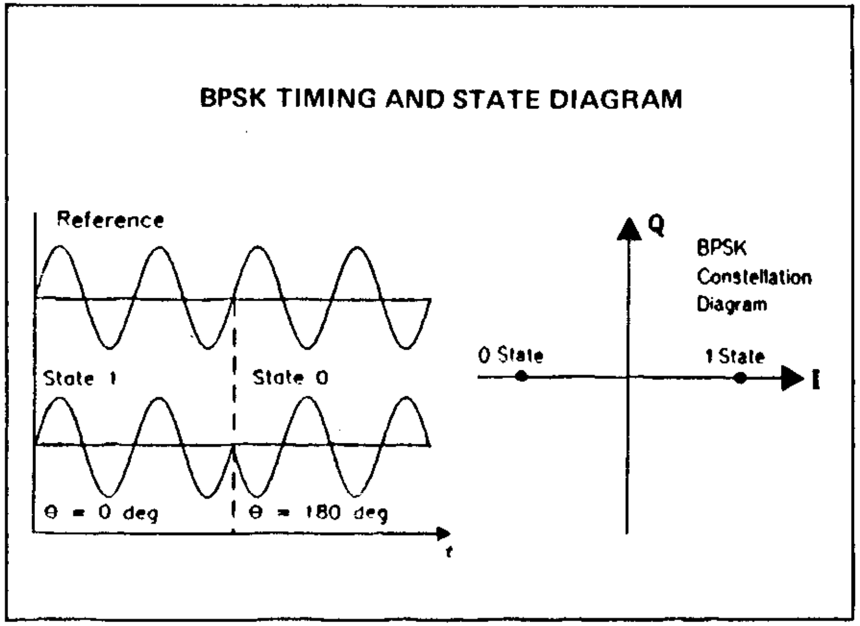
Having seen how the digital radio is interfaced to the network and how the digital signal is processed in the radio, we must now consider how this information is to be transmitted across the radio link by modulation of the microwave carrier signal.



The sinusoidal microwave carrier can be defined in terms of its magnitude, frequency and phase relative to an arbitary reference. In most digital radio systems the frequency of the carrier is fixed* so we need to consider only the phase and magnitude (or amplitude). As this slide shows, the phase and magnitude can be represented in polar or vector co-ordinates as a discrete point in the so-called I-Q Plane. I stands for in-phase (i.e. phase reference) and Q for quadrature (i.e. 90 degrees out of phase). We can then also represent this point by vectorial addition of a certain magnitude of in-phase carrier with a certain magnitude of quadrature carrier. This is the principle of I-Q modulation.

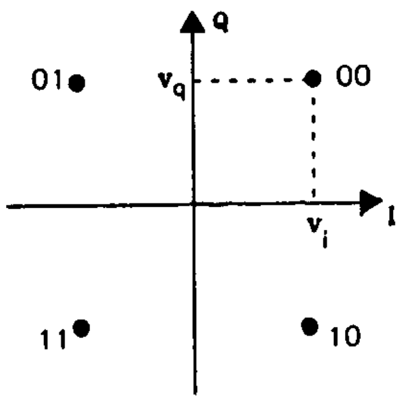
By forcing the carrier to one of several pre-determined positions in the I-Q plane, we can then transmit encoded information - each position or state representing a certain bit pattern, which can be decoded at the receiver.

* *Some low-capacity radios and satellites use frequency shift keying (FSK) and minimum shift keying (MSK) but these are not included in the present explanation.*



One of the simplest modulation schemes is Bi-Phase Shift Keying (BPSK or 2-PSK). Here the carrier magnitude is constant and to transmit either a '0' or a '1' the phase is "keyed" or switched between 0 and 180 degrees. Provided the receiver has a stable phase reference it can decide whether a '0' or a '1' was transmitted and regenerate the data stream. Notice that in this simple scheme only one bit of information is associated with each state so the carrier phase may have to be keyed at the bit-rate.

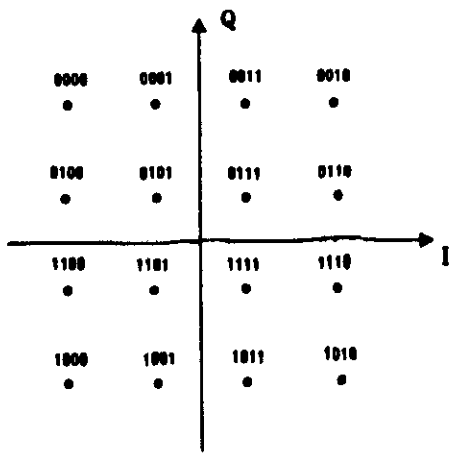
QPSK MODULATION



Quadrature Phase Shift Keying (QPSK or 4-PSK) again uses constant carrier magnitude but now four different phase values (i.e. 45, 135, 225 and 315 degrees) are used. The modulation phase-states can be generated by adding together appropriate amplitudes of in-phase and quadrature carrier (V_i and V_q), or alternatively by phase-shifting the microwave carrier directly using an electronically switched phase shifter such as waveguide stubs or delay lines.

Because we now have four discrete states we can transmit more information per state - in this case, as you can see, 2-bits of binary data or a symbol are encoded on each of the states. As serial data is taken 2 bits at a time to form the symbol, the symbol-rate is half the bit-rate. Intuitively you can probably deduce correctly that QPSK would only require half the bandwidth of BPSK for the same bit rate as its symbol-rate is half.

16 QAM STATE DIAGRAM



The quest for greater bandwidth efficiency has led to ever more complex modulation schemes. 16 QAM which stands for "Sixteen State Quadrature Amplitude Modulation" takes 4 bits of serial data and encodes them as a single phase state thereby reducing the symbol rate to one quarter of the bit rate. In order to generate this type of modulation the I and Q carriers need to take four different levels of amplitude (+3, +1, -1, -3) depending on the code being transmitted.

Comment:

Perhaps you will have noticed the way the phase states or symbols are assigned with 4 bit words. Adjacent states differ by only 1 bit so that if an error is made in the receiver in determining the transmitted state only 1 error will be generated, not 4. This is called Gray Coding. In practice things are not quite as simple as this since the data needs to be differentially encoded to resolve the phase-rotational ambiguity in the receiver. This means that some of the data bits are encoded as a change in phase state rather than absolute phase and this leads to error propagation or multiplication. This increase in error-rate due to differential encoding is sometimes called the Gray-Code penalty.

For more information see: "Differential Encoding of Multiple Amplitude and Phase Shift Keying Systems" by W. J. Weber, IEEE Transactions on Communications, Vol Com-26 No 3 March 1978 pp 385-391.

SYMBOL RATE CALCULATION

NUMBER OF STATES (OR SYMBOLS) $M = 2^N$

WHERE N = NUMBER OF BITS PER SYMBOL

$\therefore$ **SYMBOL RATE** $= \dfrac{\text{BIT RATE}}{N}$

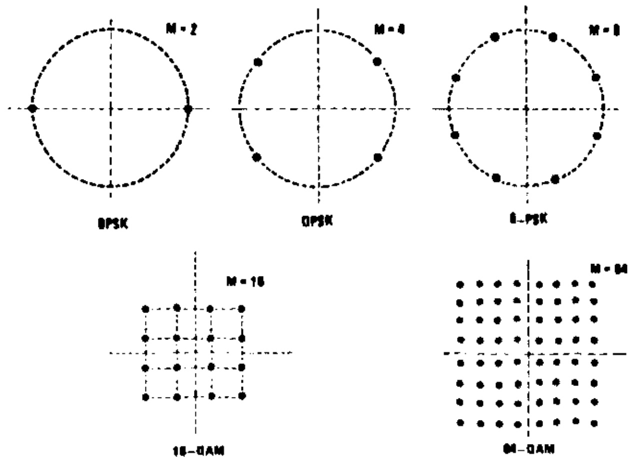
FOR 64 QAM $N = 6$

$2^6 = 64$ **STATES**

$\therefore$ **SYMBOL RATE** $= \dfrac{\text{BIT RATE}}{6}$

Calculating the **Symbol-Rate** is quite straightforward as this slide shows.

SOME TYPICAL MODULATION SCHEMES

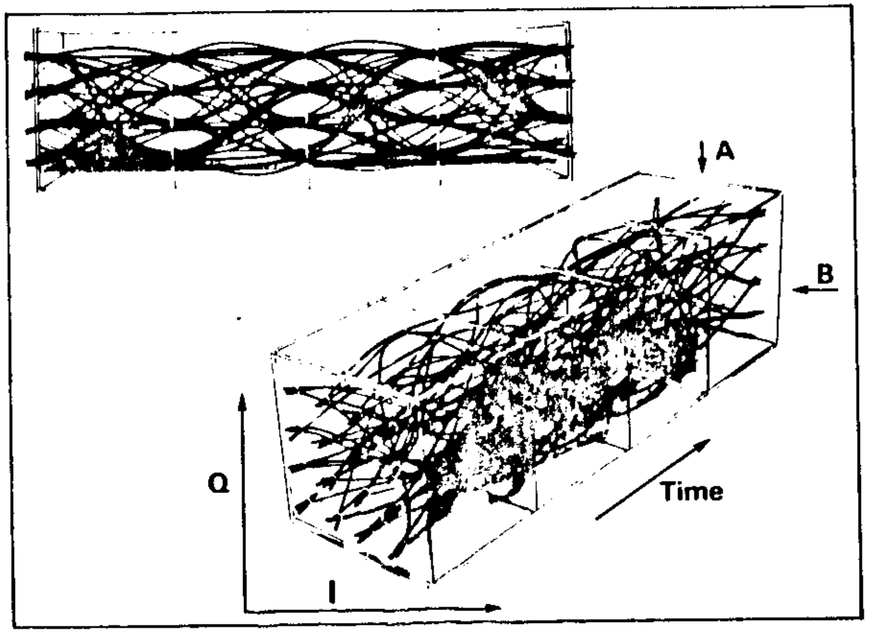


In summary here are the most common digital radio modulation schemes from simple BPSK to complex 64 QAM. It is interesting to compare the bandwidth efficiency of these schemes with analog FM radio when transmitting telephony. When carrying 64 kbit PCM, only 64 QAM can match an FDM/FM radio! So why does anyone bother with simple modulation schemes? The answer is that with 64 QAM the states are so close together that the immunity to noise and interference is greatly reduced compared with BPSK and QPSK. In hostile or noisy conditions (e.g. satellite) the simple schemes are favoured. In high-capacity line-of-sight systems where signals are strong, bandwidth efficiency is often considered more important. Research is now going on into 256 QAM systems, which will be operational by the late 1980s.

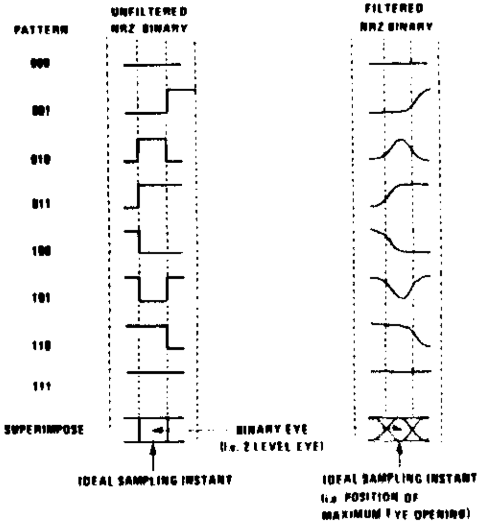
See Appendix A for more information on modulation schemes.

EYE DIAGRAMS AND CONSTELLATIONS

So far we have considered the phase states the carrier can take in the I-Q plane as static values. Of course, a real radio is supplied with a continuous bit-stream and so the symbol values or carrier states will change at the symbol-rate of the radio. At each symbol-timing instant a new phase-state value will be set up and, in between, there will be a transition.

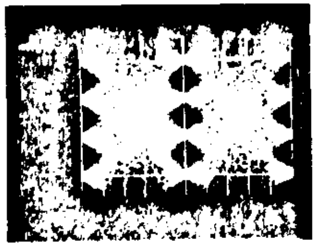


This 3-dimensional model of 16 QAM modulation shows the many transitions the carrier phase and magnitude makes between successive symbol values. The symbol timing clock is represented by the vertical planes and this is the instant when the receiver decides which of the 16 states was transmitted. Of course, the radio only transmits one state at a time but if we overlay many time intervals (as for example in the repetitive sweep of an oscilloscope) then we build up a complex picture of many transitions, as the next slide shows.

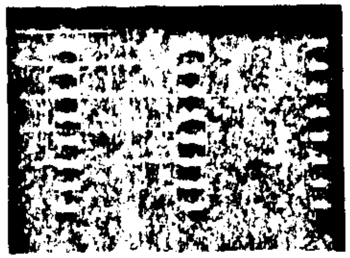


Eye diagrams, as they are called, may be thought of as viewing, on a storage oscilloscope, sections of a digital symbol pattern superimposed on top of each other. This slide shows the case for a simple NRZ binary data stream with various contributions of bit-patterns superimposed on each other. For a wideband system the transitions are instantaneous so the "eye" is square. A practical system, as we shall see later, must be filtered so the transitions are smoothed and we get the familiar eye-diagram shape. The optimum sampling instant is at the point of maximum eye opening where there is the highest probability of correctly detecting the binary state.

PRACTICAL EYE DIAGRAMS

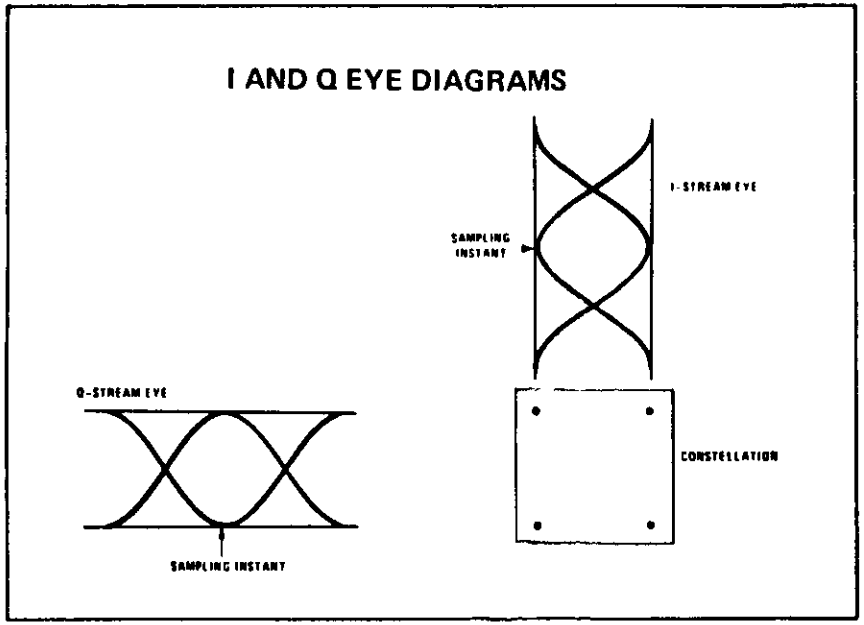


16 QAM EYE PATTERN

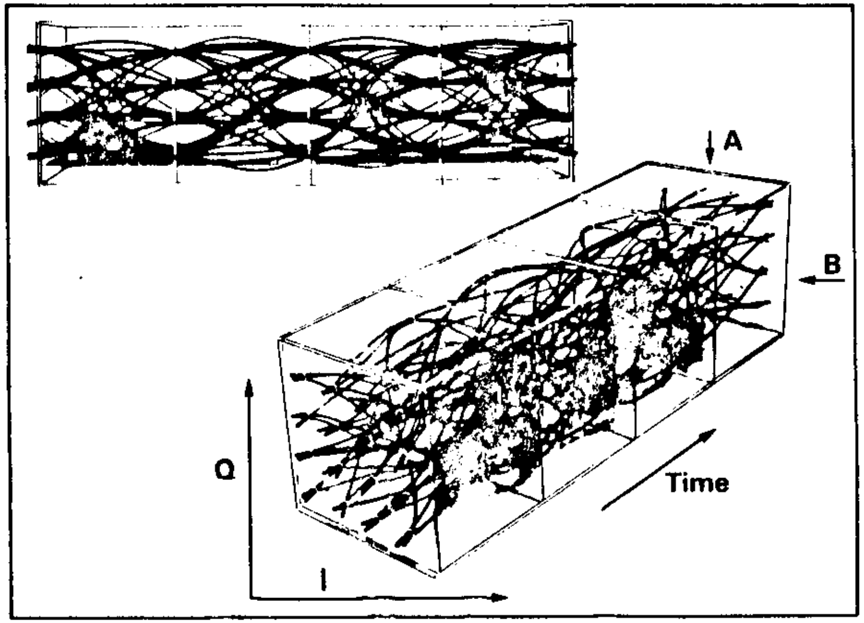


64 QAM EYE PATTERN

In digital radios the reconstructed data streams often exhibit multi-level eye patterns as, for example, in 16 QAM (4 levels) and 64 QAM (8 levels). A QPSK system has a two level or binary eye. You can see that with more complex modulation schemes the state values are closer together which means they have less tolerance to noise and interference as well as misalignment. You may also notice that the eye-width is narrower on 64 QAM than 16 QAM. This means that the system is less tolerant to errors or jitter in the symbol timing instant. The next slide shows that in a digital radio there are two eye diagrams.

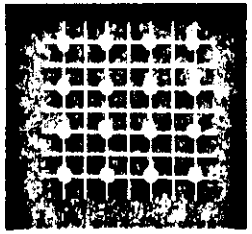


As we have seen, the digital radio modulation is constructed by separately modulating in-phase carrier (I) and the quadrature (Q). So after demodulation two independent eye diagrams are visible, one for I-stream and one for Q-stream, as shown in this slide. By displaying the value of I and Q eye diagrams at the sampling (or symbol-timing) instant and superimposing successive samples, we get the so-called constellation diagram.

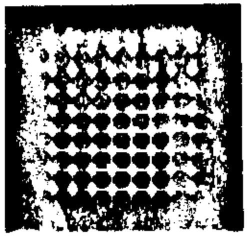


Perhaps you can see this more clearly now in the 3-dimensional model. Looking in the side (direction B) you will see the Q eye diagram, looking in the top (A direction) you will see the I eye diagram. If we build up a picture of the state values at the intersection of the vertical planes we get the constellation display shown in the next slide.

PRACTICAL CONSTELLATION DIAGRAMS



16 QAM CONSTELLATION



64 QAM CONSTELLATION

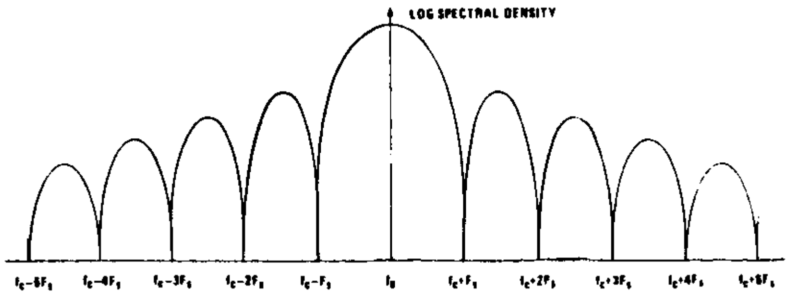
The constellation diagram is so called because the cluster of dots round each phase state looks like star clusters. Theoretically these should be single points but a practical radio suffers various impairments and noise which cause a spread on these states.

We have spent some time on this topic as an understanding of constellation diagrams and eye diagrams is very valuable in understanding degradations in the radio. The space between the phase states, the eye-opening in the eye diagram, is a measure of the quality of the radio and the probability of error.

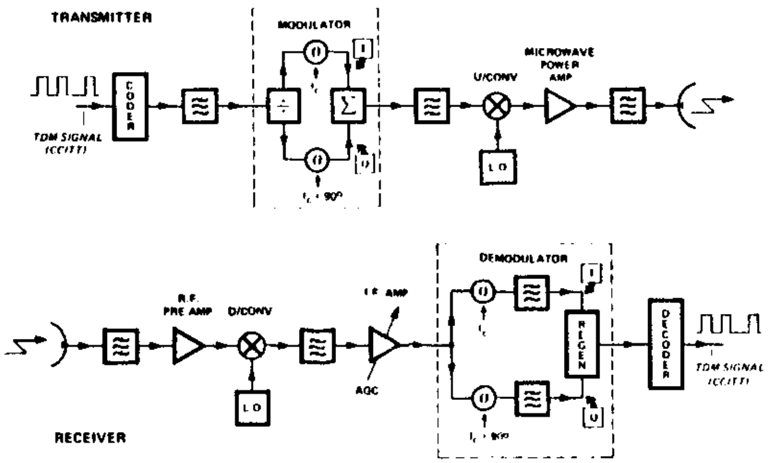
BANDWIDTH CONSIDERATIONS AND FILTERING

Having reviewed digital modulation principles we now need to consider the bandwidth required for transmission of these signals.

UNFILTERED DIGITAL RADIO SPECTRUM



DIGITAL RADIO BLOCK DIAGRAM

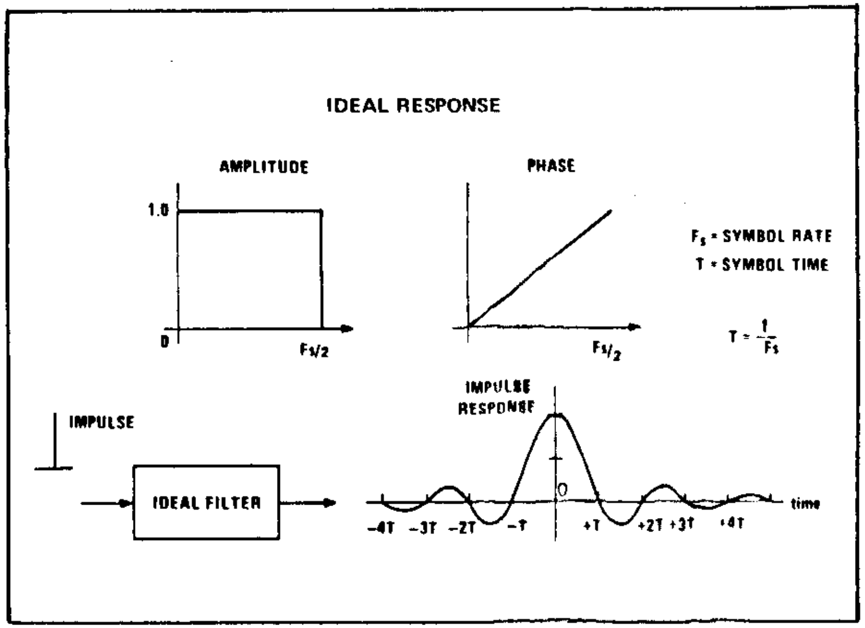


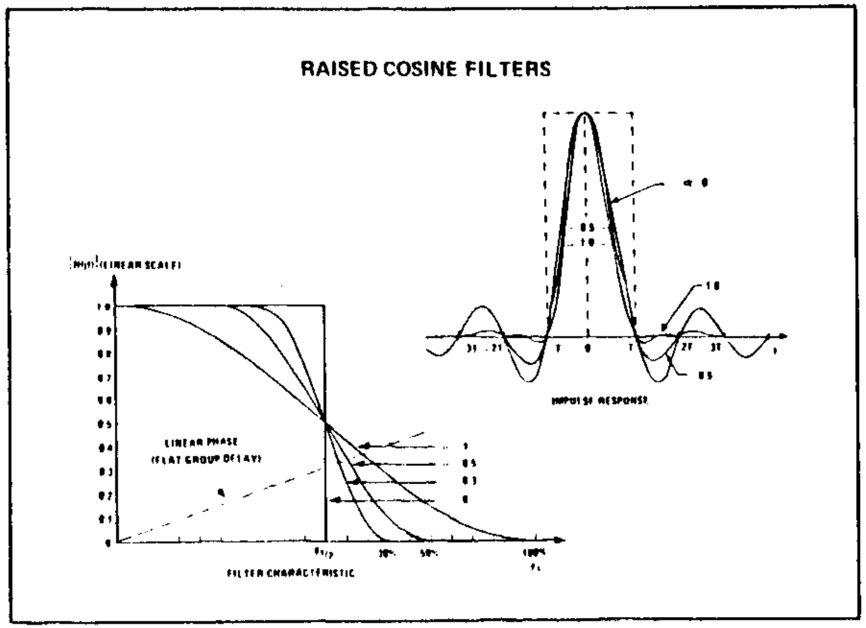
The unfiltered output of the digital radio modulator occupies a very wide bandwidth, theoretically infinite defined by the *sin x/x* characteristic. The digital signal modulating the radio is random, so the spectrum analyzer shows a noise spectrum picture with a spectral density shown in the slide. In fact, the spectrum of the radio should be independent of the data input to the radio - this is the purpose of the scrambler. The nulls in the spectrum occur at multiples of the symbol rate of the radio. The absence of the scrambler could cause a line spectrum to appear with some repetitive incoming data streams.

For practical application the radio spectrum must be restricted to avoid interference with adjacent channels. The radio filters are designed to do this while, at the same time, not degrading the data transmission.

The filtering in the radio can be implemented at a number of points. In the transmitter, the digital signal can be filtered at baseband (BB) with a low-pass filter before being applied to a linear modulator. Alternatively, the filtering could be introduced by band-pass filtering at I.F. and R.F. This will define the transmitted spectrum. In the receiver the filters are used to restrict the bandwidth of noise and to reject interference. Again the filtering could be done at R.F., I.F. or after demodulation at BB. Filtering in all three areas will contribute cumulatively to shaping of the data spectrum, but it may be that filtering must be implemented at RF or at IF to reject interference.

The second criteria for the filtering characteristic is that the overall response of all the filters in the transmitter and receiver should conform to the required shape for optimum pulse transmission. Exactly how this is shared between the filters will depend on the radio design.



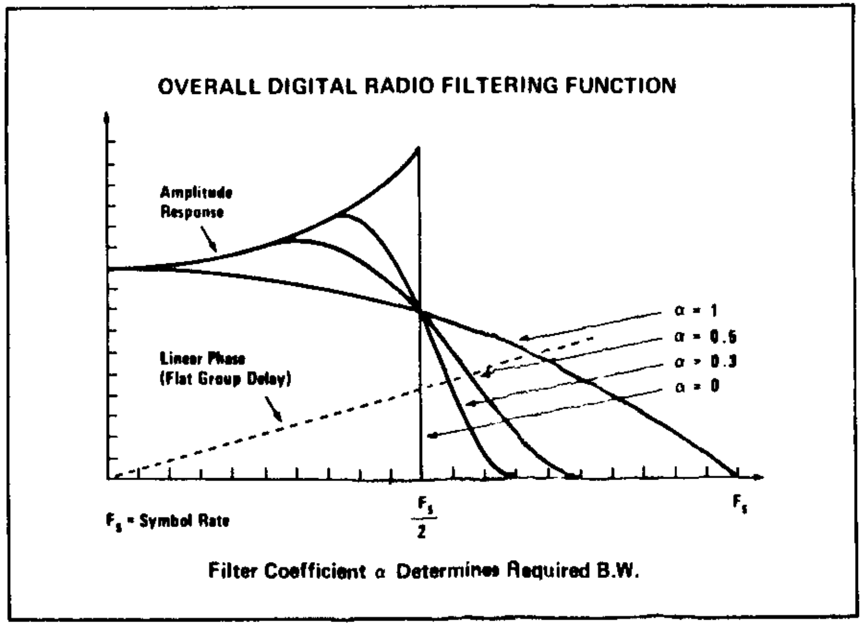


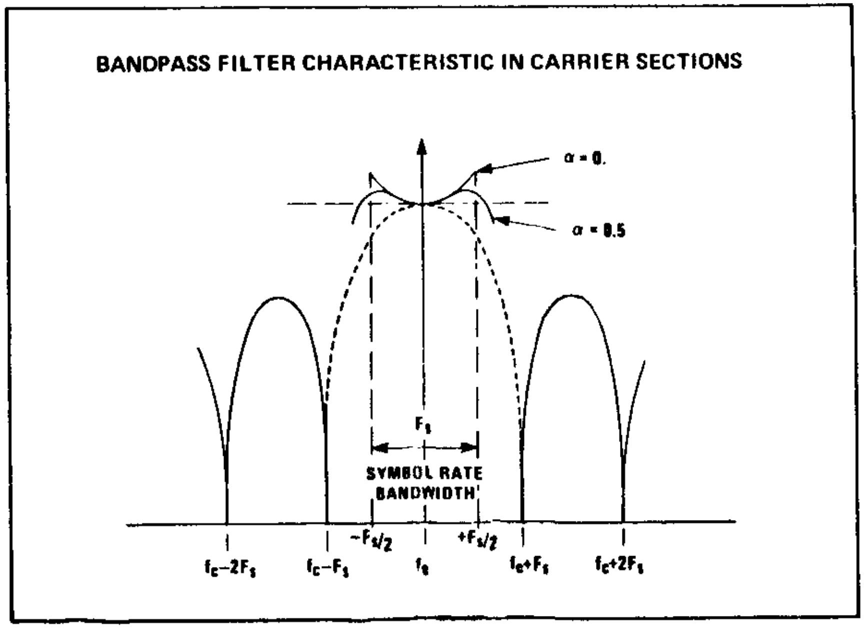
The ideal Nyquist filter response shown here is the minimum theoretical bandwidth for the transmission of impulses without intersymbol interference (ISI). The ideal filter is a "brick-wall" filter of bandwidth equal to half the symbol rate ($F_s/2$) having a linear phase response or flat group delay. In fact, the filter bandwidth is matched to the symbol-rate of the system. When this condition is met, the tails of the filter impulse response pass through zero at the sampling instants separated by the symbol time T. In a perfect world, this would mean that the symbol value at any of the sampling instants would be undistorted by the residue of surrounding symbols. This is what is meant by zero intersymbol interference, and the Nyquist criteria. If these conditions are not perfectly met the result will be ISI and closure of the "eye diagram".

References: *H. Nyquist "Certain Topics in Telegraph Transmission Theory" AIEE Trans. April 1928*
E.D. Sunde "Theoretical Fundamentals of Pulse Transmission" BSTJ May 1954
R.W. Lucky "Functional Analysis Relating Delay Variation and Intersymbol Interference in Data Transmission" BSTJ Sept. 1963.

Nobody has ever built an ideal brick wall filter, so instead one of the other realizable filters in the raised cosine family is chosen. As you can see they meet the zero ISI criteria but occupy more bandwidth. The filtering factor α (alpha) refers to the excess bandwidth over the minimum Nyquist bandwidth. For example, an alpha=0.5 filter has 50% excess bandwidth.

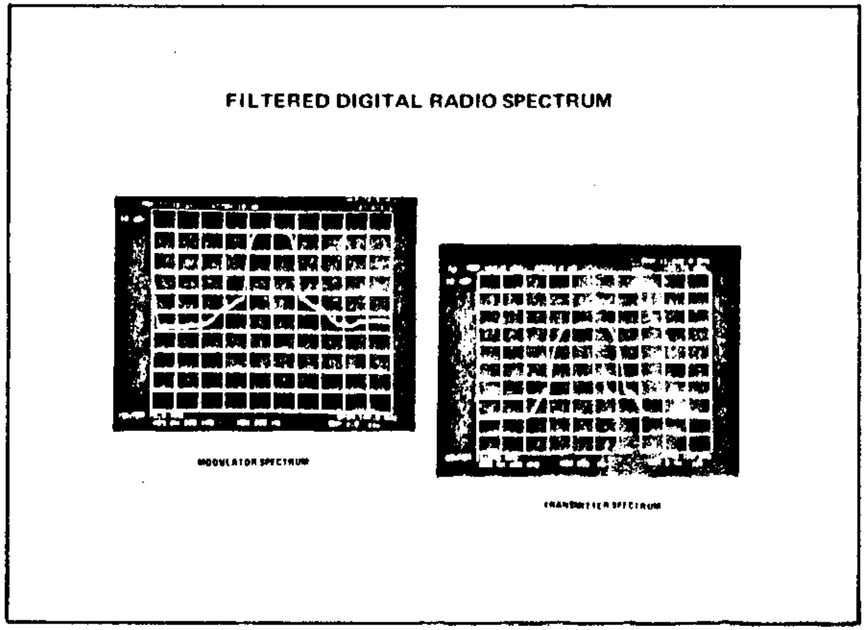
Note that all these filters are symmetrical about $F_s/2$. It is important that the filter has the correct bandwidth for zero ISI - a filter which is too wide is as bad as one which is too narrow, a rather different consideration compared with analog systems.

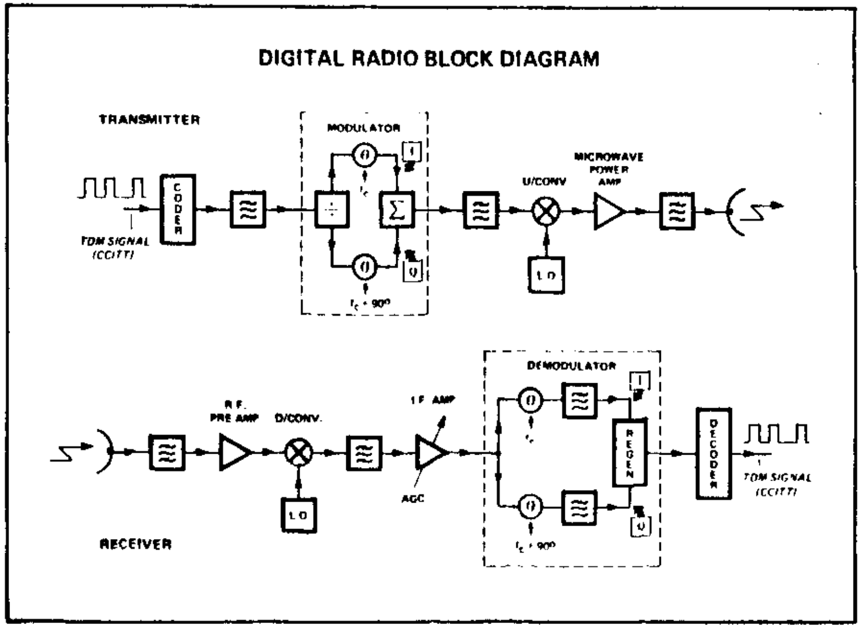




We have discussed the response of raised cosine filters to very narrow impulses, but a practical radio transmits pulses of width equal to the symbol timing period. The Fourier transform of an impulse is an infinitely wide flat spectrum (white noise) which after passing through an ideal Nyquist filtering appears as a rectangular noise spectrum. The fourier transform of this frequency spectrum is the zero ISI impulse response. To transmit pulses with zero ISI we need to shape the passband to compensate for the *sin x/x* roll-off of the pulse spectrum. This is the reason for the peaking towards the band-edge sometimes called "pre-whitening: as it produces a 'flat' filtered spectrum similar to white noise. Alternatively some systems may simulate the impulse by transmitting relatively narrow pulses rather than full width pulses.

The filters we have looked at so far are all low-pass filters at baseband. Because we are dealing with amplitude modulation of the I and Q carriers, exactly the same criteria apply for bandpass filters at I.F. or R.F with the filter characteristic mirrored about the carrier frequency. As we are now dealing with a double-sideband signal we need twice the bandwidth, so the **minimum bandwidth required in the carrier section is equal to the symbol-rate or the symbol-rate bandwidth.**





This slide shows the effect of filtering in the modulator of a 16 QAM radio, in this case, part of the raised cosine filtering of alpha=0.3. Notice that the *sin x/x* sidelobes are supressed and the "flat-top" of the in-band spectrum due to the "pre-whitening". This is a 70MHz I.F. radio with symbol rate of 22.5M symbols/second or 22.5M baud. The second photograph shows the spectrum at the transmitter output. The approximate -3 dB bandwidth of the spectrum is also equal to the symbol rate of 22.5 MHz.

The overall filtering function we have been considering is the effect of cascading all the filters in the transmitter and receiver from the output of the coder in the transmitter to the input of the regenerator in the receiver. The overall response must have flat group-delay, but the main bandshaping is usually shared between transmitter and receiver for example a square-root raised cosine filter characteristic in each.

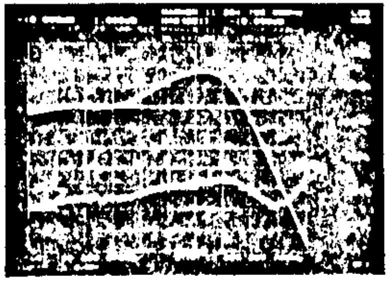
Perhaps this will be done in the I.F. filters, or maybe predominantly in BB low-pass filters with R.F. and I.F. sections being flat. Individual filters will not necessarily have the raised cosine response we have discussed and of course will not always have a flat amplitude response familiar in analog radio.

Comment:

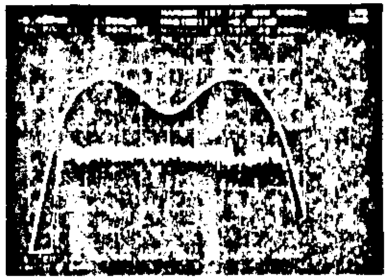
Practical radio filters may not have exactly the theoretical response described in this section. Modern computer optimization techniques enable a variety of amplitude and group delay characteristics to be synthesized which approximate to the zero ISI requirement.

Another variant in filter design is the so called partial response system (PRS) or correlative system. In this design the channel bandwidth is deliberately restricted to less than the Nyquist bandwidth so that controlled ISI produces a multi-level signal. An adaptive filter or correlative detector is used in the receiver. Examples of these systems are 9 QPR (filtered QPSK) and 49 QPR (filtered 16 QAM). In common with other complex modulation schemes, greater bandwidth efficiency is achieved at the expense of noise immunity.

PRACTICAL DIGITAL RADIO FILTERS



(a) BB FILTER



(b) IF FILTER

This slide shows some practical filter characteristics measured on the HP3577A Network Analyser. These are examples of raised cosine filtering at baseband (LPF) and I.F. (BPF). Over the passband the group delay is effectively flat. The markers have been positioned at the Nyquist frequency ($F_S/2$)

(a) Baseband Filter for a 90Mb/s 16 QAM radio with a symbol-rate of 22.5Mbaud, $F_S/2=11.25$MHz.

(b) I.F. bandpass filter (140MHz) for a 140Mb/s 16 QAM radio with a symbol-rate of 35Mbaud $F_S/2=17.5$MHz.

CHECK YOUR KNOWLEDGE

A DESIGNER IS CONSIDERING THE RADIO CHANNEL BANDWIDTH FOR A NEW RADIO WITH THE FOLLOWING DETAILS:

BIT RATE	:	**140 Mb/s**
MODULATION	:	**16 QAM**
RAISED COSINE FILTER	:	$\alpha = 0.5$

WHAT IS THE BANDWIDTH REQUIRED AT RF?

This little calculation uses the ideas we have reviewed so far. If you can work it out you have a good understanding of digital radio principles.

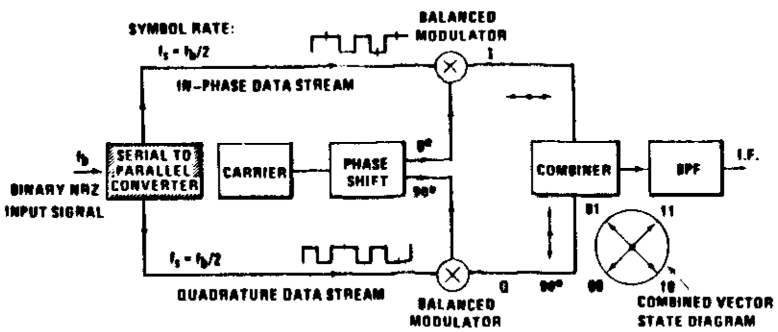
By now you will realize that digital radio design is significantly different from analog radio, particularly in the area of modulation and filtering.

See page 82 for solution.

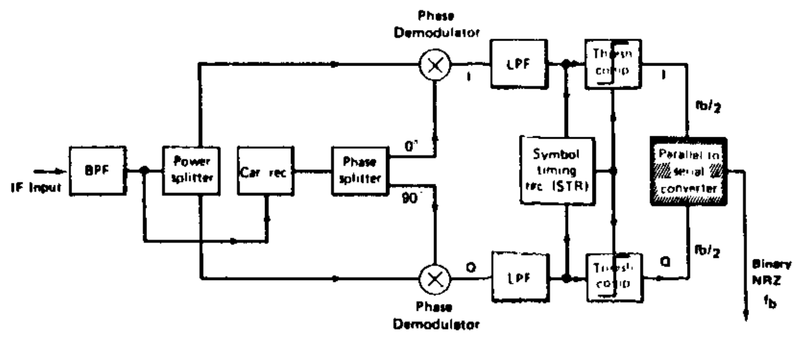
MODULATOR AND DEMODULATOR IMPLEMENTATION

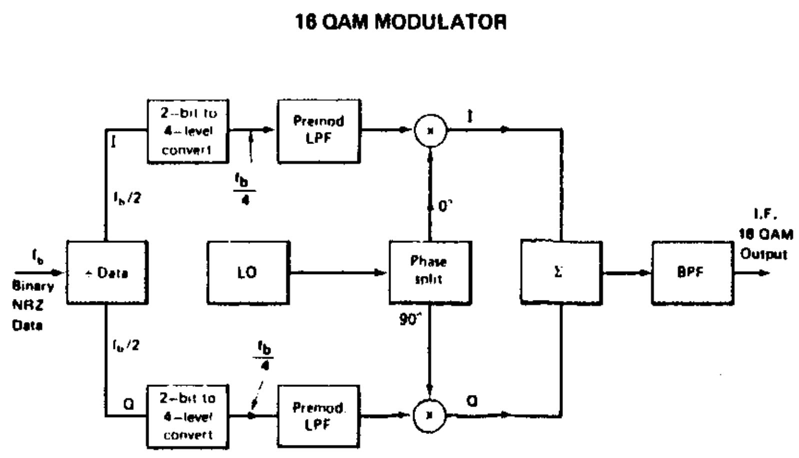
In the next few slides we will look in more detail at the design and operation of the modulator and demodulator in the digital radio. We will look in some detail at I/Q technique for generating the modulated signal, and finally we will discuss switching type modulators sometimes used in 2-PSK and 4-PSK systems.

QPSK MODULATOR



This slide shows the simplified block diagram of a QPSK (4-PSK) modulator. Earlier we saw how the digital modulation could be generated by adding together modulated in-phase and quadrature carriers. In QPSK the incoming bit-stream is divided into two parallel streams so that one bit is fed simultaneously to both I and Q balanced modulators to construct the 2 bit symbols. The carrier output from the modulator is switched under the control of the digital bit-stream and by adding together the I and Q outputs the phase state diagram is generated. In this case the bandlimiting filter is a bandpass filter at I.F., though, provided the modulators are linear, the filtering could have been implemented with LPF filters before the balanced modulators, thereby shaping the spectrum of the incoming pulses. Practically, some band-limiting is required before the modulators, otherwise the very wide *sin x/x* spectrum will fold around d.c. in an I.F. modulator and overlay the desired central lobe of the spectrum.

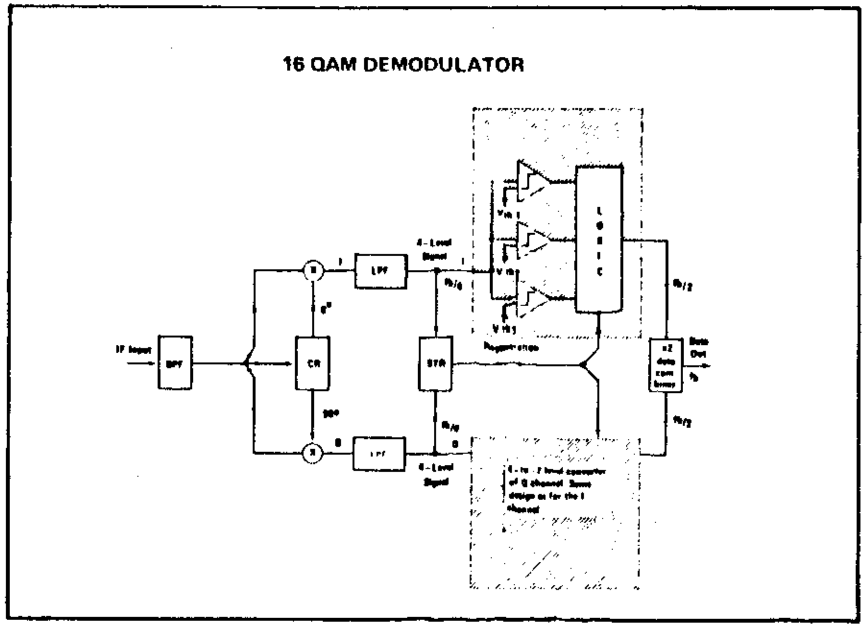




The QPSK demodulator works in a similar way to the modulator, extracting the I and Q streams by demodulation using in-phase and quadrature carrier signals. The demodulator is more complicated because it must recover a carrier signal and timing signal from the incoming I.F. Carrier recovery is usually implemented using a non-linear process such as frequency multiplication followed by a phase-locked loop. Symbol-timing is recovered from the demodulated data stream by a tuned circuit or phase-locked loop filtering out the clock component in the data stream. The scrambler in the transmitter ensures there is always a clock component independent of the data fed to the radio input.

The demodulator I and Q streams are filtered to remove unwanted I.F. signals and then passed into threshold detectors where a signal is sampled by the symbol-timing clock to determine whether a '1' or '0' is present and to regenerate the data stream. It is during this sampling and regeneration process that errors occur as we shall see later when we consider the effects of noise.

The 16 QAM modulator is similar to the QPSK modulator except that I and Q carriers are now each modulated by a 4-level signal. Two serial bits are converted to one of four voltage levels (by a simple DAC) which control the output of the balanced modulators. This shows how the symbol rate becomes $f_b/4$. In 64 QAM 3 bits are taken to generate one of 8 voltage levels on both I and Q streams. For these systems it is necessary to use linear modulators otherwise the phase state diagram will be distorted since the four voltage levels will not translate directly to the correct carrier level.

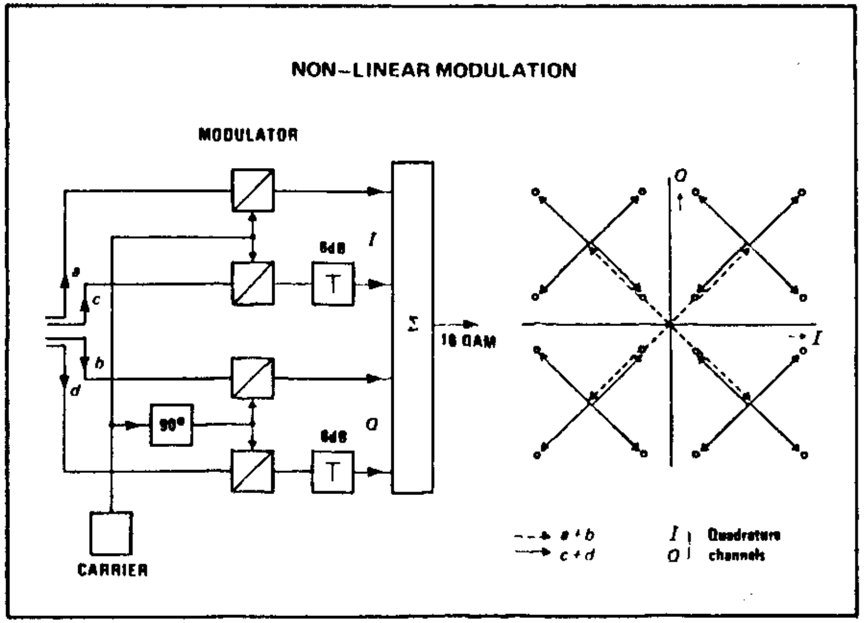


Again the 16 QAM demodulator is similar in outline to the QPSK demodulator except that the threshold detector needs to be more complicated as it must decide which of 4 possible values the signal has. With a knowledge of the encoding scheme used for the 16 phase states, the logic can be designed to regenerate the 4 bit words per transmitted symbol and so recreate the original bit stream.

In 16 QAM and 64 QAM systems the stability and accuracy of the modulator and the demodulator are very critical to the overall performance of the radio. Some designs use adaptive techniques to reduce the effect of drifts.

Comment:

Some radio and satellite systems use so called Offset Keyed or Staggered modulation. In these systems a delay of half a symbol-time is introduced between the I and Q data streams, so that the modulation envelope is not synchronized on both I and Q carriers. This has the advantage of slightly reducing the peak-power handled by the transmitter. Systems in use include Offset QPSK (OKQPSK) and Offset or Staggered QAM.

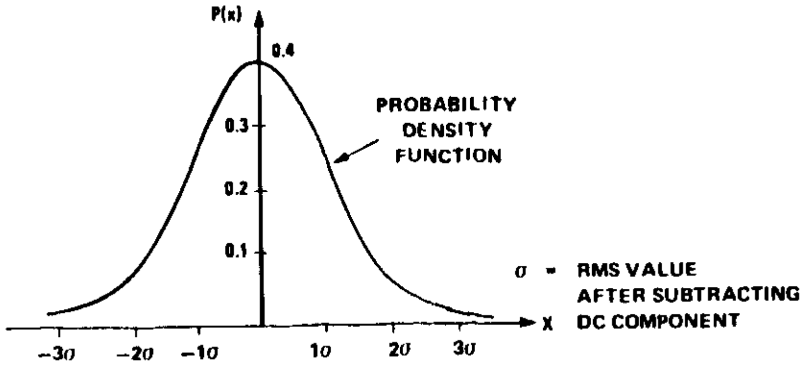


Some modulators use non-linear or switching techniques for generating the carrier phase and amplitude. One popular example is the use of PIN diodes to switch delay lines or waveguide stubs to generate BPSK or QPSK modulation at microwave. Another example is shown in this slide where a 16 QAM constellation is generated by adding together the outputs of four balanced modulators operating in saturation. These systems can have higher stability but require the bandlimiting filters to work at I.F. or R.F.

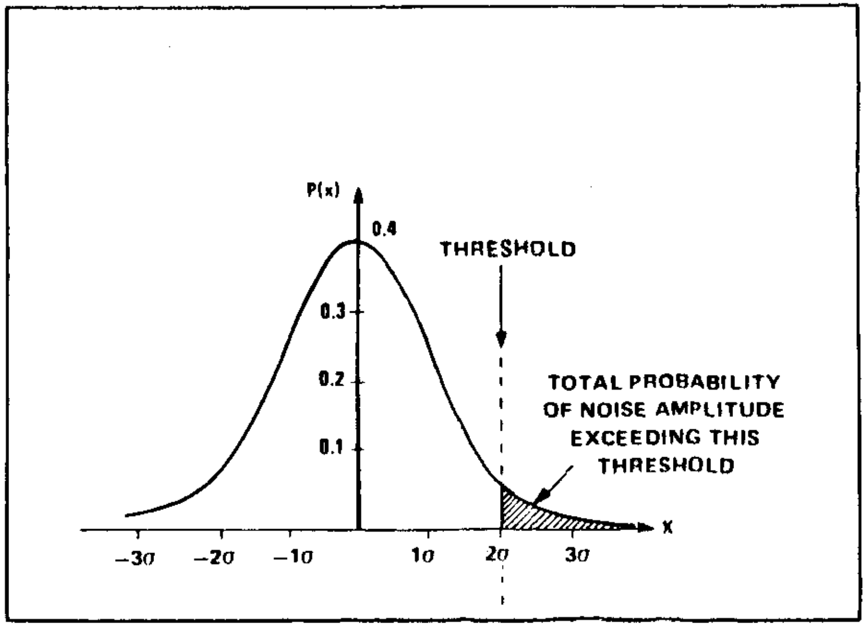
EFFECTS OF NOISE

In this last section on Digital Radio Theory we will look at the effects of noise in digital transmission and the specific definitions used in digital radio.

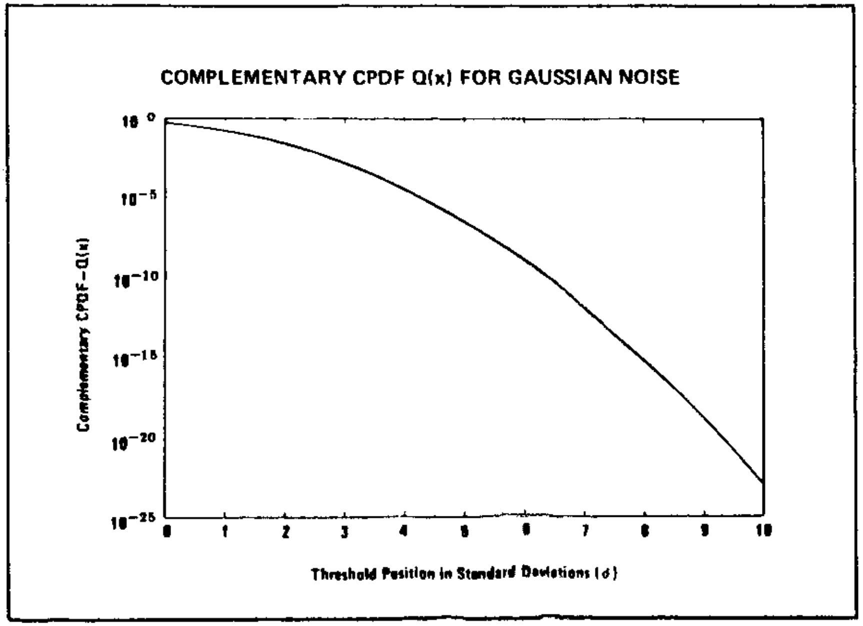
GAUSSIAN DISTRIBUTION



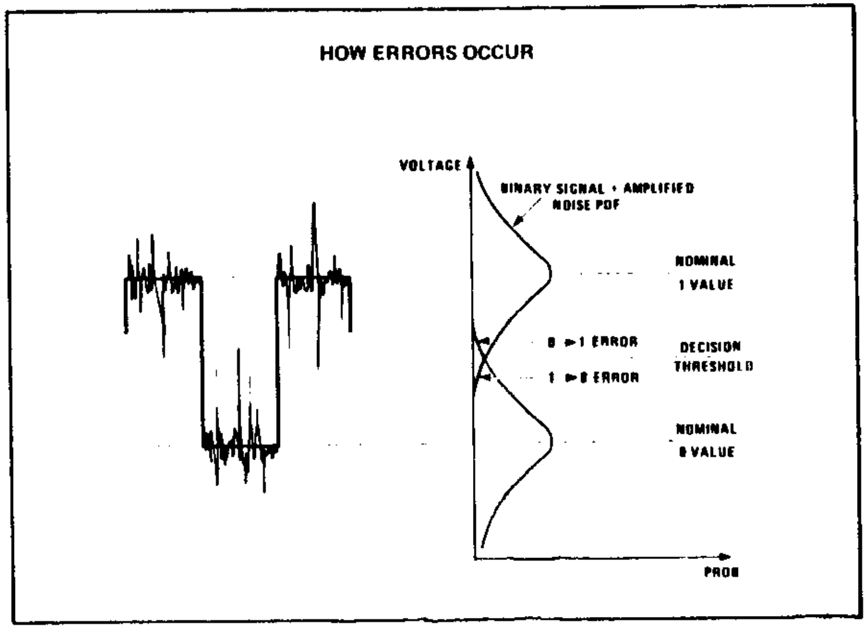
In order to understand the effects of noise in digital radio we need to consider the statistical properties of random noise. One of the most important statistical properties of a random variable is the probability of having a specified value indicated by the probability density function (PDF). The thermal noise generated in the front-end of the receiver can be represented by the Gaussian Distribution shown in this slide. The familiar bell-shaped curve shows the probability (vertical axis) of certain peak amplitude, x (horizontal axis) being reached. Notice the horizontal axis is calibrated in multiples of σ (sigma), the standard deviation or RMS value of the noise. In other words this is a normalised curve, the absolute values depending on the RMS value or power level of a particular noise generator.



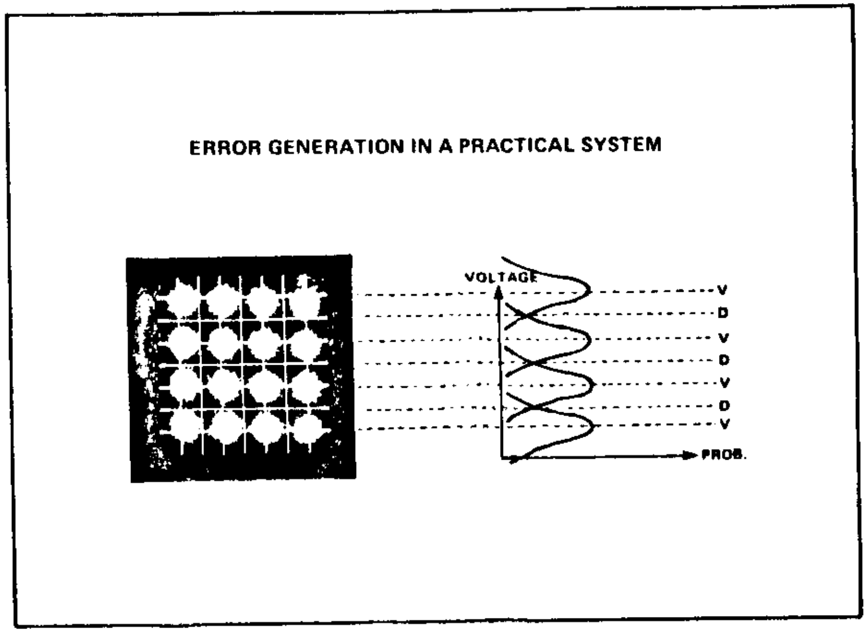
The probability of noise amplitude lying in a given range is the integral of the area under that portion of the curve. Obviously the area under the total curve (-infinity to +infinity) must equal unity (i.e. the signal must always exist somewhere within that range). Of particular interest in digital systems is the total probability of the noise amplitude exceeding a threshold value. This is called the **complementary cumulative probability distribution function** (Complementary CPDF), and is the area under the curve from the threshold value to infinity.



Plotting the complementary CPDF as a function of the threshold value (expressed in multiples of the standard deviation) we get the curve shown in this slide. You can see that the complementary CPDF becomes a very small value as we move out along the Guassian PDF. The complementary CPDF cannot be evaluated explicitly, but has to be calculated through a numerical approximation called the "error-function".



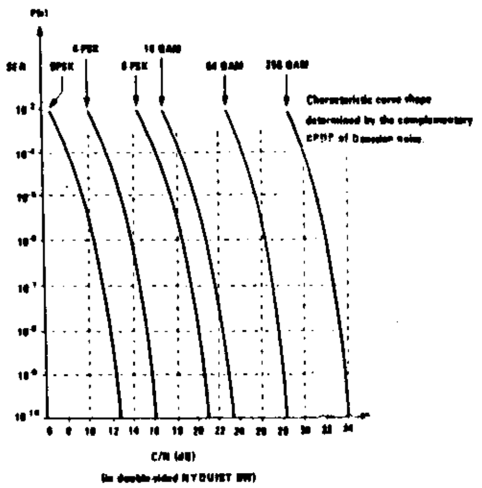
This slide shows how the superposition of noise on a digital signal generates errors. In the simple binary case shown here, a decision threshold will determine whether the received symbol (at a sampling instant) is detected as a "1" or "0". When noise is superimposed on the "1" and "0" values the PDF's shown in the slide indicate there will be a probability of a "1" being interpreted as a "0" and vice versa. This probability is the area under the curve on the wrong side of the decision threshold, which is given by the complementary CPDF as we saw in the previous slide.



A particular radio may have a more complex arrangement as, for example, in the 4-level 16 QAM system where there are three decision thresholds. Exactly the same arguments and mathematical concepts apply the only difference being one of scaling - as the phase-states or levels are closer together a lower noise power will generate the same error probability.

In fact if we take the ratio between the carrier (or signal power) and the noise power in a digital radio we can plot a set of universal curves independent of absolute levels.

THEORETICAL C/N vs P(e) CURVES

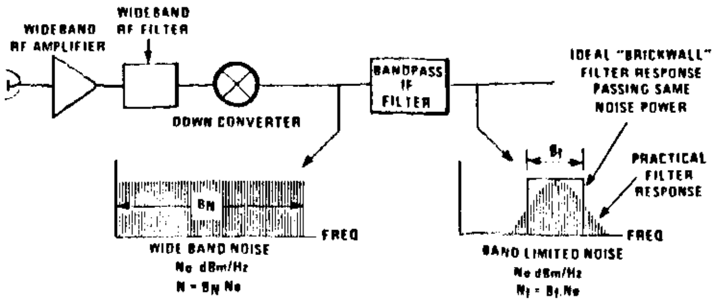


The curves on this slide show the theoretical relationship between the error probability (or symbol error rate SER) and the carrier to noise ratio C/N. As you can see all the curves have the same shape determined by the complementary CPDF for Gaussian Noise but are shifted horizontally depending on the complexity of the modulation scheme. As you might expect a 64 QAM radio requires a much better C/N ratio than 4-PSK for a given SER.

You can also see that, in common with other digital systems, digital radio exhibits a pronounced threshold, being effectively error-free down to quite low C/N ratios and then degrading rapidly to an unusable condition in 4-5dB change in C/N. The steepness of these curves means that high accuracy is required in setting the C/N value when making measurements.

These theoretical curves assume perfect modulation and demodulation with no inter-symbol interference. The noise power is also defined in the **Double-Sided Nyquist Bandwidth or Symbol-rate Bandwidth** which as we saw earlier is the theoretical minimum transmission bandwidth. In the next slide we can see the importance of defining bandwidth in noise power measurements.

NOISE BANDWIDTH



Consider the simplified diagram of a digital radio receiver. At the IF filter input, some amount of carrier power, C, and noise power N will be present. This noise power will be spread over a large but finite frequency range and will have a given bandwidth (B_N) and noise density (N_0) - i.e. noise level expressed as power in 1Hz bandwidth. At the I.F. filter output - assuming zero insertion loss for simplicity - the unmodulated carrier power will still be C. The noise-density will likewise remain unaffected, but the noise power will be reduced by the filter. The filter's noise bandwidth (B_f) is defined as the bandwidth of the equivalent "brick-wall" filter which would deliver the same noise power.

Clearly the C/N ratio measured at these two points would be different and if the bandwidths are undefined the results become confusing and meaningless. The noise power can be defined in the theoretical symbol-rate bandwidth or in the practical bandwidth of the radio.

CARRIER POWER DEFINITION

- UNMODULATED POWER — THEORETICAL BASIS
- MODULATED POWER (OR SIGNAL POWER) — LOWER BECAUSE OF SPECTRUM TRUNCATION

A second uncertainty which can arise in the setting of C/N is the definition of carrier power.

The theoretical curves are referenced to the **unmodulated carrier power** which, for a wideband system with PSK modulation, would be the same as the **modulated power**. As we have seen, however, practical systems must restrict the bandwidth and this truncation of the spectrum results in a reduced level for modulated power (maybe 1 or 2 dB less). In QAM systems results are further complicated by the presence of amplitude modulation - so for these systems an average modulated power is often used.

So you can see, for practical measurements, it is necessary to define the carrier power measurement for consistency and comparison of results.

BANDWIDTH INDEPENDENT UNITS

- CARRIER POWER TO NOISE DENSITY C/No dB Hz

WHERE $No = \frac{\text{NOISE POWER}}{\text{DEFINED RECEIVER BANDWIDTH}}$

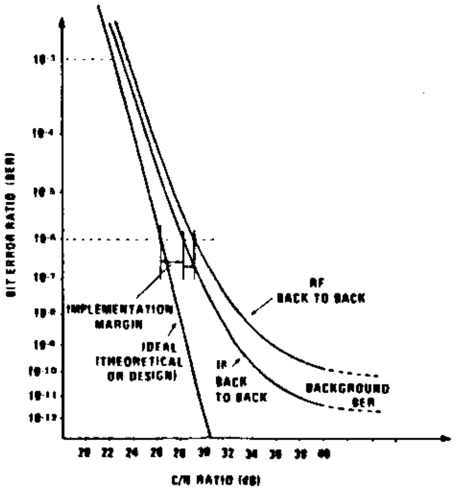
- ENERGY PER BIT TO NOISE DENSITY E_b/No dB

WHERE $E_b = \frac{\text{CARRIER POWER}}{\text{BIT RATE}}$

To avoid the ambiguities and uncertainties which can arise due to the definition of receiver bandwidth in the C/N measurement, bandwidth independent units can be used. These specify the performance in terms of noise density (N_0) which, as we saw earlier, remains unchanged despite filtering which results in a change in noise power.

When comparing the performance of different systems with more-or-less efficient practical filtering it is very valuable to use these units, particularly E_b/N_0 which gives a measure of a system's capability to transfer information in a given noise environment. E_b/N_0 is also a popular measurement unit in satellite systems. (See Product Note 3708-1 for more information on these units and the relationship to received signal level.)

PRACTICAL C/N versus BER CURVES



When we look at the performance of a practical radio, the results depart from the theoretical values in the way shown in this slide*. The difference between theory and practice is sometimes called the **implementation margin** and results from all the imperfections that can occur in practical radio.

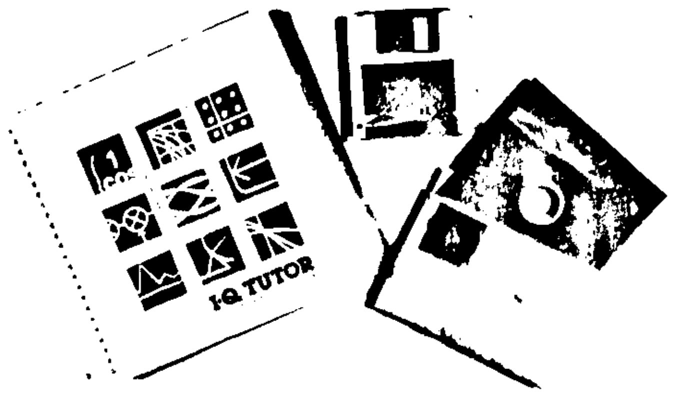
The poorer the performance, the greater the required C/N for a given bit error ratio (BER). At high C/N ratio the digital radio performance becomes asymptotic to the low-level background (or dribble) BER.

* *The practical results shown, are for a 64-QAM radio, and are plotted on "error-function" paper which has a vertical scale such that the theoretical curve plots as a straight line. Deviations from this line for practical systems are then clearly seen.*

Comment:

Normally a radio will be worse than the theoretical curve, i.e. it will require a higher C/N ratio for a given BER. The exception is for systems using forward error correction (FEC) when the practical system can have an overall performance better than theoretical, in which case, bandwidth is being exchanged for better BER. Some line-of-sight radios use this technique, and it is quite common in satellite systems.

IF YOU WOULD LIKE TO KNOW MORE



....HP I-Q TUTOR

That brings us to the end of our overview of digital radio theory, but if you would like to learn more, Hewlett-Packard has available an I-Q Tutor which comes complete with software to run on a 200 Series Controller such as HP9816 and HP9836. This package simulates many of the effects in digital radio and satellite, and provides a good practical understanding (Order Number HP 11736A, and HP 11736B for running on the IBM compatible Vectra System).

For text-book information on theory see

Feher, K; "Digital Communication Microwave Applications" (Prentice Hall Inc., 1981)

HP Product Note: 3708-1 "Noise and Interference Effects in Microwave Radio Systems" (Publication No.: 5953-5487)

PART 2

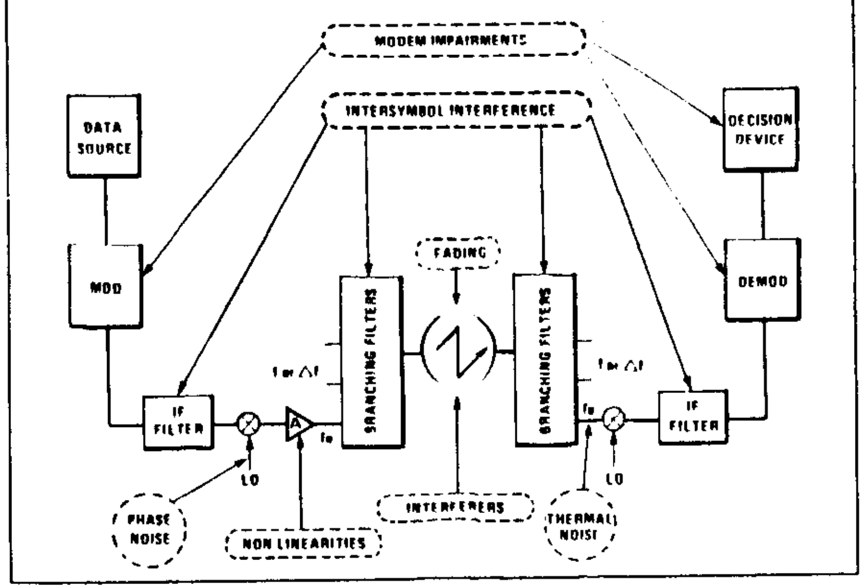
PRACTICAL PROBLEMS AND MEASUREMENT SOLUTIONS

DIGITAL RADIO
PRACTICAL PROBLEMS
AND MEASUREMENT SOLUTIONS

- DIGITAL RADIO IMPAIRMENTS
- PROPAGATION PROBLEMS AND STATISTICAL ERROR ANALYSIS
- RADIO EQUIPMENT IMPERFECTIONS
- CARRIER TO NOISE VERSUS BER TESTING
- MEASURING INDIVIDUAL IMPAIRMENTS
- SUMMARY OF DIGITAL RADIO TESTING REQUIREMENTS

In the second part of this presentation, we will consider the problems and impairments that may occur on a practical digital radio. We will consider propagation problems and equipment impairments under the headings shown on this slide, concluding with a summary of testing requirements. As each problem is considered, we will review the testing techniques and suitable measurement equipment.

BLOCK DIAGRAM OF A RADIO LINK WITH IMPAIRMENTS



A practical digital radio can suffer from a number of impairments which give rise to error generation in the system. The most common causes of degradation are illustrated on this slide.*

As you can see, some of these impairments are due to propagation and interference effects and are external to the radio equipment, while others are due to imperfections in the digital radio itself. We will look at these two categories separately as they require a different approach to measurement.

* *"Comparison of High-Level Modulation Schemes for High-Capacity Digital Radio Systems" by Michel Borgne. IEEE Transactions on Communication, Vol. Comm-33 No. 5 May 1985. pp 442-449.*

PROPAGATION PROBLEMS
AND
STATISTICAL ERROR ANALYSIS

First we will consider the effects of propagation on the error performance of the radio. As will be explained in the following slides, since propagation conditions vary continuously, the performance of the system can only be assessed on a statistical basis over a period of time (e.g. 1 month).

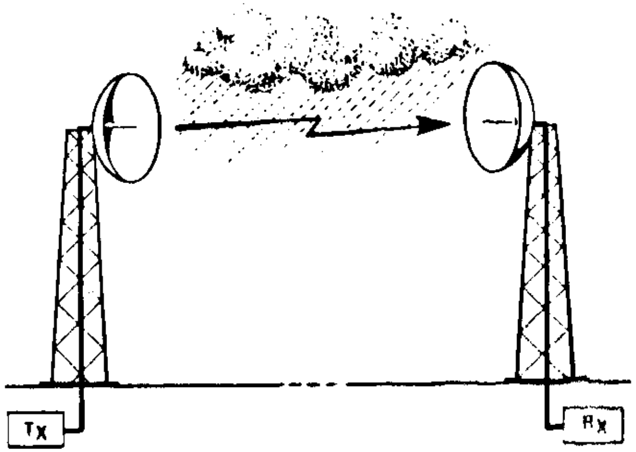
ERRORS DUE TO PROPAGATION PROBLEMS AND INTERFERENCE

- LOW RECEIVED SIGNAL LEVELS DUE TO FLAT FADING (EG HEAVY RAIN)
- MULTIPATH PROPAGATION CAUSING FREQUENCY SELECTIVE FADING WHICH GENERATES INTER-SYMBOL INTERFERENCE (ISI)
- INTERFERING SIGNALS FROM SATELLITES AND OTHER MICROWAVE RADIOS, PARTICULARLY ANALOGUE

Propagation effects are intermittent and time varying, often depending on weather conditions. Two common problems are **flat-fading** and **frequency selective fading.**

Interference from satellite signals, other microwave radios and external sources such as radar systems also cause eye-closure and therefore reduce the fade margin of the radio. We will consider the measurement of interference later in this seminar.

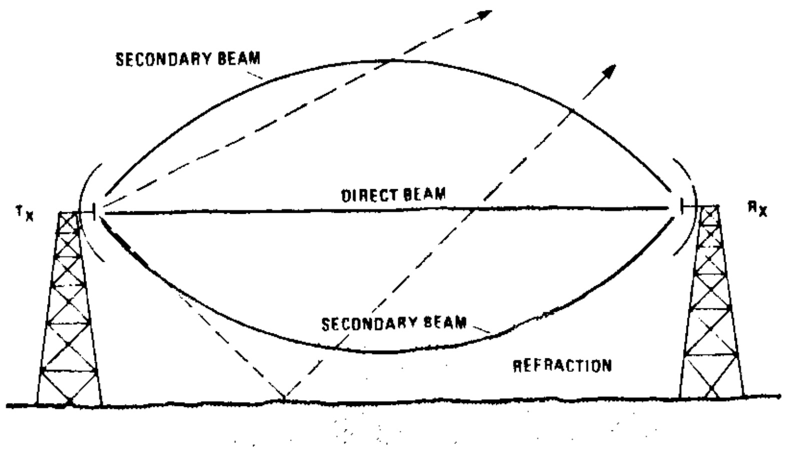
HEAVY RAIN CAUSES FLAT-FADING



Flat-fading as its name implies is a non-frequency dependent attenuation of the input signal and typically occurs during periods of heavy rain particularly at higher microwave frequencies (see Appendix B). When the signal level is low, the C/N ratio is degraded and errors occur in the transmitted data. Obviously the better the design and alignment of the radio (i.e. the greater the "eye opening"), the greater its fade margin. The fade margin is defined as the amount of signal attenuation (e.g. 50dB) from normal received level that is possible for a given Bit Error Ratio (BER) such as 10^{-3}.

A more serious problem, however, for most digital radio systems is multipath propagation which we will discuss in the next slides.

MULTIPATH PROPAGATION



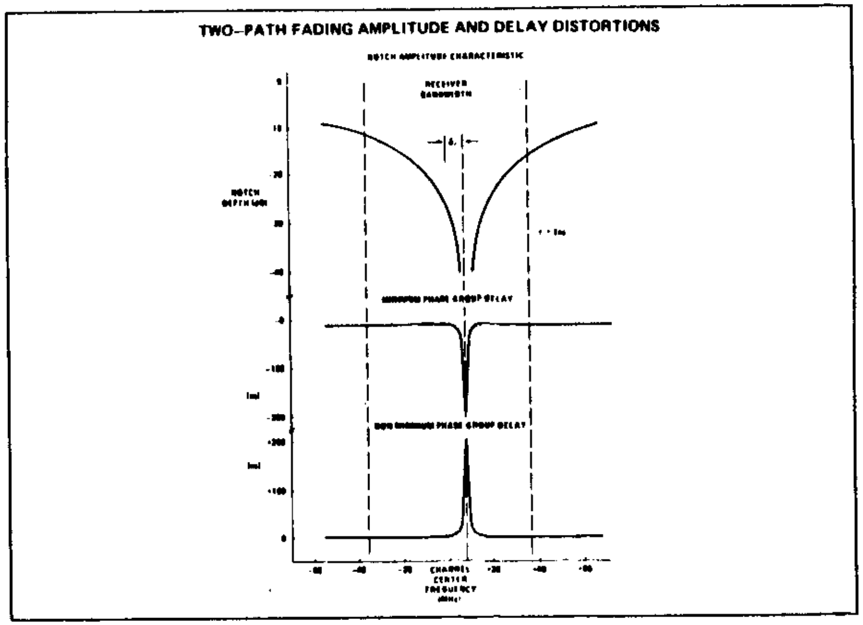
All terrestrial microwave radio systems can suffer from multipath propagation where the receiver antenna receives not only the direct signal but also a secondary signal which is slightly delayed relative to the direct beam. The secondary beam* bends due to the varying refractive index of the air. The phenomena is particularly prevalent in hot summer weather and when the radio path is across water. The result is frequency-selective fading.

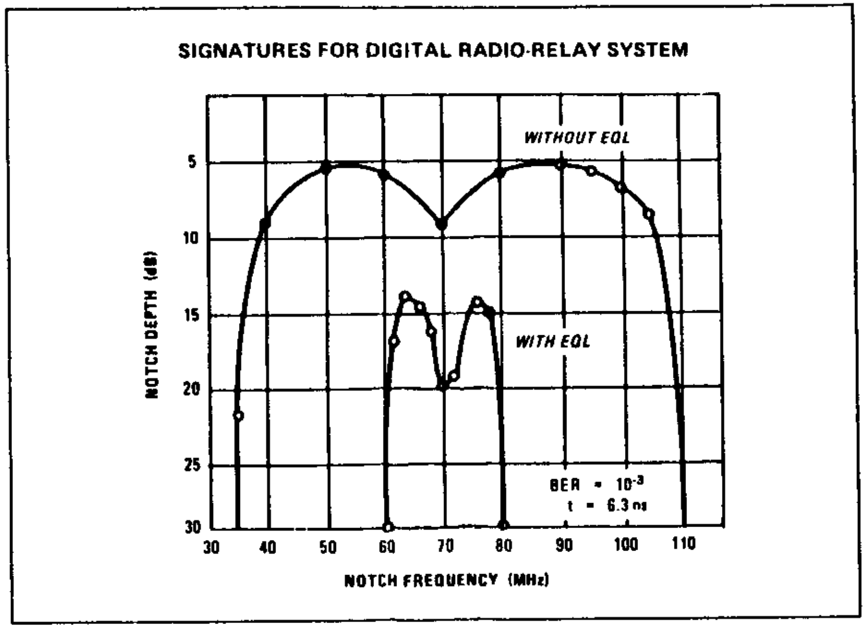
The degree of multipath fading is heavily dependent on the "hop" length - the longest hops requiring very careful attention in the system design. For this reason radio links operating at microwave bands below 15 GHz with relatively long hops tend to suffer predominantly from multipath propagation whereas higher-frequency systems with shorter hops are affected mostly by flat fading from attenuation due to rain.

* Secondary beam can also result from reflections off buildings etc.

For more information on Multipath fading see:

"Multipath Fading Channel Models for Microwave Digital Radio" by Rummler, Coutts and Liniger. IEEE Communications Magazine, November 1986, Vol 24, No 11, pp 30-41.





Frequency-selective fading gives rise to notches or slopes in the amplitude response across the radio channel. The group-delay is similarly affected, and as shown in the slide, the shape will invert if the fade becomes non-minimum phase, i.e. when the secondary beam amplitude is greater than the direct beam. The depth and frequency of the notch vary continuously during multipath fading and create very serious inter-symbol interference in a digital radio.

The presence of multipath activity can be identified in-service by monitoring the received digital radio spectrum on a **spectrum analyzer**. The normally stable and symmetrical spectrum will be distorted as the multipath fade passes through the band. Multipath can also be identified **in-service** on the **HP 3709B Constellation Display** (see page 79).

The result is short periods of very high error ratio which can cause loss of synchronization in the radio itself or in associated multiplex equipment.

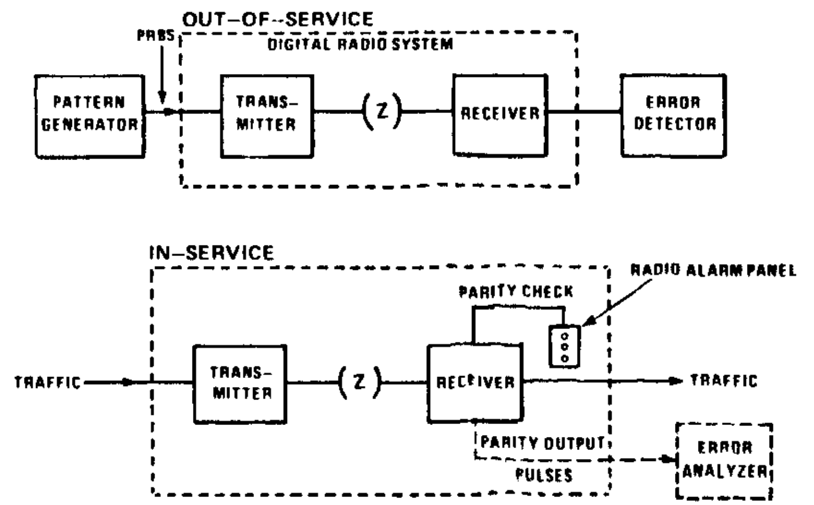
Comment:

In order to overcome this serious problem, counter-measures such as **diversity reception** *(spatially separated antennas providing a combined signal) and* **adaptive equalizers** *in the receiver must be employed, particularly in 16 QAM and 64 QAM systems. The adaptive equalizer cancels out the effect of the multipath fade either in the frequency-domain at I.F. or in the time-domain at B.B. Later in the presentation we will see just how sensitive the higher modulation schemes are to non-flatness.*

The effectiveness of adaptive equalizers can be checked by plotting the so called **"signature"** or **"M-curve"** which is the locus or plot of maximum notch depth as a function of notch frequency for a given BER such as 10^{-3}. These signatures can be plotted for a variety of conditions including minimum and non-minimum phase fades. The better the adaptive equalizer, the lower the "signature" indicating that deeper notches can be tolerated.

In the factory the multipath fade is usually simulated by means of a passive network at I.F. or more accurately at microwave. These simulators are usually "custom-designed" for particular frequencies and system configurations.

ERROR PERFORMANCE TESTING



Error performance measurements can be made in two ways:

(a) **Out-of-service**, where the traffic is removed and a pseudo-random binary sequence (p.r.b.s.) is applied to the transmit terminal, and the received data stream checked bit-by-bit for errors. Sequence lengths of 2^{15}-1 and 2^{23}-1 are specified by CCITT. This is the preferred method for assessing the performance of the radio particularly during commissioning since every bit is checked for errors. Normally the pattern-generator and error-detector are connected at the coded CCITT interface on the terminal. Alternatively, the connection may be made at a binary data and clock interface depending on the terminal design.

(b) **In-service**, where the radio operates normally carrying revenue-earning traffic and the error performance is measured internally by parity checking on data blocks. This works quite well at moderate or low error ratios, but becomes inaccurate during burst of errors, for example during multipath fading, when there is a possibility of parity error cancellation in the data block. The result of this simple test is usually displayed on the radio control panel. Alternatively the parity error detection may be available as an electrical pulse which can be connected to the "external error input" of the error analyzer.

For more information see:-

CEPT Standards - HP 3764A Training Manual (Pub No 5953-5444)
N. American Standards - HP 3781/82B Training Manual (Pub No 5952-3290)

DIGITAL RADIO MEASURE OF PERFORMANCE

BIT ERROR RATIO (BER) =

$$\frac{\text{NUMBER OF ERRORS COUNTED IN THE MEASUREMENT INTERVAL}}{\text{TOTAL NUMBER OF TRANSMITTED BITS IN THE MEASUREMENT INTERVAL}}$$

Once the errors have been counted the result can be expressed as the bit error ratio (BER). This is the basic measure of error performance. Obviously the result can be integrated over different periods of time depending on the degree of averaging required. For statistically stable results it is desirable to count at least 100 errors per gating period. This can lead to very long measurement periods at low BER values.

PROPAGATION PROBLEMS REQUIRE STATISTICAL ERROR ANALYSIS

CCIR RECOMMENDATION 594 FOR 2500 KM LINK
(64 kb/s UNIDIRECTIONAL CHANNEL)

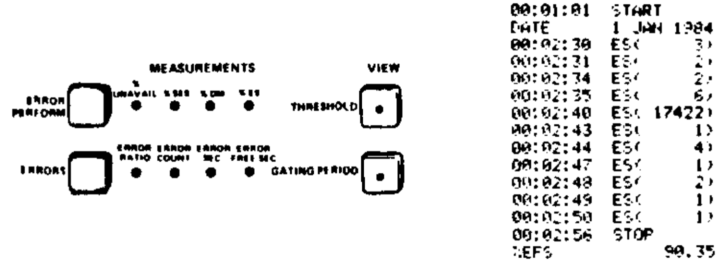
- **BER WORSE THAN 1 x 10^{-6} FOR LESS THAN 0.4% OF ANY MONTH.**
- **BER WORSE THAN 1 x 10^{-3} FOR LESS THAN 0.054% OF ANY MONTH.**
- **ERRORED SECONDS SHOULD NOT EXCEED 0.32% OF ANY MONTH.**
- **RESIDUAL BER SHOULD NOT EXCEED 5 x 10^{-9} (15 MINUTE INTEGRATION).**

Because of the time-varying characteristics of the error-performance, the measurements must be classified statistically for the % time that the error performance is worse than certain thresholds. The latest CCIR Recommendation 594 is shown in this slide and relates to a unidirectional 64kb/s channel over a hypothetical reference circuit of 2500km. For shorter length systems the % time and residual BER value would be scaled in proportion to the length (L/2500).

The recommended measurement period is one month and usually this is taken to be the worst propagation month for a particular route depending on whether it is dominated by multipath propagation or rain attenuation.

These revised recommendations are intended to be compatible with a high-grade circuit in the CCITT Recommendation G821 for the error performance requirements of the ISDN. The addition of the errored-second objective also recognises the importance of burst errors in microwave radio systems, caused in particular by multi-path propagation.

ERROR ANALYSIS SECTION OF A DIGITAL TRANSMISSION ANALYZER PROVIDING SIMULTANEOUS RESULTS IN ALL CLASSIFICATIONS



In order to classify the performance according to the CCIR/CCITT recommendations, simultaneous anaylsis of all the error-performance criteria is necessary during the measurement period of up to one month.

This slide shows the 8 simultaneous measurements available on the HP3764A Digital Transmission Analyzer. The error-performance categories are those defined in CCITT Recommendation G821.

% Unavailability - 10 sec period at worse than 10^{-3}
% Severely Errored Seconds (% SES) - Available second periods worse than 10^{-3}
% Degraded Minutes (% DM) - Available periods worse than 10^{-6}
% Errored Seconds (% ES) - Available seconds with one or more errors

Comment:

These relate closely to the CCIR recommendation shown in the previous slide, and in both cases the performance relates to a single 64kb/s channel. Normally, of course, digital radio measurements are made at higher multiplexed rates and the question arises as to how these results relate to the 64kb/s criteria. It is usually accepted that taken over a sufficiently long period, the BER criteria transfer directly to the high-speed measurement. For the error-second criteria it is necessary to know the number of errors that occur in an errored second on the high-speed system. From that an estimate of %ES at 64kb/s can be made. An example of error-second error count is shown in the printout on this slide.*

* *"Experimental Results on Digital Radio-Relay Quality compared with new CCIR Objectives" by Damosso, Dionis, Zenobio and Fabbin. European Radio Relay Conference, Munich Nov 1986.*

PROPAGATION PROBLEMS REQUIRE STATISTICAL ERROR ANALYSIS

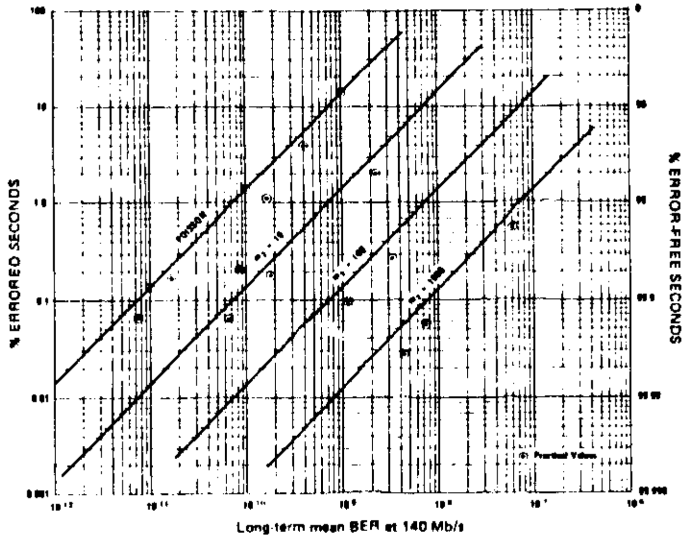
BELLCORE TA-TSY-000236 (JULY 1986)
FOR 4/6/11 GHz DIGITAL RADIO (1 WAY, 1 HOP)

FOR A TEST PERIOD OF AT LEAST 5 CONSECUTIVE DAYS:

- **ERRORED SECONDS SHOULD NOT EXCEED 0.04% OF THE PERIOD.**
- **NO MORE THAN 2 SEVERELY-ERRORED SECONDS CONTAINING AT LEAST 2800 ERRORS**

In the USA another error performance standard has been proposed by Bellcore in TA-TSY-000236. This considers the performance of a single radio hop, presumably at the multiplexed DS3 rate (44.736 Mb/s). This is an easier test to perform on a hop-by-hop basis than CCIR Rec 594, but it is harder to relate the results to the overall link performance as seen by the service user at 64 kb/s.

ERROR ANALYSIS HELPS TO DIFFERENTIATE BETWEEN FLAT-FADING AND MULTIPATH PROPAGATION



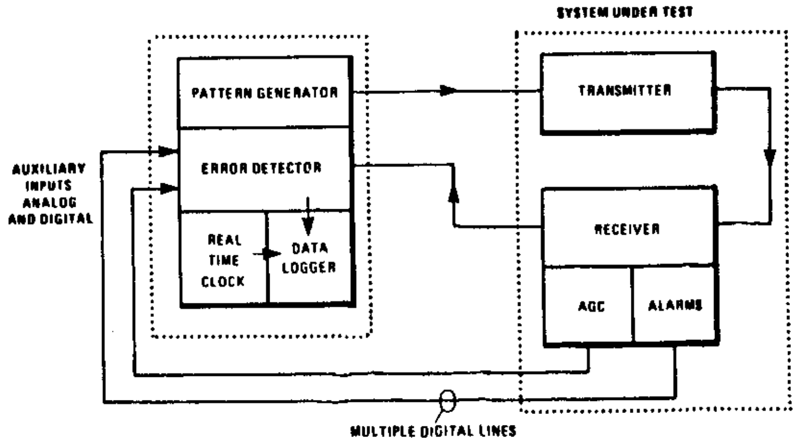
Another way of using the error performance measurements is to analyse how the errors occur on a practical system. On a digital radio system errors may occur in an even distribution due to noise when a flat fade occurs or they may occur in short bursts during multipath activity. The long-term BER value is only dependent on the total number of errors counted, whereas % errored seconds will depend on how the errors are distributed in time (an errored second could contain just one error or many thousands if an error burst occurs). For example a BER of 10^{-9} at 140 Mb/s would cause 14% ES if the errors occurred as single events, but less than 0.2% ES if they occurred on average in groups of 100.

This relationship can be plotted on the graph shown in this slide where %ES are plotted against the mean BER for successive measurement periods (e.g. 24 hours). We can see then whether the errors occur in a random Poisson distribution characteristic of a noise limited system, or whether they occur in clusters. (The straight lines are derived from the theoretical relationships for Poisson and Neyman distributions, m_2 indicating the cluster size).

For more information see:

"Error Performance Requirements for Future ISDNs"
by Ron McDowall, Hewlett-Packard (Pub No 5954-7938)

"Performance of Digital Radio-Relay Systems"
by G. Hart, British Telecommunications Engineering, pp 201-209 Vol 3, Oct. 1984.



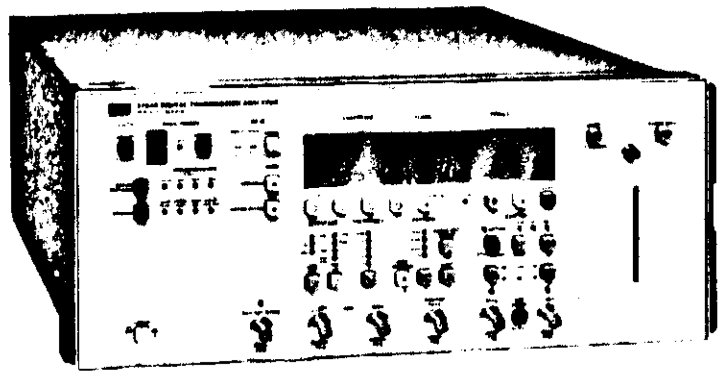
Often when monitoring the error-performance of a radio system, it is useful to correlate error occurrence with other events such as radio alarms and a.g.c. voltage (as an indication of fade depth). On some error detectors, these signals can be checked and recorded with time along with the error measurements.

This has a substantial benefit in reducing the amount of equipment required for a complete test of the radio, avoiding the need for data acquisition equipment and system software to correlate the measurements.

Hewlett Packard manufactures a wide range of pattern-generators and error-detectors suitable for North American or European hierarchies. The latest products incorporate comprehensive error-performance analysis and data-logging.

HP 3764A DIGITAL TRANSMISSION ANALYZER

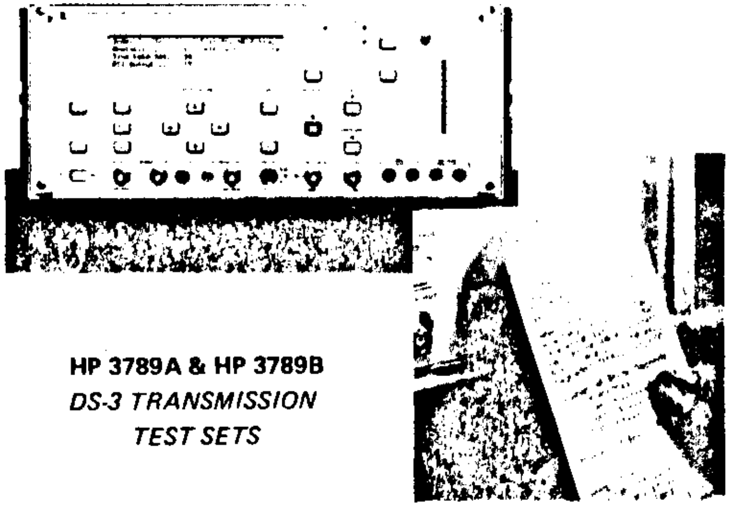
(FOR CEPT RATES 2, 8, 34 & 139 Mbit/s)



The HP 3764A Digital Transmission Analyzer contains a pattern generator and error detector at the standard CEPT bit-rates of 2, 8, 34 and 139 Mbit/s. It has many valuable features for digital radio testing:

- Simultaneous error analysis in all G.821 categories.
- Non-volatile memory for ease-of-use and protection of long-term measurement results during power failure.
- Results recorded on internal printer or on 80 column external printer (HP 2225A Thinkjet).
- Analog and digital auxiliary inputs.
- Standard CCITT clock frequency offsets available with Opt 005.

For the manufacturer of 139 Mbit/s radios, the HP 3764A Opt 002 provides complete error performance and jitter testing in one economical package.



HP 3789A & HP 3789B
DS-3 TRANSMISSION
TEST SETS

The HP 3789A and HP 3789B contain pattern generator and error detector, and operate exclusively at the DS3 rate which is the standard transmission interface in North America. The HP 3789A/B has many useful features for the digital radio engineer, including:

- Comprehensive simultaneous error analysis in all categories with limit testing.
- In-service and out-of-service tests at DS3.
- Six DSX3 outputs for loading multiple channels.
- Analog and digital auxiliary inputs.
- Results recorded on internal printer or on 80 column external printer (HP 2225A Thinkjet).
- Non-volatile memory for ease of use and results protection.
- Optional DC operation from station batteries.

In addition the HP 3789B provides jitter measurement at DS3 and can demultiplex a selected DS1 tributary allowing full in-service testing of individual traffic channels.

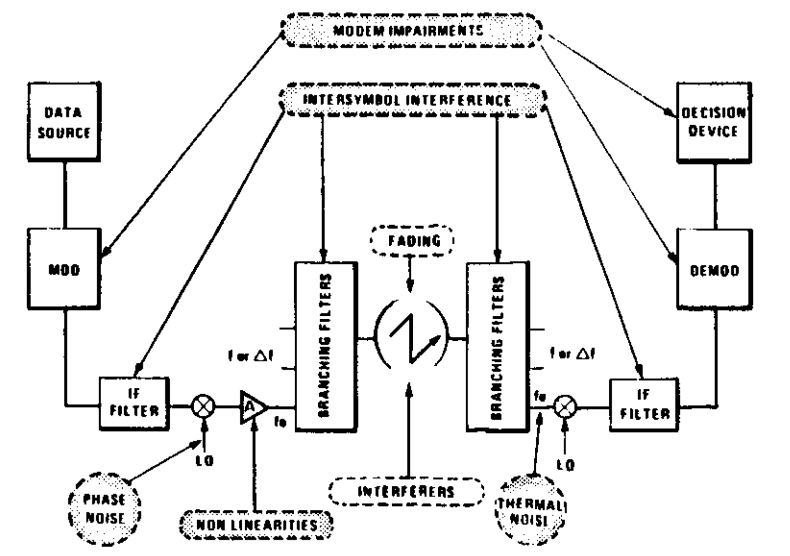
For measurements at DS-1, DS-1C and DS-2 see HP 3780A, HP 3781B, HP 3782B and HP 3787B data sheets.

Demonstrate appropriate HP Pattern Generator and Error Detector

RADIO EQUIPMENT IMPERFECTIONS

Having reviewed propagation effects and the need for statistical error analysis to measure overall system performance of an installed link, we will now consider the other half of the problem which is the imperfections in the radio equipment itself.

BLOCK DIAGRAM OF A RADIO LINK WITH IMPAIRMENTS



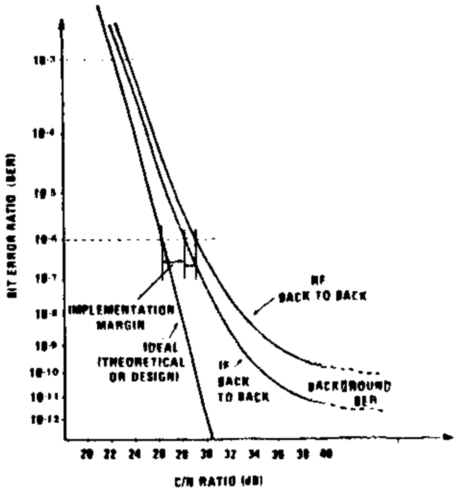
Some of these equipment impairments are illustrated in this slide, and as shown each of the imperfections is generally associated with a particular block in the radio. In the next slide the main equipment imperfections are listed, and we will discuss the measurement of these in the remainder of this seminar.

ERROR GENERATING IMPERFECTIONS IN THE TRANSMISSION EQUIPMENT

- ADJACENT CHANNEL INTERFERENCE AND SPECTRUM OCCUPANCY
- MISALIGNED TRANSMISSION CHARACTERISTICS (GROUP DELAY, AMPLITUDE RESPONSE, RETURN LOSS) CAUSING INTER-SYMBOL INTERFERENCE (ISI)
- POWER AMPLIFIER NON-LINEARITY
- PHASE NOISE (JITTER) ON RECOVERED CARRIER AND CLOCK
- THERMAL NOISE
- MISALIGNED MODULATOR AND DEMODULATOR STATES

All the above equipment imperfections cause "eye" closure and therefore reduce the radio's margins for handling low RSL and interference. Unlike propagation effects, these mechanisms give rise to long-term degradation and are responsible for low-level continuous background error ratio which is present even with high RSL. All these degradations can be quantified in terms of the Carrier to Noise (C/N) penalty or "implementation margin" resulting from the eye closure. It is these degradations that cause the practical C/N versus BER to deviate from the theoretical performance as shown in the next slide.

PRACTICAL C/N versus BER CURVES



When we look at the performance of a practical radio, the results depart from the theoretical values in the way shown in this slide. The difference between theory and practice is sometimes called the **implementation margin** and results from all the imperfections that occur in a practical radio.

The poorer the performance, the greater the required C/N for a given bit error ratio (BER). At high C/N ratio the digital radio performance becomes asymptotic to the low-level background (or dribble) BER.

We can quantify the loss of the C/N margin for each of the individual impairments in the IF and R.F. sections as shown in the example in the next slide.

(Note: The practical results shown, are for a 64-QAM radio, and are plotted on "error-function" paper which has a vertical scale such that the theoretical curve plots as a straight line. Deviations from this line for practical systems are then clearly seen.)

Comment:

Normally a radio will be worse than the theoretical curve, i.e. it will require a higher C/N ratio for a given BER. The exception is for systems using forward error correction (FEC) when the practical system can have an overall performance better than theoretical, in which case, bandwidth is being exchanged for better BER. Some line-of-sight radios use this technique, and it is quite common in satellite systems.

SYSTEM DEGRADATION BUDGET IN A TYPICAL HIGH-SPEED (90 Mb/s) RADIO SYSTEM

DEGRADATION CAUSED BY	DEGRADATION (dB)
1. MODEM IMPERFECTIONS – IF BACK-TO-BACK:	
1.1 Phase and amplitude errors of the modulator	0.1
1.2 Intersymbol interference caused by the filters in a back-to-back modem	1.0
1.3 Carrier recovery phase noise	0.1
1.4 Differential encoding/decoding	0.3
1.5 Jitter (imperfect sampling instants)	0.1
1.6 Excess noise bandwidth of receiver (demodulator)	0.5
1.7 Other hardware impairments (temperature variations, aging, etc)	0.4
Modem total:	2.5 dB
2. RF CHANNEL IMPERFECTIONS	
2.1 AM/PM conversion of the quasi-linear output stage	1.5
2.2 Band limitation and channel group delay	0.3
2.3 Adjacent RF channel interference	1.0
2.4 Feeder and echo distortion	0.2
Channel total:	3.0 dB
Total modem and channel degradation:	5.5 dB

This table shows a typical budget for the C/N penalty due to individual impairments. This indicates that the required C/N ratio will need to be 5.5dB higher than theoretical for a given BER. This means that higher transmit power or lower receiver noise figure will be required for a given fade margin or system gain. Every extra dB of transmit power is very expensive particularly when linearity is important as in QAM systems.

The C/N penalty or implementation margin is the most important measure of performance for the design and alignment of a digital radio, and later as we look at individual impairments we will always relate the measured parameter to the **equivalent C/N degradation**. For those of you who have worked with analog FM/FDM radio systems, this concept is analogous to the summation of intermodulation noise in pico-watts (pW and pW0p).

Our first task, however, is to consider how we can accurately measure the implementation margin and the practical C/N versus BER curve.

TWO METHODS OF MAKING C/N VERSUS BER MEASUREMENTS

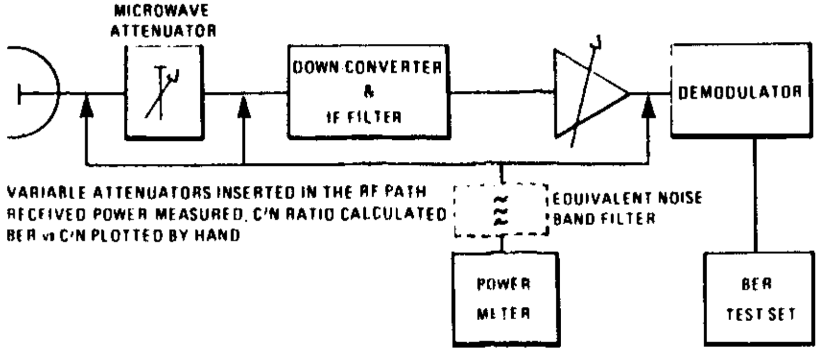
- **THE TRADITIONAL FADE METHOD USING AN ATTENUATOR AT THE RECEIVER INPUT (VARYING C)**
- **THE ADDITIVE NOISE METHOD USING A NOISE GENERATOR (VARYING N)**

We will now discuss two methods of making this measurement. Firstly the traditional method of fade simulation and secondly, the additive noise method using the HP 3708A and how it can overcome the problem experienced with the traditional method.

In either case it is important to define the carrier power as either the modulated or unmodulated signal particularly when results are to be compared (see page 29 & 30). The modulated carrier power will normally be less than the unmodulated power due to spectrum truncation in the filtering of the radio and due to AM in the modulation scheme, e.g. QAM. Differences of 1-2dB or more are normal.

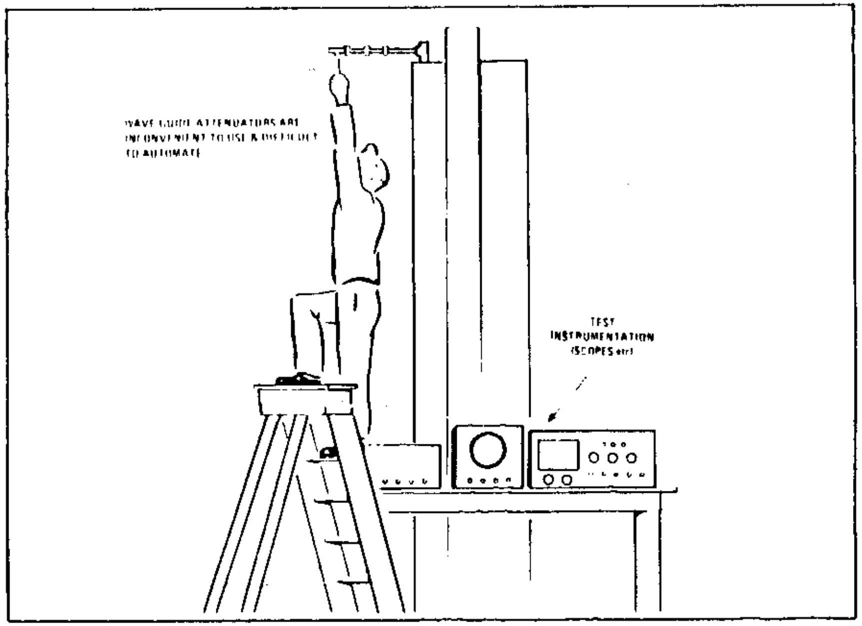
FADE SIMULATION

Traditional Method

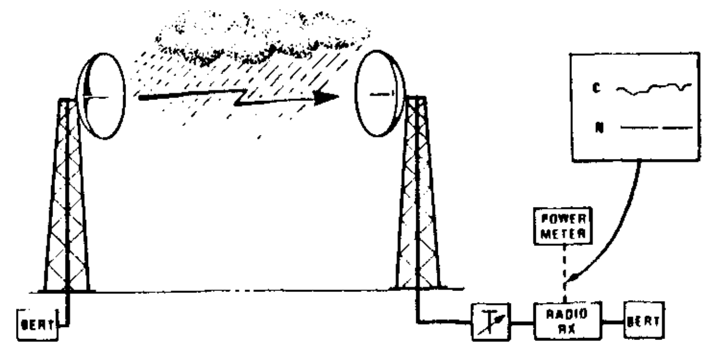


In this method an attenuator is connected at the microwave input of the receiver so that the incoming signal can be attenuated into the receiver noise. A power meter is used to check the effective C/N ratio in the I.F. strip* or to measure the received signal level (RSL) if that is the parameter being used to plot the BER curve. In the latter case, the levels are very low so it is normal to measure the signal entering the receiver and then to rely on the attenuator accuracy to set the RSL, assuming the incoming signal level has not changed! When the noise level is measured at some point in the I.F., the measured level will probably be higher than the noise level at the regenerator following the demodulator. This is because additional filtering may exist after the measurement point. This can be accounted for by making the measurement via an equivalent noise band filter at I.F.

* *Measuring the effective C/N requires a measurement of (C+N) and then N. In order to do this, it is necessary to switch the A.G.C. amplifier to "manual" and switch the transmitter off (to remove C), requiring communication and co-operation with the transmit-end and multiple measurements.*



PROPAGATION VARIATION WILL CAUSE A VARIATION IN CARRIER POWER AND CHANGE IN C/N WITH POSSIBLE LOSS OF SYNCHRONISATION AT LOW C/N

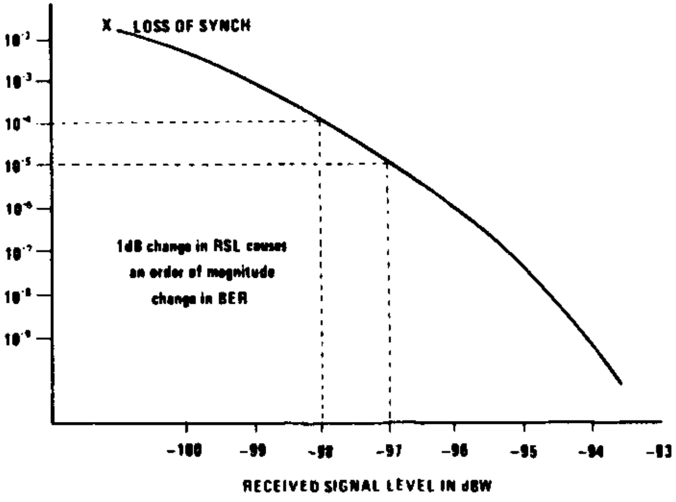


Sometimes the inaccessibility of the waveguide attenuator makes the measurement slow and it is difficult to automate. Matching problems and inherent attenuator inaccuracy at microwave frequencies reduces the reliability and repeatability of the measurements and increases the probability of operator error.

Furthermore, in the field-maintenance environment, it is undesirable to interfere with the microwave sections of the radio because of the danger of misalignment.

There is an additional problem in field measurements due to propagation effects which will cause a variation in carrier power. The noise level is fixed by the receiver front-end noise figure which to a first approximation is constant, so the C/N varies, typically 1-2dB even on a good day. This will lead to measurement errors and unstable BER readings. This is particularly true at low BER values where long gating times are required for statistically stable measurements.

A SMALL CHANGE IN RECEIVED SIGNAL LEVEL WILL CAUSE A LARGE CHANGE IN BER



The effect of the varying received level (or C/N ratio) can easily be seen on this slide where a 1dB level change (quite normal) could result in an order of magnitude change in BER.

Furthermore, as the receiver is working close to its minimum receivable level, any further reduction in carrier level, for example, a fade of 2-3dB could cause of loss of synchronisation in the radio and the results would then be completely invalid.

SOME TYPICAL PROBLEMS EXPERIENCED IN C/N TESTING USING THE TRADITIONAL METHOD

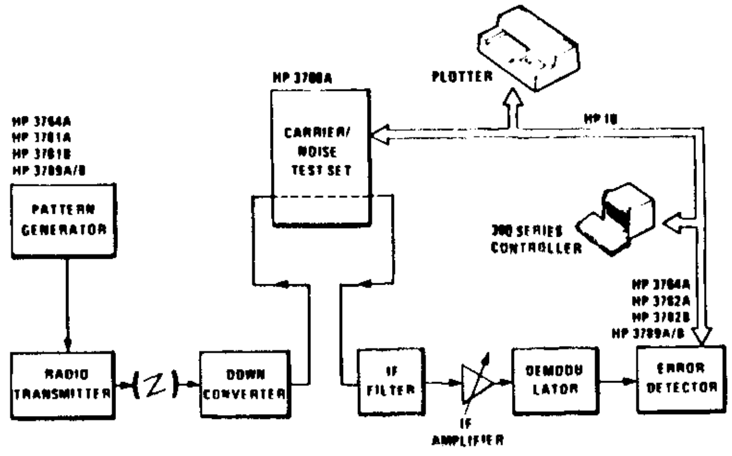
- CONFLICTS BETWEEN FACTORY AND FIELD/END USER MEASUREMENTS BECAUSE OF INACCURACY/REPEATABILITY
- UNSTABLE MEASUREMENTS IN THE FIELD DUE TO VARYING CARRIER LEVELS, AND POSSIBLE LOSS OF SYNCHRONISATION
- DIFFICULT TO RESOLVE SMALL CHANGES IN C/N MARGIN DUE TO INHERENT INACCURACY/REPEATABILITY
- TIME CONSUMING AND PRONE TO OPERATOR ERROR BECAUSE MANY ADJUSTMENTS REQUIRED TO PLOT A COMPLETE C/N CURVE (DIFFICULT TO AUTOMATE)
- UNCERTAINTY IN RELATING THEORY AND MEASUREMENTS, AND DETERMINING C/N PENALTY BECAUSE OF UNKNOWN NOISE PARAMETERS AND BANDWIDTHS

If you have tested digital radios using the traditional fading method, you will probably have experienced one or more of these problems. Most of them relate back to the uncertainty and variation in the received signal level and the inaccuracy of microwave attenuators. In other words, there are too many uncontrolled variables for repeatable measurements.

HOW DOES ADDITIVE NOISE AND INTERFERENCE TESTING USING THE HP 3708A OVERCOME THESE PROBLEMS?

We will now consider how the HP 3708A Noise and Interference Test Set can overcome these problems. Firstly, we will describe how the additive noise test is made.

ADDITIVE NOISE AND INTERFERENCE TESTING USING HP 3708A



In additive noise testing, the digital radio receiver operates at normal unattenuated levels so that the effect of receiver noise figure is negligible and possible loss of synchronisation is minimised. The IF signal in the receiver path is connected through the HP 3708A, set to the appropriate System Bandwidth, so that high crest-factor* noise can be added to the radio signal. The carrier-to-noise ratio is then accurately known and the BER is checked using the pattern generator and error detector.

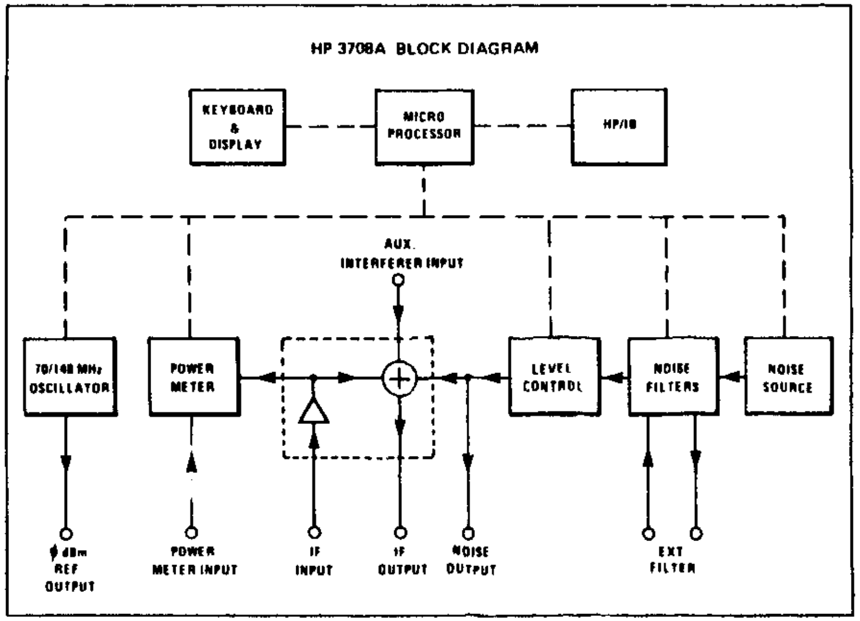
The configuration shown in this slide performs automatic measurements with HP-IB controllable BER testers. The measurements can also be made manually with a wide range of non-programmable BER testers.

* Comment:

In order to obtain accurate measurements at low BER values it is necessary for the crest factor of the noise to be >15dB. This is the minimum specified level under worst-case conditions for the 3708A and typically is better than 20dB allowing accurate measurements down to 10^{-12} BER.

For more information see:

"Additive White Noise and Interference Testing of Digital Radio Systems with Noise of Finite Crest Factor" by Geoff Waters and Ivan Young (Publication No: 5954-2007)

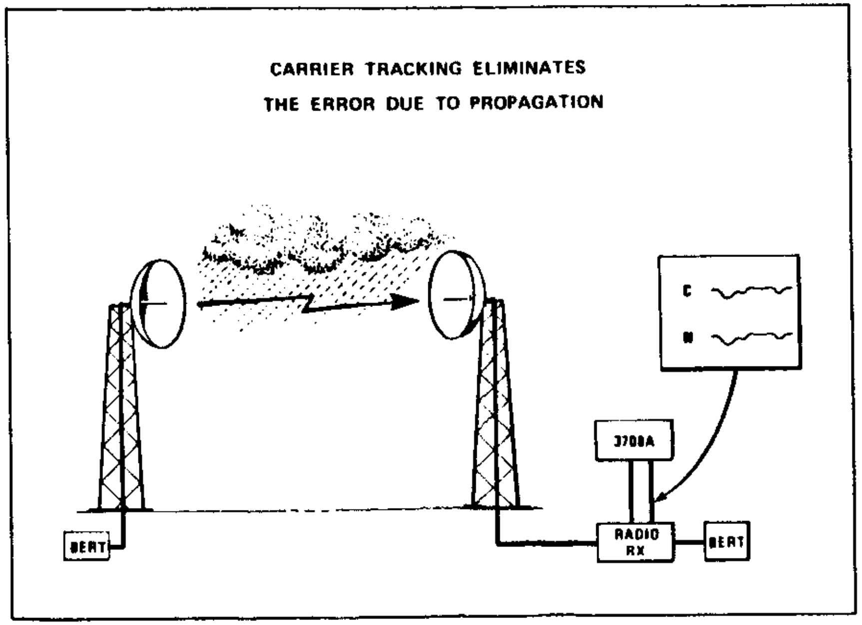


In this simplified block diagram of the HP 3708A, you can see that the radio signal is passed through the injection assembly via the I.F. input and output ports.

The injection assembly has 0dB loss to the radio signal and is specified to have negligible transmission impairments for signal levels up to +5dBm. The signal level is measured using the high-accuracy power meter and the microprocessor then automatically controls the noise source level to establish the desired C/N ratio at the I.F. output. The power meter is also designed for fast response (<10msec) so that the microprocessor can control the noise source to automatically track any variations in the received carrier level and maintain a constant C/N ratio.

Comment :

The noise level is controlled by a switched attenuator and a continuously variable PIN diode attenuator. When the C/N ratio is set-up the microprocessor sets the switched attenuator so that the PIN diode is in the middle of its range. This allows a rapid variation of the noise level over +/- 5dB without operating the switched attenuator ensuring fast response to fades.



This slide demonstrates the carrier-tracking capability where the noise level tracks the varying carrier level to keep a constant C/N ratio over the measurement period.

3708A CARRIER TO NOISE RATIOS

1. CARRIER POWER TO NOISE POWER C/N dB
2. CARRIER POWER TO NOISE DENSITY C/No dB Hz
3. ENERGY PER BIT TO NOISE DENSITY E_b/No dB

$$No = \frac{N}{B_e} \qquad E_b = \frac{C}{f_b}$$

$$C/No \text{ dB Hz} = (C/N) \text{ dB} + 10 \log_{10} (B_e)$$

$$E_b/No = (C/N) \text{ dB} - 10 \log_{10} \left(\frac{f_b}{B_e} \right)$$

B_e = RECEIVER NOISE BANDWIDTH IN Hz

f_b = BIT-RATE IN Hz OR BITS/SEC

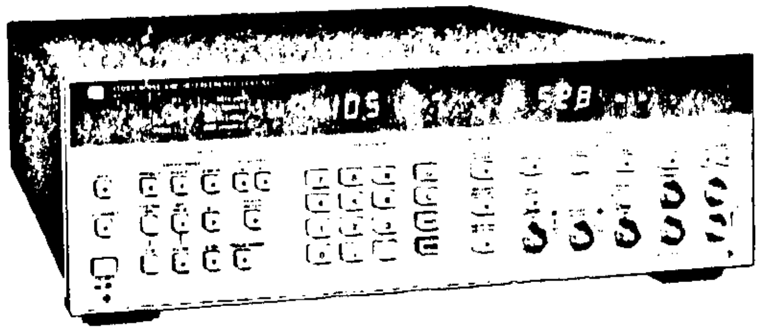
There are three units commonly used in carrier to noise testing available on HP 3708A: carrier to noise (C/N), carrier to noise-density (C/N_0) and energy-per-bit to noise-density (E_b/N_0). C/N_0 and E_b/N_0 are independent of bandwidth and particularly useful in R&D work for comparing the efficiency of different systems. These measurements require the test-set to have an accurately known noise-density for injection into the radio I.F.

C/N needs to be related to a specified system bandwidth, for example, the bandwidth of practical receiver filters or to the theoretical minimum Nyquist bandwidth (or symbol-rate bandwidth) (see page 29).

The HP 3708A is designed with accurate noise bandwidth filters so that noise-density is calibrated. The HP 3708A can also measure the noise bandwidth of external I.F. filters.

See Product Note 3708-1 (Publication No. 5953-5481) for more details on the theory.

HP 3708A NOISE & INTERFERENCE TEST SET



This slide shows the front-panel of the HP 3708A Noise and Interference Test Set. On the left-hand side you can see the keys that control the modes of operation and the measurement units. You will notice a key in this area labelled **SYSTEM BANDWIDTH** which allows the entry of a user defined bandwidth for referring the C/N ratio value (for example, the symbol-rate bandwidth or receiver bandwidth).

On the right-hand side, you can see the main I.F. input and output connectors and the selection of internal filter bandwidths. The bandwidth quoted above each key is the flat bandwidth. Normally the band of noise chosen should be wider than the radio I.F., the bandwidth being subsequently defined by the radio filtering. The HP 3708A can test radio systems using 70 MHz and 140 MHz IFs.

The standard instrument has 75 ohm interfaces, 50 ohm version available as Option 001.

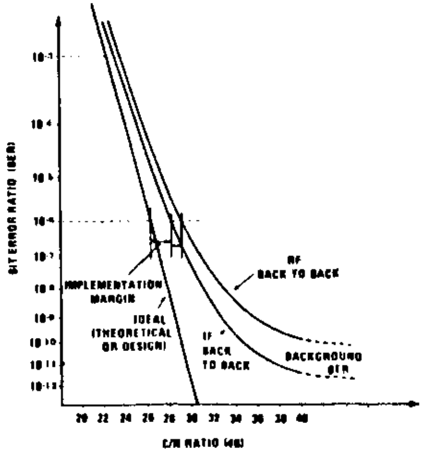
See Product Note 3708-2 for more details on using the HP 3708A.

THE HP 3708A OVERCOMES THE PROBLEMS DESCRIBED EARLIER BY THE FOLLOWING ADVANTAGES:

- **ACCURACY AND REPEATABILITY BETTER THAN 0.1 dB TYPICALLY**
- **ACCURACY INDEPENDENT OF VARYING CARRIER LEVEL IN FIELD APPLICATIONS**
- **SIMPLE OPERATION AND AUTOMATION SAVES TIME AND REDUCES OPERATOR ERROR**
- **ACCURATE NOISE DENSITY AND SYSTEM BANDWIDTH DEFINITION**

The accuracy and repeatability of the HP 3708A test method means that measurements can be made on individual radios at different times and results used for reliable comparison of degradation in performance.

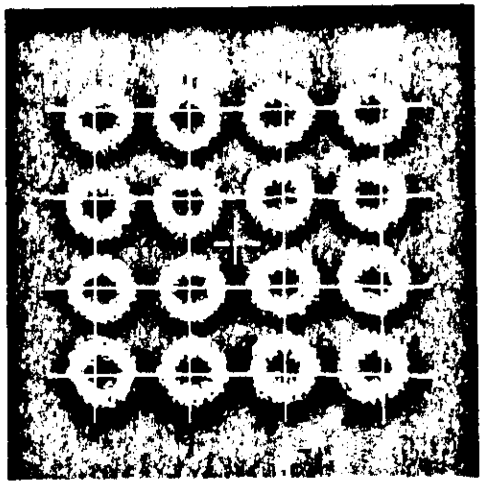
PRACTICAL EFFECTS OF IMPERFECTIONS ON THE C/N VS BER CURVE



So the practical BER versus C/N curves, shown here, can be accurately and quickly plotted using the HP 3708A. The only remaining problem is the long gating time (maybe several days) required in the BER test in order to measure the very low background BER at high C/N values. It is important to know the background BER as when many radios are cascaded, the overall error performance may become unacceptable. This is particularly relevant now that the latest version of CCIR Rec. 594 includes the RBER specification (page 38). An overall value of 5×10^{-9} for 2500 km requires a single hop specification of 10^{-10} to 10^{-11}. Clearly the test is very time consuming, particularly if a re-test is required after adjustment.

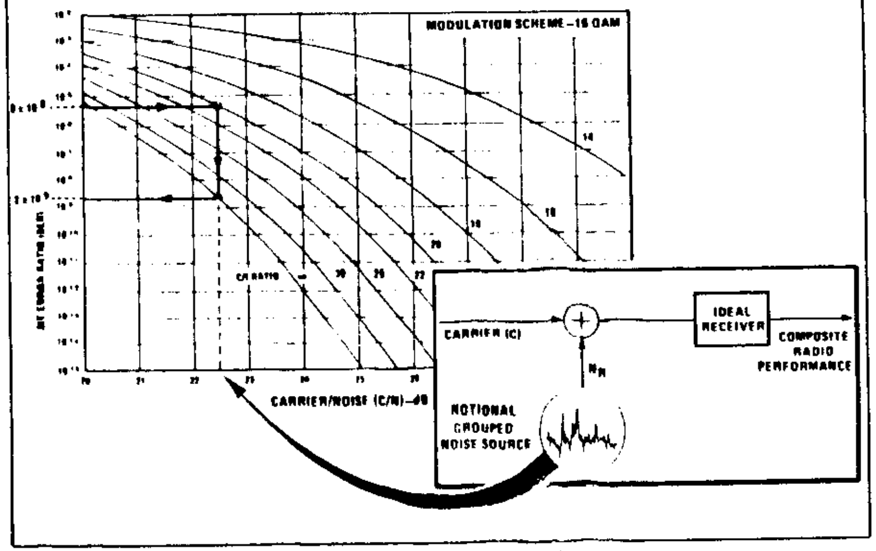
The measurement time can be substantially reduced by using a C/I test in which a sinusoidal interfering signal is added to produce a controlled amount of "eye" closure. The increased BER is then easily measured and the true background BER can be estimated. This effect is shown in the next slide.

C/I TESTING USING SINUSOIDAL INTERFERING TONE AT 71.5 MHz



This constellation display of a 16-QAM digital radio with an interfering tone added shows clearly the reduction in space between the individual phase-states which causes the increased BER value. Estimation of the true background error ratio is made using the graphs illustrated in the next slide.

GRAPHICAL METHOD OF ESTIMATING RESIDUAL BER



The principle of this technique depends on representing all the imperfections in the radio as equivalent to a notional grouped noise source which is summed into an ideal receiver. A level of C/I is chosen (in this 16 QAM example C/I = 20dB) which gives a reasonable BER with the radio under test (e.g. 10^{-3} to 10^{-6}). From the BER value and C/I value we can then find the equivalent C/N ratio for the notional internal grouped noise source (in this example 22.5dB). If we assume this has a gaussian distribution, then we can estimate the residual BER from the intersection with the C/I = infinity curve as shown in this slide. The curves are calculated theoretically for an ideal radio with varying amounts of sinusoidal interference. These allow the estimation of the equivalent notional noise source and from that the effective background BER from the C/I= infinity curve.

Comment:

There are a number of assumptions in this technique which are described more fully in these two HP publications:-

Product Note 3708-3 "Determination of Residual Bit Error Ratio in Digital Microwave Systems" (Pub No 5953-5490)

"Methods for Estimating Residual BER in Digital Radio" by Ian Kennedy Compston (Pub No 5954-7942).

DETERMINATION OF RADIO RESIDUAL BER

MAJOR ADVANTAGES OF HP 3708A C/I TEST METHOD

- **REDUCED MEASUREMENT TIME**

  EXAMPLE : 100 Mbit/s DIGITAL RADIO
  RESIDUAL BER (APPROX) 10^{-12}

  TRADITIONAL METHOD TEST TIME : 3 DAYS (APPROX FOR 30 ERRORS)
  HP 3708A METHOD TEST TIME : 2-3 MINS (APPROX)

- **MEASUREMENT ACCURACY**

  IN FIELD MEASUREMENTS, THE PROPAGATION CONDITIONS ARE UNLIKELY TO REMAIN STABLE OVER A 3 DAY MEASUREMENT PERIOD.

The simple calculation on this slide demonstrates the time-saving possible using the C/I test. Estimation of residual BER becomes feasible in field installations and test-time can be saved in production, allowing repeated tests to be made for evaluation of adjustments.

In high-density modulation schemes such as 16-QAM and 64-QAM, excessive background BER can be a significant problem.

HP 3708A SOFTWARE

- FULLY AUTOMATIC PLOTTING OF C/N AND C/I CURVES INCLUDING SEQUENCING
- STORAGE OF MEASUREMENT CURVES FOR COMPARISON
- FLEXIBLE PLOTTING FORMAT INCLUDING USE OF GRAPH PAPER
- STORAGE OF MEASUREMENT CONFIGURATION AND PARAMETERS FOR EASY MEASUREMENT EXECUTION

The HP 3708A software demonstrates the power of a fully automated system incorporating HP 3708A, HP 3764A or HP 3789A/B, 300 Series Controller and plotter. Once the software is loaded, measurements can be initiated and results plotted and stored with a few key-strokes. Automatic measurements can be left running unattended, or the sequence repeated several times to detect variations in radio performance.

For more details see:

Product Note 3708-4 "HP 3708A Demonstration Guide" (Publication No 5954-9551).

Demonstrate the HP 3708A and the system

MEASURING INDIVIDUAL IMPAIRMENTS

Having reviewed the overall measure of digital radio performance, the C/N versus BER test, we will now consider in more detail the measurement of individual impairments.

ERROR GENERATING IMPERFECTIONS IN THE TRANSMISSION EQUIPMENT

- ADJACENT CHANNEL INTERFERENCE AND SPECTRUM OCCUPANCY
- MISALIGNED TRANSMISSION CHARACTERISTICS (GROUP DELAY, AMPLITUDE RESPONSE, RETURN LOSS) CAUSING INTER-SYMBOL INTERFERENCE (ISI)
- POWER AMPLIFIER NON-LINEARITY
- PHASE NOISE (JITTER) ON RECOVERED CARRIER AND CLOCK
- THERMAL NOISE
- MISALIGNED MODULATOR AND DEMODULATOR STATES

This slide shows some of the more common equipment imperfections the overall effect of which can be evaluated by the C/N versus BER test.
The C/N versus BER test yields some diagnostic information as it is sometimes possible to check parts of the radio (for example the modulator/demodulator back-to-back, or a transmitter against a standard receiver, or to separate the implementation margin of the microwave section and transmit amplifier, and so on). However, we often need to measure an impairment directly and interpret the result in terms of C/N penalty contribution to the implementation margin, and in the remaining part of the seminar we will look at some of the impairments listed on this slide.

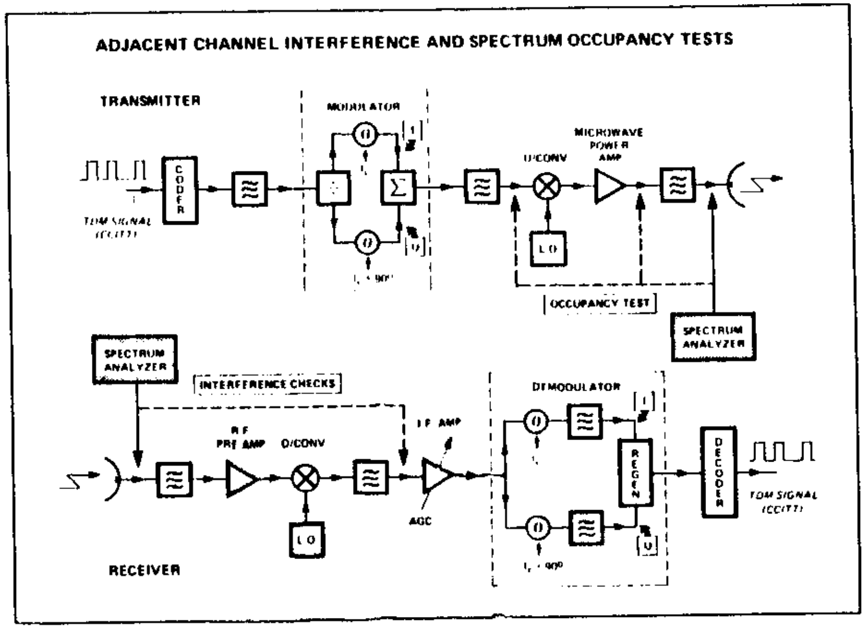
Comment:

In the theoretical analysis given for each individual impairment, the C/N degradation is calculated assuming that the impairment is present on an otherwise perfect radio. With a practical system which inevitably has a variety of imperfections causing eye-closure, the sensitivity to a specific impairment will be increased and the C/N degradation greater than predicted. The curves shown in the following pages should be treated, therefore, as the minimum expected C/N degradation, or conversely the equivalent level of impairment should be taken as an absolute maximum specification.

SPECTRUM OCCUPANCY AND INTERFERENCE

- **ADJACENT CHANNEL INTERFERENCE FROM TRANSMITTER (SPECTRUM OCCUPANCY)**
- **CHECK LEVELS OF INTERFERENCE AT RECEIVER INPUT**
- **MEASURE SUSCEPTIBILITY TO INTERFERENCE**

The first group of impairments all relate to the generation of interference by the transmitter and the effects of interference at the receiver.



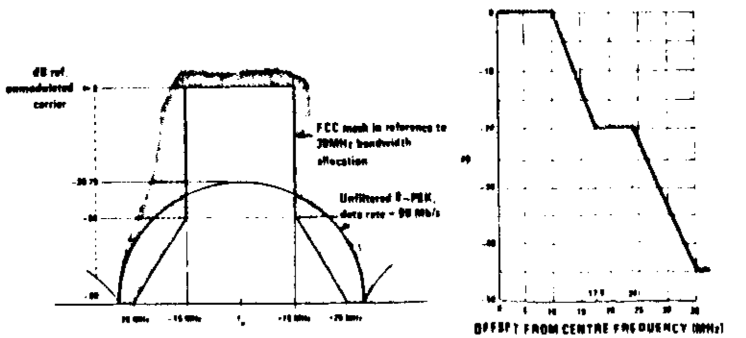
These two tests are generally made with a spectrum analyzer. The Spectrum Occupancy test is a measure of how well unwanted sidebands and spurious signals have been suppressed by the successive filters in the transmitter. To minimize interference to adjacent radio channels it is very important that the radio complies with the occupancy mask laid down by the local regulating authority (e.g. the FCC in the USA or a PTT in a European country etc.). Some examples are shown in the next slide.

The levels of interference present at the receiver can also be checked using a spectrum analyzer with the associated transmitter switched off. Sources of interference include:

- **Adjacent Channel** - due to poor out-of-band suppression from adjacent transmitters.
- **Co-channel** - from another radio on the same frequency possibly using an opposite polarization.
- **External Sources** - such as radar systems.

Interference causes eye-closure in the demodulator and results in a C/N penalty or loss of receiver sensitivity.

TYPICAL OCCUPANCY MASKS

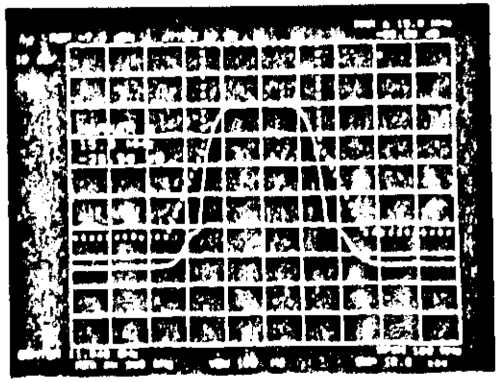


FCC MASK FOR 30 MHz CHANNEL ALLOCATION
(MASK IS REFERRED TO THE LEVEL IN A 4 kHz BANDWIDTH
WITH REFERENCE TO THE UNMODULATED CARRIER POWER)

UK MPT1403 MASK FOR 13 GHz 34 Mb/s RADIO
(THIS IS A POWER DENSITY MASK, INDEPENDENT OF
MEASUREMENT BANDWIDTH)

This slide shows two typical occupancy masks in use in the industry. The out-of-band response is measured relative to either the unmodulated carrier (or total mean power level of the spectrum) or the mid-band power density (dBm/Hz). An example of the first case is the FCC mask and for this test it is necessary to take account of the noise bandwidth of the spectrum analyzer filters. The relative power-density type of mask (e.g UK MPT 1403) is independent of the filter bandwidth provided a sensible filter is chosen (i.e. less than 10% of symbol-rate bandwidth).

OUTPUT TRANSMIT SPECTRUM OF A 90 Mb/s 16QAM RADIO WITH FCC 30 MHz MASK OVERLAYED



THE VERTICAL POSITION OF THE MASK HAS BEEN ADJUSTED TO TAKE ACCOUNT OF MEASUREMENT BANDWIDTH/FCC BANDWIDTH (+19.5dB) AND SPECTRUM ANALYZER DETECTOR (−2.5dB)
THE REFERENCE LEVEL HAS BEEN SET TO THE AVERAGE POWER LEVEL OF −7.5 dBm.)

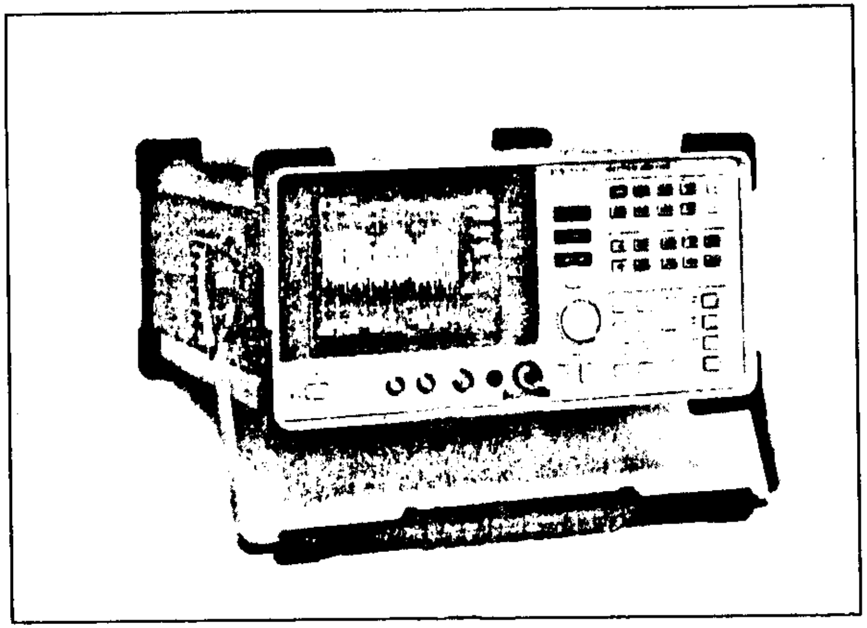
This slide shows the measurement of a 16-QAM radio against the FCC 30MHz mask. The top of the mask is referenced to the average power level of -7.5dBm. This value can be calculated from the spectrum picture by knowing the noise bandwidth of measurement filter or it can be measured directly with a power meter.

This display also illustrates a potential cause of operator error: the displayed level of the spectrum appears to be -27.5dBm so there is a temptation to set the input attenuator for a low-level signal when in fact the power at the analyzer input is -7.5dBm. The measurement would be in error because of intermodulation and compression.

For more information on the many uses of spectrum analyzers see:

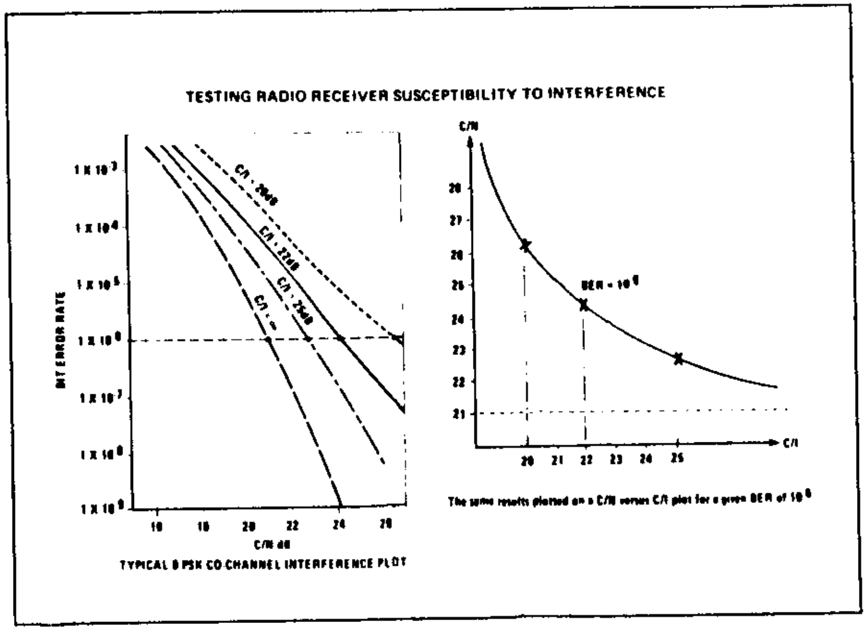
"Practical Applications of Spectrum Analyzers for Digital Radio Measurements" by Jim Kaylor and Hugh Walker (Pub No 5954-7944)

Hewlett-Packard produces a wide range of Spectrum Analyzers from portable instruments for field use to very high-performance fully- programmable and synthesized instruments for R&D and Manufacturing applications. Most instruments can provide a hard-copy output to a plotter, and the latest products can store instrument settings and masks for simple go/no-go operations.



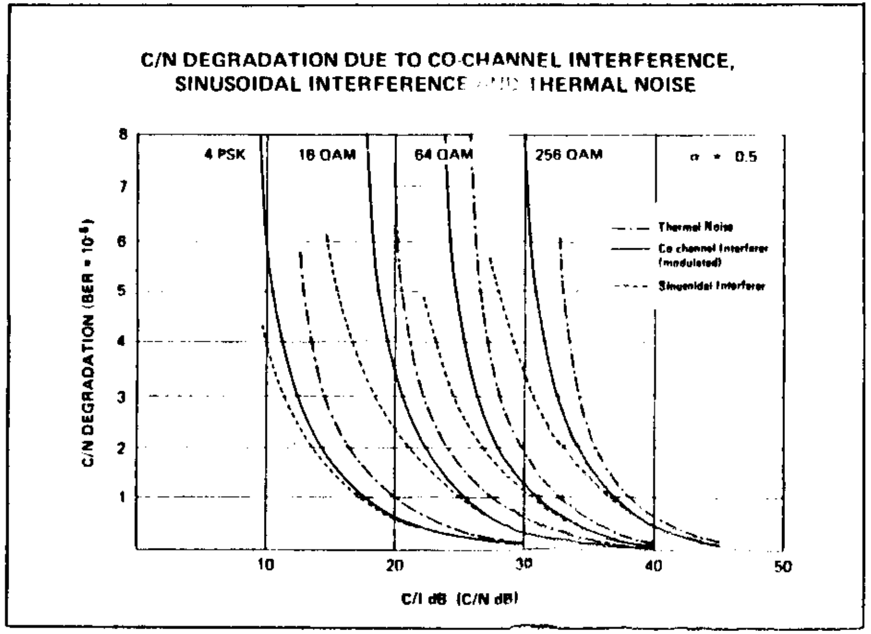
The HP 8562A/B portable spectrum analyzers (1 kHz to 22 GHz) make ideal test tools for digital radio, particularly in the field environment since they meet military requirements for ruggedness. The synthesized tuning, trace storage and averaging, and the capability to recall from non-volatile memory, instrument configurations by user-defined names, ensure easy, reliable and accurate measurements of digital radio signals. Many of the detector and filter characteristics are automatically calibrated and a plug-in test-and-adjustment module speeds troubleshooting and adjustments.

Demonstrate an appropriate Spectrum Analyzer



It is sometimes necessary to specify the performance of the receiver with varying levels of adjacent channel or co-channel interference by plotting a family of C/N versus BER curves for varying levels of interference as shown in this slide. For a given BER (e.g. 10^{-6}) the effective loss of receiver sensitivity (or C/N penalty) can be measured as a function of C/I so that it is then possible to plot a C/N versus C/I curve as shown in the slide.

The HP 3708S Noise and Interference Test System is particularly valuable in this test, as multiple C/N versus BER curves can be plotted automatically saving a great deal of time. This interference can be injected either at the microwave receiver input, or at I.F. via the wideband Auxiliary Interferer Input on the HP 3708A.



To complete the section on interference, here are some theoretically generated curves*, for various modulation schemes, showing the carrier to noise degradation resulting from varying levels of in-band interference. As you might expect, the more complex modulation schemes require higher levels of carrier to interference (C/I) for satisfactory operation. You will also see that the degradation varies depending on the type of interference. Sinusoidal or CW interference causes the least degradation, an uncorrelated co-channel modulated interferer rather more, and thermal noise the most.

<u>Comment</u>:

For moderate levels of interference it is often acceptable, for testing and analysis, to treat the co-channel interference as thermal noise - use a HP 3708A rather than a digital radio. As you can see, this gives a slightly pessimistic answer - sometimes referred to as the thermal noise approximation (TNA). Many digital radio impairments present in small amounts can be represented by an equivalent noise source - this is the basis of the RBER prediction method described earlier (page 51).

* *"Distortion Analysis of Digital Radio-Relay Systems using 4-256 QAM" by Biester and Glauner, European Radio Relay Conference, Munich, November 1986.*

TRANSMISSION IMPAIRMENTS

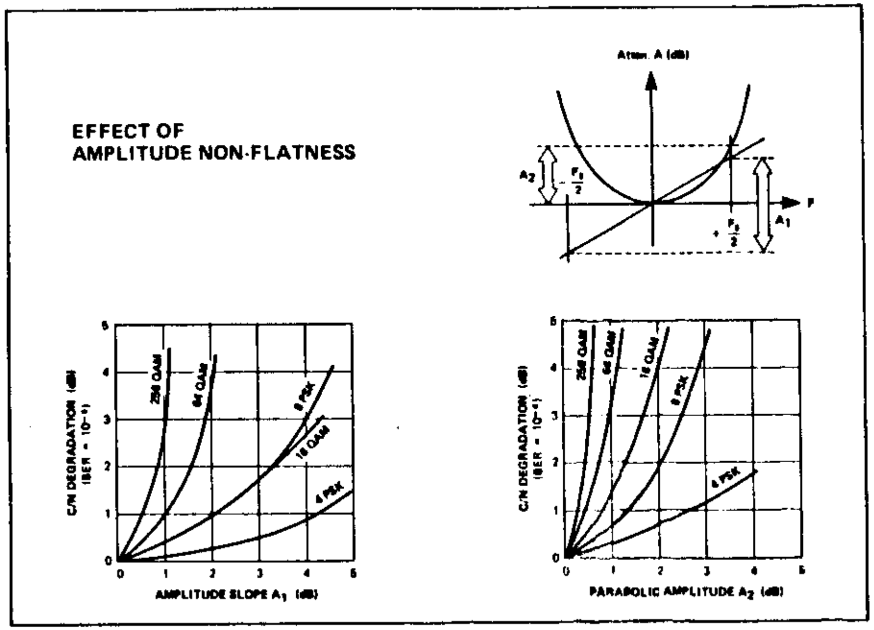
- **RADIO FILTERS**
- **WAVEGUIDE/ANTENNA AND PROPAGATION EFFECTS**

We will now consider the second item on the list of equipment imperfections - misaligned transmission characteristics such as amplitude response, group delay and return loss.

In the theory section (page 19) we saw the importance of having the right overall transmission characteristic of the raised-cosine amplitude response with flat group-delay. If the response departs from these characteristics the result will be ISI and degradation in C/N margin.

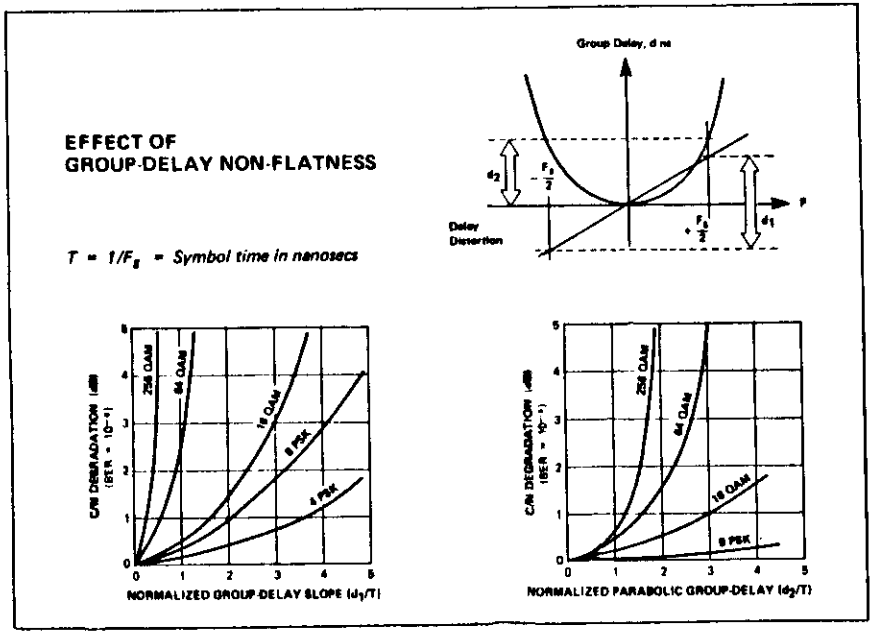
Imperfections can occur in the filters of the radio, in waveguide and antenna sections and, of course, in the line-of-sight path.

The next three slides will give you an idea of the effect of such imperfections on the implementation margin or C/N penalty.



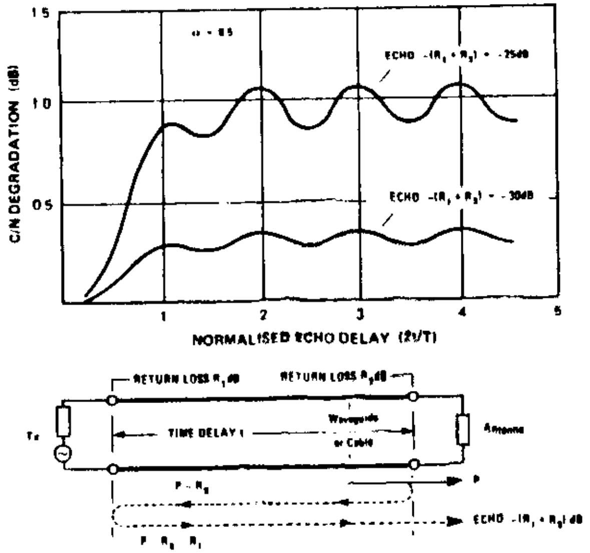
This slide shows the carrier to noise degradation resulting from amplitude slopes and parabolic amplitude in the radio passband. The amplitude response is measured across the symbol rate bandwidth and the C/N degradation interpreted from the curves depending on the modulation scheme. Notice how much more sensitive the higher modulation schemes are - if you are working with 64QAM, amplitude response must be better than 0.5 dB across the band.

These curves also show why the higher modulation schemes are so sensitive to multipath fading.



This slide shows a similar set of curves for group-delay non-flatness. We measure the group-delay in nanoseconds over the symbol rate bandwidth and normalise the result by dividing by the symbol time T (= 1/Fs). We can then read off the C/N degradation depending on the modulation scheme. You can see that group-delay slope is the most important effect, and for example on a high-capacity 64 QAM radio, it would be necessary to equalize the group delay to better than 0.5 nsecs.

C/N DEGRADATION DUE TO ECHOES ON A 16 QAM SYSTEM



What is the effect of reflections or echoes in digital radio? These may be due to poor return loss in filters, waveguide imperfections, antennas and moding effects in waveguides particularly during multipath fading. The analysis is quite simple as this slide for a 16 QAM example shows. If the echo or reflected signal is delayed by more than one or two symbol times (T to 2T) then the echo signal can be treated as an uncorrelated co-channel interferer with a C/I value equal to the sum of the two return losses, $(R_1 + R_2)$ as shown in this slide. This is the worst case limiting condition. To evaluate the effect on C/N degradation we refer to the C/I plots on page 57 and read off the degradation for the particular modulation scheme. Alternatively, if we treat the reflected signal as noise with a C/N ratio equal to $(R_1 + R_2)$, then we can read off the theoretical C/N versus BER curves (page 29) the resulting residual BER.

You can see that complex modulation schemes require quite good return loss (20 - 25 dB for 64 QAM, 25 - 30 dB for 256 QAM). Remember there is a cumulative effect in the transmit and receive waveguide/antenna systems in each hop.

In the next slides we will look at how transmission measurements are made.

For more information on the source of these curves see:

"Modulation Techniques for Microwave Digital Radio" by Noguchi; Daido and Nossek. IEEE Communications Magazine, Oct 1986, Vol 24, No 10, pp21-30.

"Distortion Analysis of Digital Radio-Relay System Using 4 to 256 QAM" by Biester and Glauner, European Radio Relay Conference, Munich, November 1986.

"Effects of Waveguide Echoes on Digital Radio Performance" by Wu and Achariyapaopan, Globecom 85, paper 47.5.

TWO CLASSES OF TRANSMISSION MEASUREMENTS

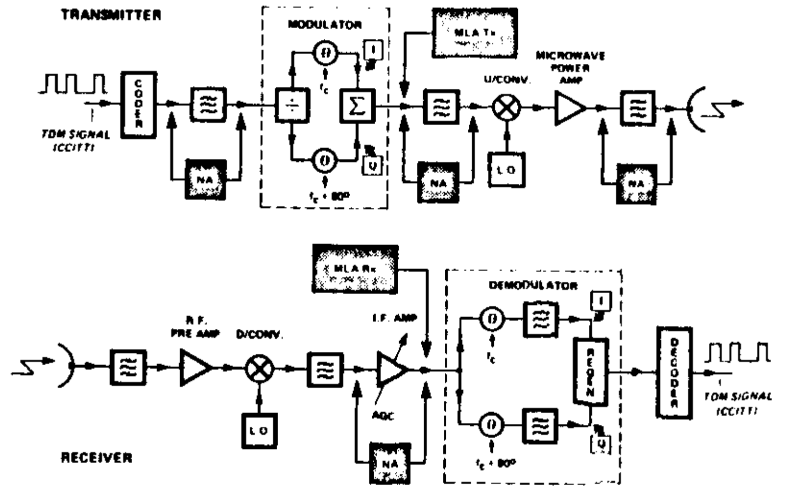
- **COMPONENTS SUCH AS FILTERS, EQUALISERS, AMPLIFIERS ETC**
- **'END-TO-END' TESTING OF COMPLETE RADIO SYSTEM**

Component testing is usually undertaken with a network analyzer and can be a fully automated measurement at the test bench.

"End-to-end" measurements require a Microwave Link Analyzer (MLA) which is able to recover its own phase reference for group delay measurements as distinct from the network analyzer which requires a local reference. The MLA can make measurements across a complete installed radio link, and across frequency translations.

These applications are illustrated in the block diagram of the next slide.

TRANSMISSION MEASUREMENTS



Network Analyzers (NA) are used for aligning subassemblies such as filters, amplifiers, combiners and so on, and because they operate over wide bandwidths, are useful for checking the out-of-band response of filters.

Two types of network analyzers are in use:-

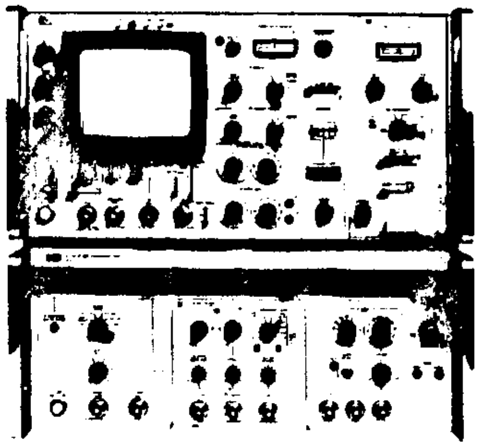
Vector Analyzers - which measure group-delay as well as amplitude response and return loss (reflection).

Scalar Analyzers - which measure only amplitude response and return loss.

Because of the higher cost of microwave vector analyzers, scalar analyzers are often used for measuring components such as waveguide filters, combiners, waveguide matching units etc., the assumption being made that when the amplitude response and return loss are optimized, the phase response or group-delay will be correctly aligned (for a minimum-phase network).

> *Hewlett-Packard produces a very wide range of scalar and vector network analyzers for B.B., I.F. and microwave bands, many of which have become industry standards of performance.*

HP 3711A/12A MICROWAVE LINK ANALYZER AN INDUSTRY STANDARD



When "end-to-end" measurements are required across a complete radio system including frequency translations, a Microwave Link Analyzer (MLA) must be used. The HP 3711/12A MLA operates at both of the commonly used I.F.'s of 70MHz and 140MHz with sweep widths of 50MHz and 100MHz respectively.

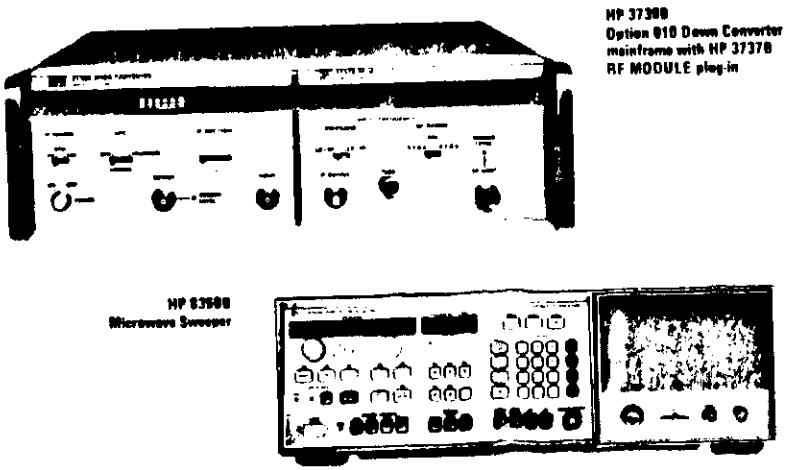
It has very low residual non-flatness in amplitude response and group-delay and furthermore performance is guaranteed between any HP 3711A Transmitter and HP 3712A Receiver (not just pairs). This is important in "end-to-end" testing where it is not possible to "normalize" out intrinsic non-flatness, a technique commonly used in network analyzer measurements.

The HP 3711A/12A is also well suited to testing non-flat digital radio filters, since the intrinsic AM/PM conversion in the MLA receiver is held at a very low level. If AM/PM is present in the receiver, spurious group-delay slopes would appear when measuring devices with non-flat amplitude responses common in digital radio. The HP 15570B AM/PM test network enables the MLA receiver to be checked very precisely ensuring accurate group-delay measurements at all times.

For more information see:

"Technique Simplifies AM-to-PM Measurements" by I. Matthews, Microwaves and RF, August 1984, pp108-112.

EXTEND MLA MEASUREMENTS TO MICROWAVE FREQUENCIES

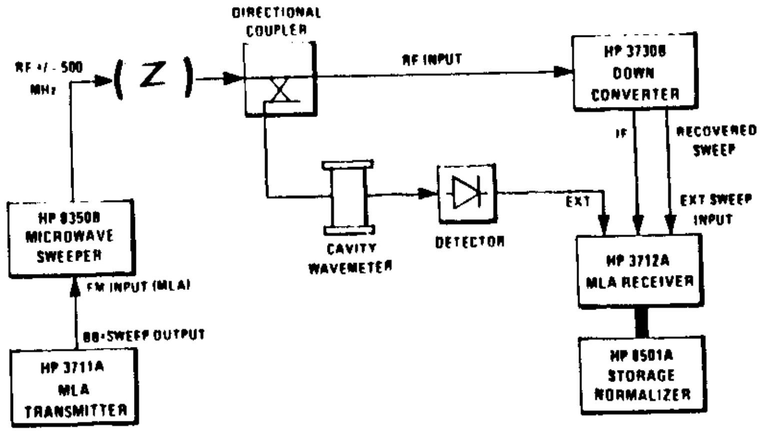


The flexibility of MLA measurements can be extended to microwave frequencies using the HP 3730B Down Converter and HP 8350B Microwave Sweeper. A range of plug-ins is available for these units covering the various microwave radio bands from 1.7GHz to 14.5GHz. Consult the HP 3730B and HP 8350B Data Sheet for more information.

Using these instruments, amplitude response and group-delay can be measured in the transmitter up-converter, receiver down-converter, in waveguide runs, antenna systems and the complete microwave path.

The next slide shows a practical application of these measurements using the HP 3730B Down Converter.

SYSTEM CONFIGURATION FOR SWEPT MEASUREMENTS OVER BANDWIDTHS UP TO 1 GHz



This slide shows a measurement set-up for evaluating wideband devices such as antennas and for checking the transmission path end-to-end over a radio hop. Such tests might be used for identifying multipath effects.

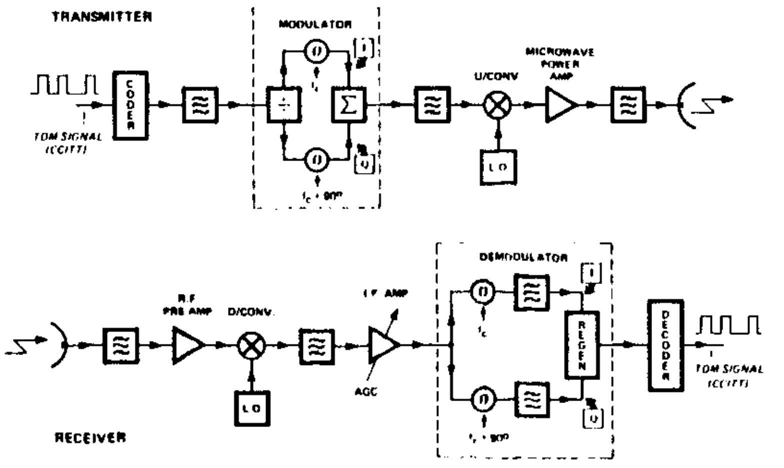
The system is capable of sweep widths of up to 1GHz at microwave which covers the bandwidth of most microwave radio bands. The HP 8350B Microwave Sweeper is modulated directly with the MLA BB + Sweep Signal. The novel aspect of this configuration 's the use of the HP 3730B Down Converter which recovers the sweep and compresses the 1GHz sweep width down to an I.F. bandwidth which can be handled by the MLA. Because of this compression the MLA markers are no longer useable, so a cavity wavemeter is used for frequency identification.

For more information see Product Note 3730-1 (Pub No 5953-5468)

POWER AMPLIFIER NON-LINEARITY

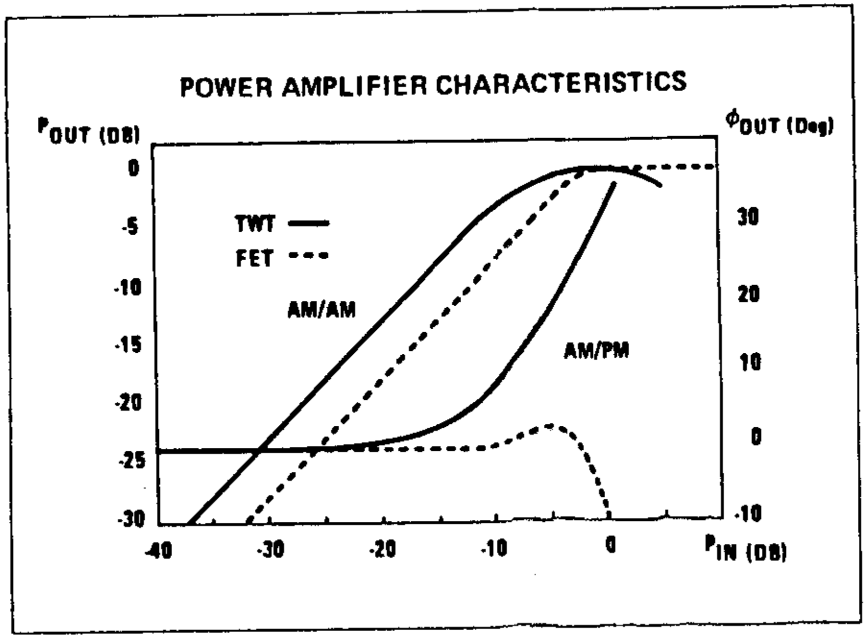
The third impairment we will consider is the effect of the non-linear distortion on the digital radio signal. This is a serious problem particularly in more complex modulation schemes such as 16-QAM and 64-QAM.

DIGITAL RADIO BLOCK DIAGRAM

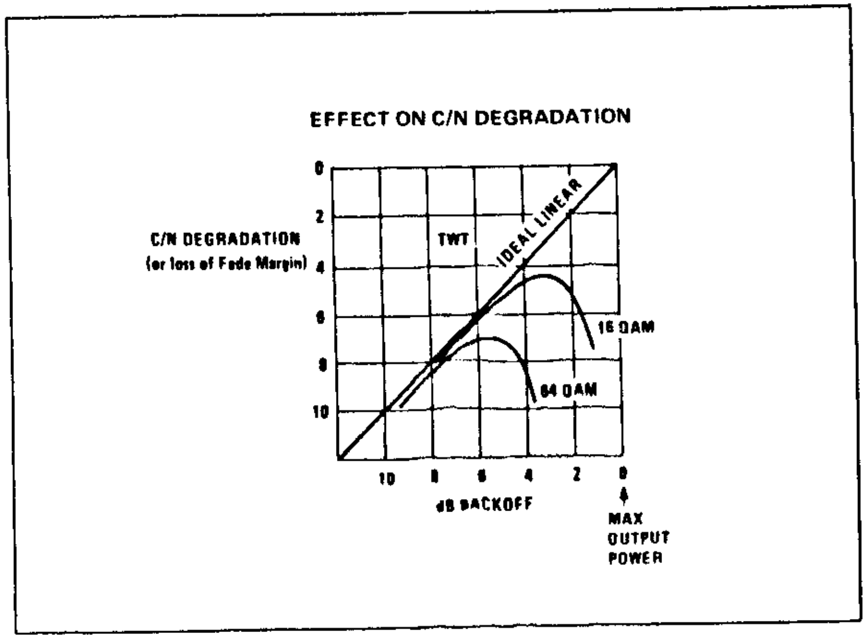


Almost all digital radio modulated signals exhibit **amplitude modulation** (AM) either intentionally as in QAM or as a result of spectrum truncation as in filtered 4-PSK. When these signals pass through a non-linear amplifier they are compressed and distorted and inter-symbol interference (ISI) and phase-state displacement result with a loss of C/N margin.

These effects can take place at any point in the carrier section of the radio where there is a non-linear device, but predominantly in the transmit amplifier because of the high-power operation. In complex modulation systems such as 64-QAM and 256-QAM it may also be necesary to check non-linearity in other sections such as up and down converter and A.G.C. amplifier.



The non-linearities take two forms. Firstly, so called AM/AM which is the familiar compression characteristics where the linear relationship between input power and output power bends over as the amplifier saturates. The second form is conversion between amplitude modulation and phase modulation or AM/PM conversion. In this case as the power increases the relative phase angle of the output changes. This is serious for digital radio as information is contained in the phase relationship. This slide shows the kind of characteristics which might be expected in travelling wave tube (TWT) and field effect transistor (FET) power amplifiers.



This slide shows the typical effect on the C/N margin or fade-margin of the radio when the power amplifier is working in a non-linear region. If the amplifiers were perfectly linear, the more output power we could transmit the greater would be the fade margin of the radio (Ideal Linear Line). Because of the non-linearities in the practical amplifier, the increasing distortion at higher power levels causes more ISI so we enter a region of "diminishing returns". Eventually the curve turns over as the distortion becomes very severe and further increase in transmit power actually results in a lower fade margin! (Analogue radio engineers may recognise a similarity to the familiar white noise V-curve test.)

You will notice the horizontal axis is calibrated in dB of "backoff" which is the amount the output power is reduced below the maximum available from the TWT. Typical "backoffs" of 10dB or more are used in complex modulation. Normally the power amplifier is preceded by a predistorter (usually at I.F.) which tries to produce the reverse AM/AM and AM/PM characteristic and cancel the non-linearity. The adjustment of these devices is quite critical in 16-QAM and 64-QAM systems.

The effects in 4-PSK are far less serious, which is another reason for its popularity in satellite systems where very high transmit powers are required.

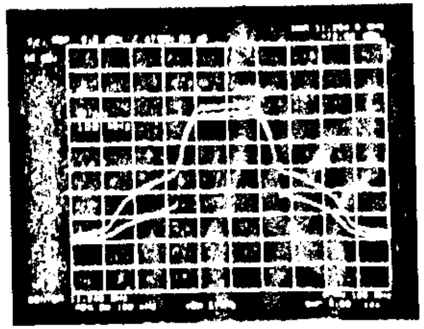
METHODS OF MEASURING NON LINEARITY AND ADJUSTING PREDISTORTER

- TWO- OR THREE-TONE INTERMODULATION TEST
- MEASURE UNWANTED SIDEBAND LEVEL
- AM/PM MEASUREMENT USING TEST NETWORK
- CONSTELLATION MEASUREMENTS

Each of the methods shown on this slide are in use for measuring non-linearity and adjusting predistorter settings. The first two are straightforward distortion measurments, and assume a strong correlation between AM/AM and AM/PM conversion but do not measure phase discrepancies directly. The AM/PM effect can be measured directly using a test network described later but does not detect AM/AM effects precisely. The problem with the first three tests is that they provide an incomplete analysis and the dynamics of the test signal may not accurately simulate the true digital radio modulation. Optimization using these methods will not necessarily give the best overall radio performance in terms of C/N margin. Probably the best method is the constellation measurement described later (page 78).

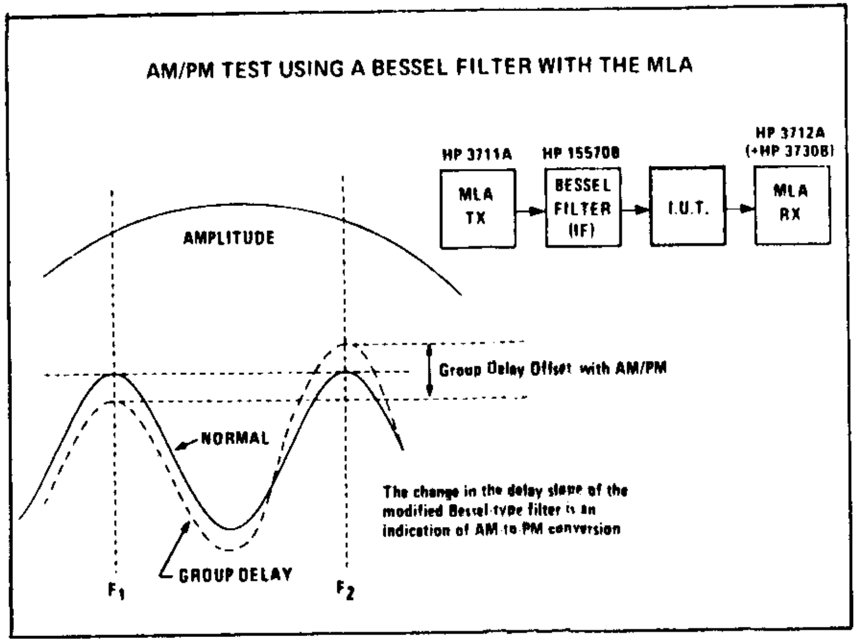
We will now review the measurement of sideband level and AM/PM conversion using the test network.

SPECTRUM RE-GROWTH FOR A 16QAM RADIO DUE TO 3dB TWT OVERDRIVE (MEASURED BEFORE MAIN TRANSMIT FILTER)

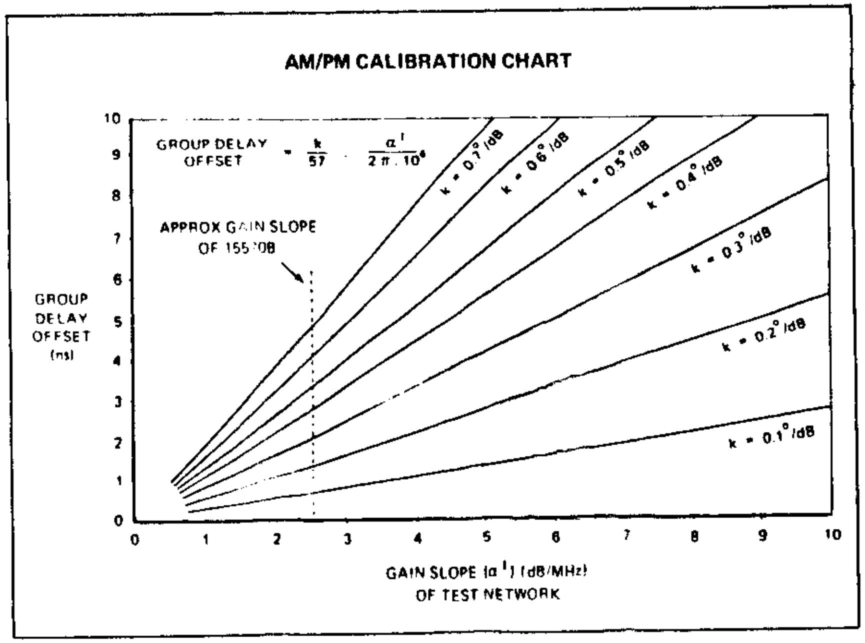


The measurement of unwanted sideband level is made simply by connecting a spectrum analyzer to the power amplifier output before the main transmit filter. Provided the input to the amplifier (e.g. TWT) is well filtered the sideband level is a measure of the non-linearity present. It is a fairly simple test but has the advantage of using the actual digital radio modulated signal with the correct dynamics.

If we wish to measure AM/PM we need to make a phase measurement. One method for doing this is described in the next two slides.



An alternative method of measuring individual components is the use of a Bessel filter (HP 15570B) in conjunction with the MLA. The filter is placed before the item under test and the non-flat amplitude response of the filter generates AM on the swept MLA test signal. The AM exercises the non-linearities of the item under test which generates incidental PM. This is interpreted by the MLA as a group-delay slope. (When measuring through a microwave device, the HP 3730B can be used to down-convert the signal to I.F.) By knowing the amplitude slopes of the filter and the group delay offset we can calculate the AM/PM conversion as shown in the next slide.



This slide shows graphically the relationship between gain slope and group-delay offset as a function of AM/PM conversion (k). One suitable test network is the HP 15570B shown on the graph. This test method is very sensitive and relatively free of calibration errors provided the MLA receiver has intrinsically low AM/PM conversion (checked using the HP 15570B).

For more information see:

"Technique Simplifies AM-to-PM Measurement" by I. Matthews, Microwaves and RF, August 1984, pp 108-112

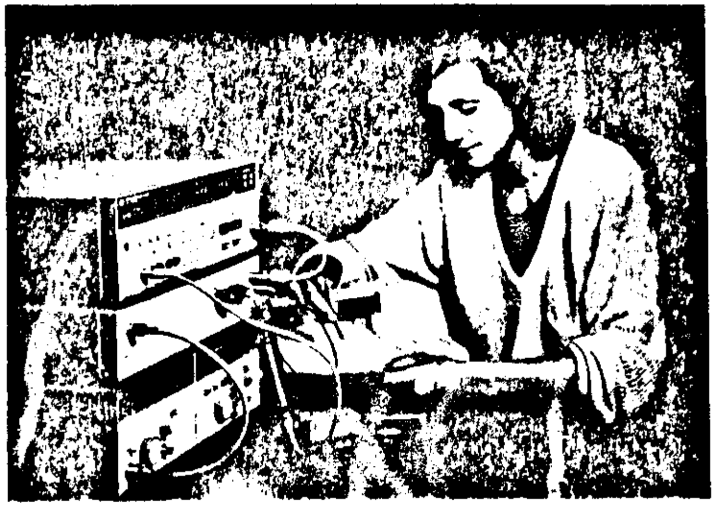
EFFECTS OF NOISE AND JITTER

- **RECEIVER NOISE FIGURE**
- **PHASE NOISE ON CARRIER SUPPLIES**
- **TIMING JITTER IN REGENERATOR**
- **JITTER SPECIFICATIONS AT NETWORK INTERFACE**

Next we will consider the effects of noise and timing jitter. Sources of noise in digital radio include receiver front-end noise figure, and phase noise in carrier supplies and recovered reference carrier. Since digital radio uses the phase relationship for transmitting information, phase noise will appear as random noise after demodulation and result in "eye-closure".

Timing jitter can appear in two areas: internal to the radio in the symbol-timing clock recovery in the regenerator, and externally at the interfaces of the radio to the digital network.

MICROWAVE NOISE FIGURE MEASUREMENT SYSTEM

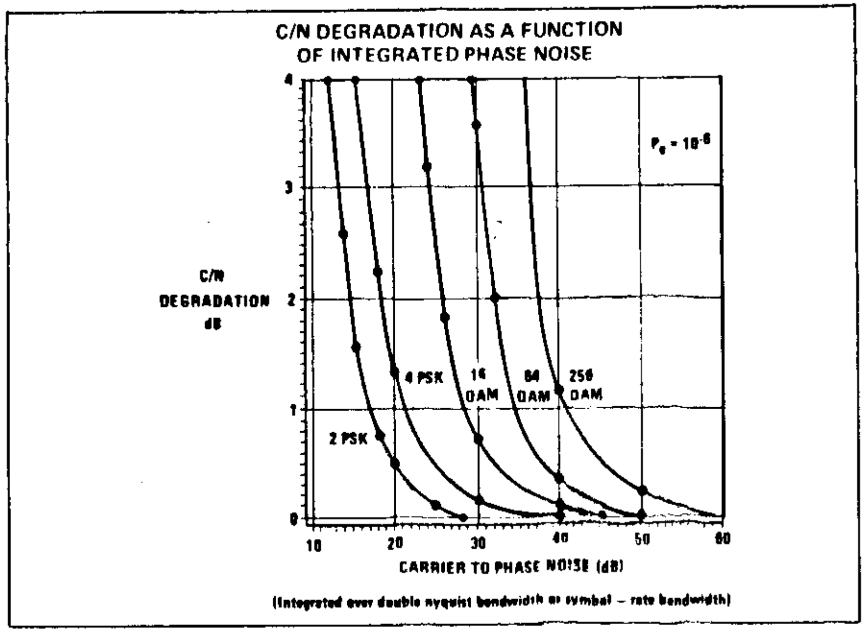


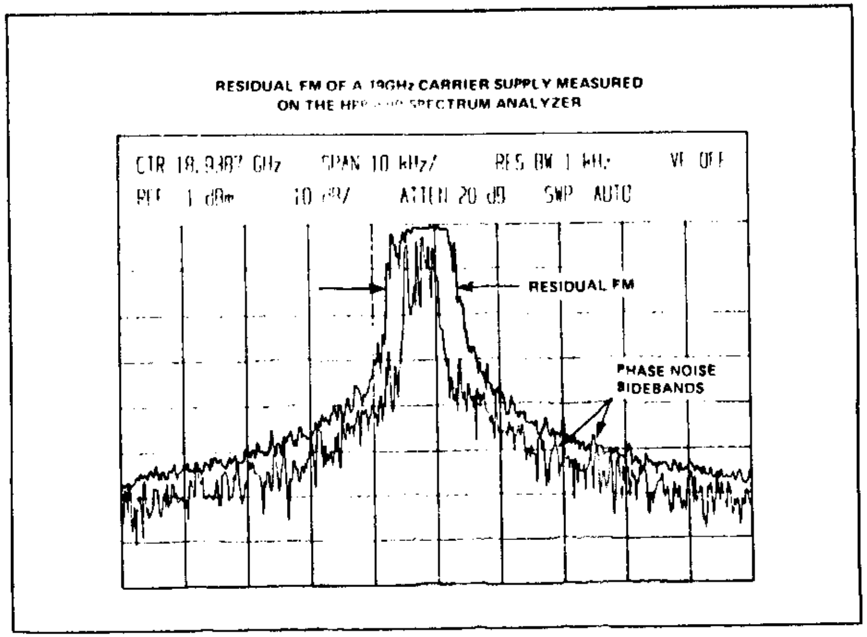
Thermal noise at the front end of the receiver is only significant in deep fades when there is a very low received signal level. The level of thermal noise is dependent on the noise figure of the front-end microwave preamplifier placed before the down converter. Noise figure is an important parameter in the overall system design as it determines receiver sensitivity and by implication the total system gain or fade margin. Usually it is measured at component level before the preamplfier is integrated into the receiver rack.

Noise figure measurements are easy to make using the HP 8970T shown in this slide. It operates between 10 MHz and 18 GHz. The output of the device under test is connected to the Noise Figure Test Set, and a calibrated noise source (HP 346B) connected to the device input. Measurements of gain and noise figure are calculated automatically.

For more information see:

Application Note 57-1 "Fundamentals of RF and Microwave Noise Figure Measurements" Hewlett-Packard, Pub No 5952-8255.





Turning now to the effects of phase noise in local oscillators, this slide shows the C/N degradation resulting from "carrier-to-phase-noise" in various modulation schemes*. The phase-noise value is the integrated value across the double-sided Nyquist bandwidth (or symbol-rate bandwidth) (see footnote).

As expected the more complex modulation schemes are much more sensitive to phase noise effects. Phase noise can be measured using a spectrum analyzer or a dedicated phase noise testing set-up.

* *"Performance of Multi-level Modulation Scheme in the Presence of Phase Noise." by A.D. Kincar and K. Feher, ICC'85 Proceedings Vol 1, pages 511-514. Published by I.E.E.E.*

<u>Comment</u>:

Phase noise close to the carrier is not usually important as it falls inside the bandwidth of the receiver carrier recovery phase lock loop in the demodulator. This in effect "tracks-off" the phase noise of the incoming signal over a bandwidth of perhaps ± 20 kHz around the carrier.

For simple modulation schemes such as 2-PSK and 4-PSK adequate phase-noise measurements can usually be made with a spectrum analyzer (particularly if it has a synthesized local oscillator). This slide shows the measurement of a 19GHz carrier supply intended for a 4-PSK radio. (The "peak-hold" facility captures the maximum excursion of the carrier residual FM.) The effects of phase-noise can also be seen on the constellation measurement as described later.

On complex schemes such as 64-QAM and 256-QAM, more sensitive phase-noise measurements are required using dedicated phase-noise test-equipment.

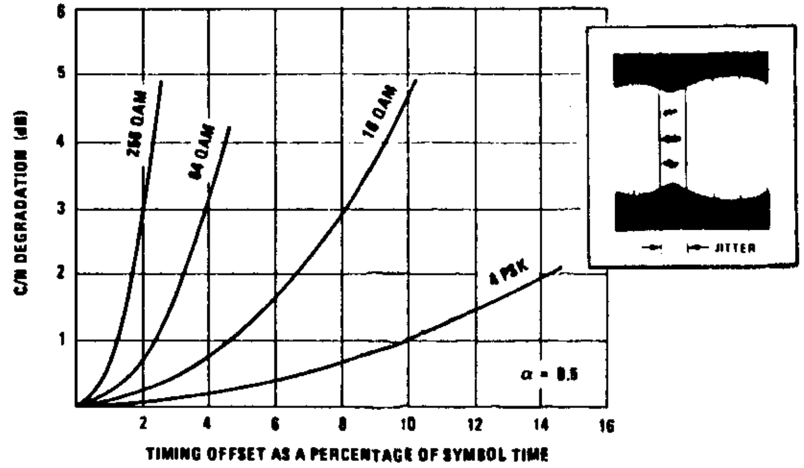
For more information see:

RF and Microwave Phase Noise Measurement Seminar (Pub No 5953-8479)

HP Product Note 11729C-2 "Phase Noise Characterization of Microwave Oscillators (Pub No 5953-6497).

"Choosing a Phase Noise Measurement Technique" by Terry Decker and Bob Temple (HP Part No 1000-1118).

C/N DEGRADATION DUE TO SAMPLING TIME ERROR

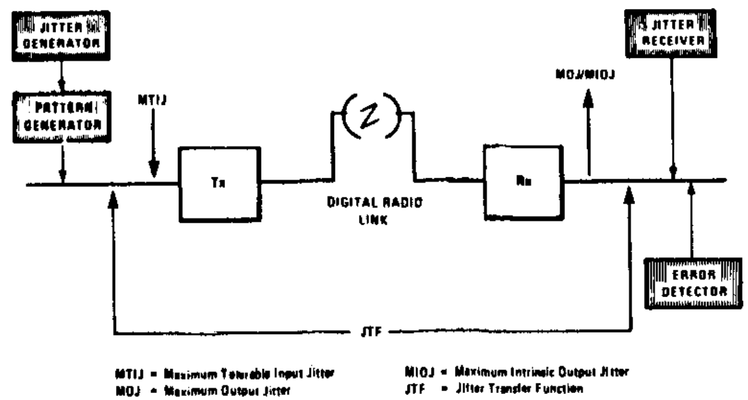


Jitter on the symbol timing clock which is used to sample the demodulator I and Q signals is also a source of errors in digital radio. If the demodulator signal is not sampled at exactly the centre of the eye-diagram (ie. at the position of maximum eye-opening), the effect is eye-closure at the sampling instant leading to an equivalent loss of C/N margin.

Jitter on the clock can be checked by measuring sideband levels on a spectrum analyzer or by observing zero-crossings on a digital processing oscilloscope such as the HP 54100. It is useful to check if the jitter is dependent on the data pattern and if it accumulates in successive regeneration.

As we saw in the theory section (page 15), the eye-diagrams for more complex modulation such as 16-QAM and 64-QAM are not only smaller in the vertical amplitude levels, but also narrower in the horizontal time axis. These systems are progressively more sensitive to timing errors and jitter as shown in slide. As you can see the timing error needs to be maintained within ± 1-2% of the symbol time (± 0.01/0.02 UI) or less.

JITTER MEASUREMENTS IN THE DIGITAL NETWORK



Perhaps the most common jitter measurements are made at the standard CCITT interface on the radio which connects with the digital network. A number of jitter specifications have been laid down by CCITT (Rec. G823 and G824*) and in Bell Technical References 43501 and 43806, for the various standard hierarchy rates. The idea is that if a piece of equipment meets the specifications at its input and output, then it can be connected freely within the digital netowrk without degrading jitter performance and causing errors.

There are three classes of measurements:

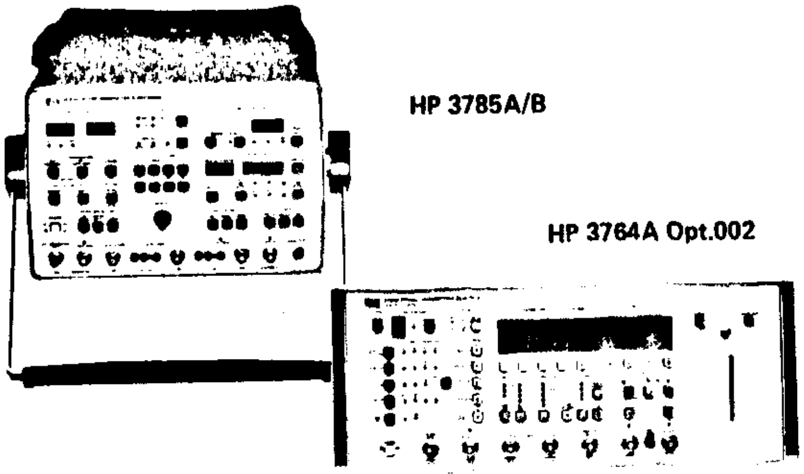
- **Maximum Tolerable Input Jitter** - which is tested by applying increasing jitter to an input data stream and determining the onset of bit errors.
- **Maximum Output Jitter (and Intrinsic Output Jitter)** - which is the level of output jitter with a jittered (or jitter-free) input signal.
- **Jitter Transfer Function** - which is a measure of how the jitter is attenuated by passing through the system, a necessary specification to prevent jitter accumulation in the network.

Jitter measurements are normally made in the factory where the equipment should be fully checked to the appropriate specification. If this has been done correctly, jitter measurements should not be necessary in the field except perhaps in very large networks where there is the danger of jitter accumulation.

* *In conjunction with ECSA T1X1.3 Committee*

For more information see: "HP 3785A/B Training Manual" Pub No 5953-6683, and *"Jitter Measurements in the Integrated Digital Network" by Ron McDowall, Pub No 5954-7939.*

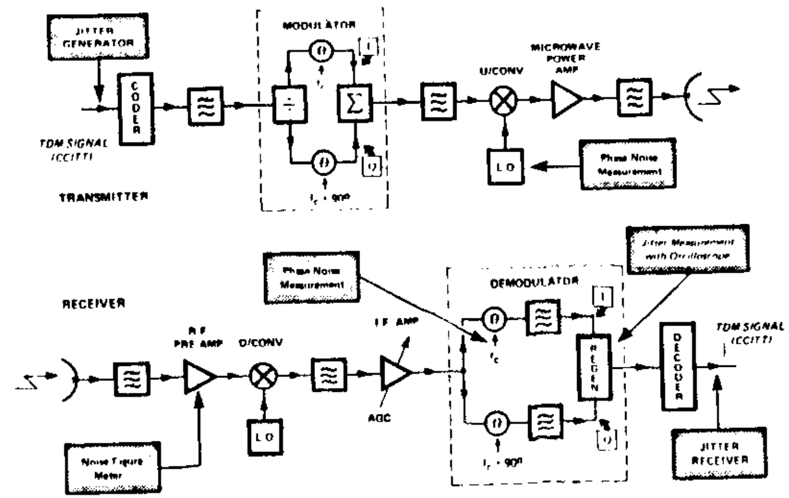
JITTER MEASUREMENTS IN THE DIGITAL NETWORK



HP 3785A/B

HP 3764A Opt.002

NOISE AND JITTER MEASUREMENTS



Jitter measurements at the network interfaces are standardized by CCITT and made using a dedicated jitter test set. The measurements are made according to the masks laid down in CCITT Specification G823 and G824.

The Hewlett-Packard products shown here are:

- HP 3785A for 2, 8 and 34 Mb/s CEPT Interfaces.
- HP 3785B for DS1, DS1C, DS2 and DS3 for North America.
- HP 3764A Opt 002 for 139 Mb/s CEPT Interface.

In this section we have discussed several different but related measurements. In summary here are the measurements referred to the digital radio block diagram. Jitter measurements at the CCITT network interfaces use a dedicated jitter generator and receiver, the other measurements are made with more general purpose equipment as the tests are not standardized.

MISALIGNED MODULATOR AND DEMODULATOR STATES

Now we will discuss the final major source of digital radio impairments, imperfections in the modulator and demodulator. Clearly if the phase states in the modulator are not correctly adjusted, the C/N margin of the system will be degraded since the state-space between the transmitted symbol and the decision threshold is reduced. We can classify the various geometrical misalignments into the categories shown in the next slide.

For more information on modulator/demodulator implementation see pages 23 to 25.

MODULATOR/DEMODULATOR IMPERFECTIONS

- **AMPLITUDE ERRORS**
- **QUADRATURE ANGLE ERRORS**
- **PHASE-LOCK ANGLE ERRORS**
- **INCORRECT DEMODULATION DECISION THRESHOLDS**

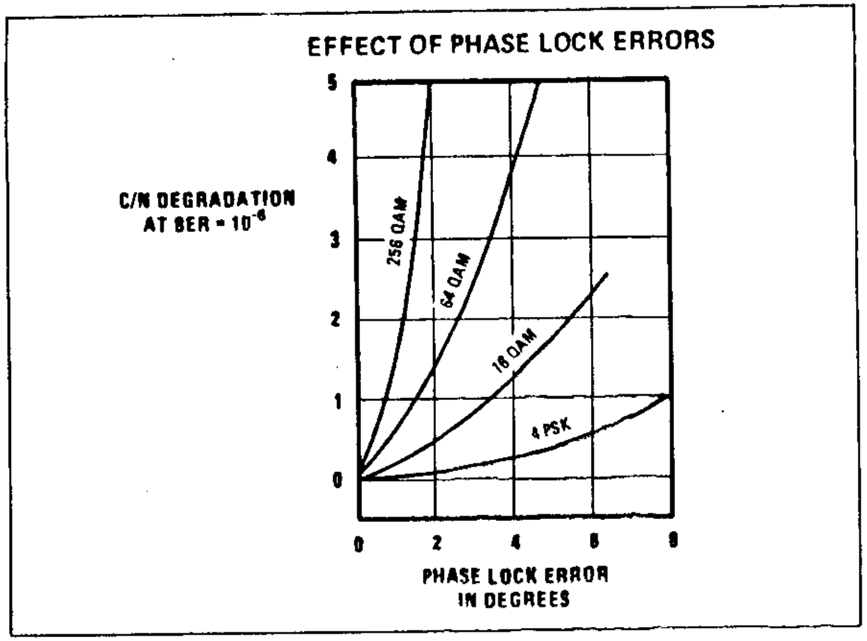
A variety of imperfections can exist in practical modulators and demodulators, the more complex the modulation scheme the more likely the problems.

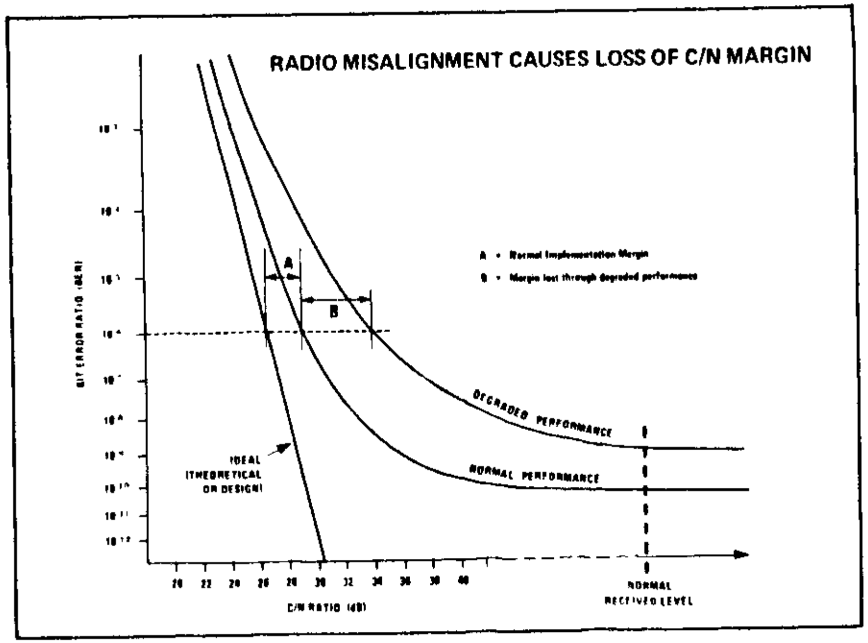
Amplitude Errors might consist of non-equal amplitude values in the four states of QPSK or unequal I and Q levels or non-linear modulation of I and Q carriers.

Quadrature Errors mean effects due to the I and Q carriers not being at exactly 90 degrees. The result is a modulation diagram which is trapezoidal or rhombic in shape.

Phase Lock Errors occur when the recovered carrier in the demodulator is not correctly phased with the incoming signal, resulting in a rotated modulation diagram.

Incorrect Demodulation Decision Thresholds are similar in effect to the amplitude errors in the modulator so that there is a reduced margin between the state value and the decision threshold.





As you might expect, these impairments are much more critical in higher modulation schemes. This slide shows the effect of phase-lock angle error in different modulation schemes and the result in terms of C/N degradation. You can see a 2 degree error in 4-PSK has little effect (this would be considered a reasonable specification for this modulation scheme) but in 64-QAM would result in an unacceptable loss of 1.5dB C/N margin. Similar graphs can be plotted for all the geometric distortions listed in the previous slide.

The study of all these impairments is a significant exercise in itself and beyond the scope of our present seminar. However, we will look at the practical measurement of some of these impairments later.

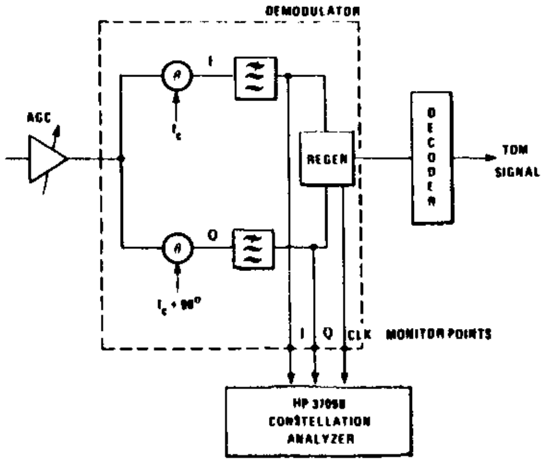
For more details on Modulator/Demodulator impairments see:

"Relating Constellation Parameters to Digital Radio Performance" by Ian Kennedy Compston (Pub No 5954-7943).

As we have seen already, all these impairments added together give rise to the normal implementation margin or separation between the theoretical and practical curve. If the radio is misaligned in any way, this implementation margin can increase substantially as this slide shows. Because of the "threshold-effect" in digital radio the only indication of degraded performance at normal received level will be an increase in the background BER. This may still be relatively low and remain undetected by built-in monitoring. However, when the radio is stressed by propagation effects or interference it will fail much earlier. How can we detect this condition before it causes failure?

One method would be to carry out a C/N versus BER test on a regular basis, but the problem here is that this test requires the radio to be taken out-of-service. Clearly we need a better way to check these impairments while the radio is still in-service. Fortunately such a test is available by looking at and analyzing the phase-state diagram or constellation of the radio.

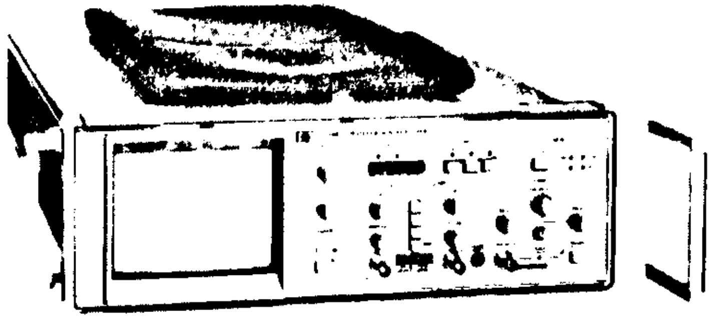
MAKING THE IN-SERVICE CONSTELLATION MEASUREMENT



To make the constellation measurement in-service, the radio must be equipped with suitable monitor points in the demodulator providing the post demodulation I stream, Q stream and symbol-timing clock. The measurements can be made using the HP 3709B Constellation Analyzer or the HP 8980A Vector Analyzer. Many radio impairments can be identified, including the modulator/demodulator geometric distortions we have just described.

If no monitor points are available for connection to a terminated 50/75 Ohm cable, then high-impedance probes can be used. Suitable accessories are available for HP 3709B and HP 8980A.

HP 3709B CONSTELLATION ANALYZER



IN-SERVICE ANALYSIS OF DIGITAL RADIO PERFORMANCE

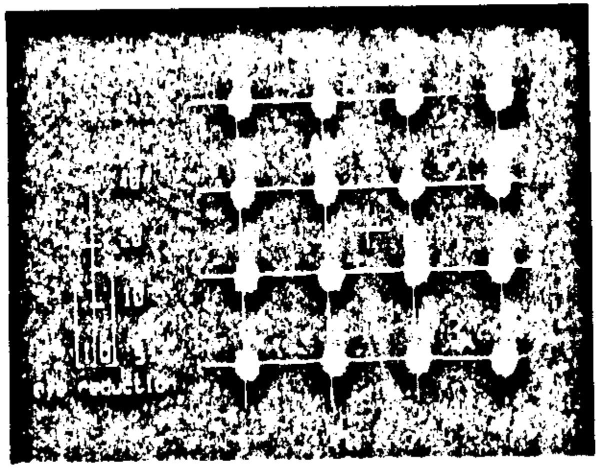
The HP 3709B Constellation Analyzer is a special purpose high-performance dual-channel sampling oscilloscope designed for digital radio measurements (for 4 PSK, 9 QPR, 16QAM, 25QPR, 49QPR, 64QAM, 81QPR and 256 QAM). It can display the radio constellation or either I or Q eye-diagrams at the push of a button. The analog circuitry is designed to handle the random digital radio signal with the minimum of distortion (ISI) and very low drift, and the autoranging timebase automatically displays 2 periods of eye-pattern for symbol rates between 1 and 80 Mbaud.

In addition to visual interpretation, the HP 3709B can make estimates of the average rms closure for all displayed phase states, lock and quadrature angle, and non-linear distortion either on a repetitive basis or in a single shot mode for higher accuracy and repeatability.

Measurements and constellation display can be dumped directly to the HP 2225A ThinkJet Printer via HP-IB.

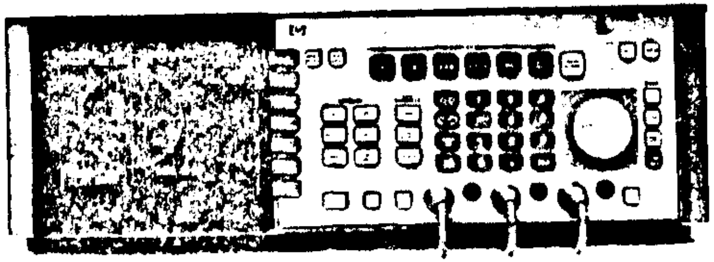
For more information on HP 3709B measurements see Product Note 3709-1 "Measurement Principles" (Publication No 5954-9546).

HP 3709B BAR-CHART DISPLAY



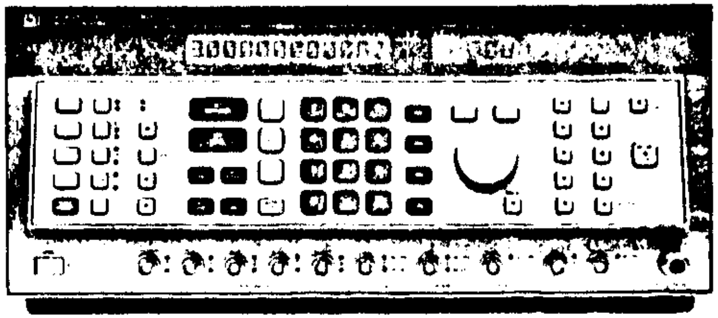
The HP 3709B can also display I and Q "eye reduction" as a bar-chart. This is a composite measurement including all the effects of individual impairments, similar to the measurement of overall C/N margin. The bar chart has a fast update rate for making adjustments.

HP 8980A VECTOR ANALYZER



The HP 8980A Vector Analyzer has similar measurements and performance to the HP 3709B Constellation Display, however it has extra facilities more appropriate to R&D work in Digital Radio. Whereas the HP 3709B displays only constellations and conventional eye-diagrams, the HP 8980A can be used as a general purpose oscilloscope and can display the transitions in the I/Q vector diagram as well as constellations with phase and magnitude markers. It is also fully programmable over the HP-IB, making it a good choice for automatic test systems. Measurement displays can be dumped directly to the HP 2225A ThinkJet Printer as with the HP 3709B. Constellation measurements can be averaged over a user-defined number of measurement points allowing a trade-off between speed and repeatability."

HP 8780A VECTOR SIGNAL GENERATOR



The HP 8780A Vector Signal Generator can be used as a reference I/Q modulator in the synthesized frequency range 10 MHz to 3 GHz. It can generate standard modulation schemes: BPSK, QPSK, 8PSK and 16QAM, and with option 064, 64QAM. The HP 8780A is valuable in the development and testing of radio and satellite receivers, and can generate a complex test signal for analysis of non-linear devices such as transmit amplifiers. The Vector Generator can be modulated either through multiple digital inputs or via analog I and Q vector inputs (eg with filtered multi-level signals).

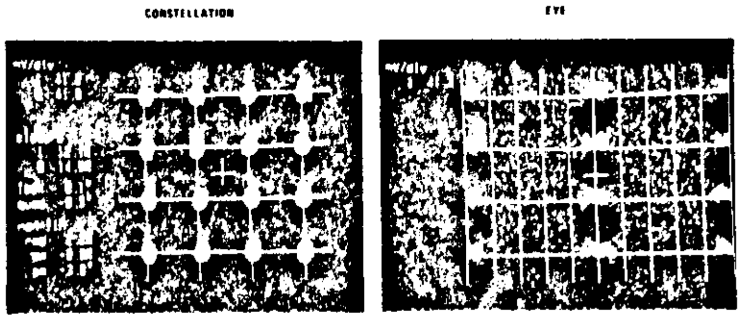
The following slides show the constellation and corresponding eye-diagram for a 16 QAM radio. They demonstrate the diagnostic power of the constellation measurement, the eye diagram showing little more than the presence of eye closure.

For more information see:

Product Note 8780-1 (Hewlett-Packard Pub No 5954-6368)

Application Note 343-1 "Vector Modulation Measurements" (Hewlett-Packard Pub No 5954-6365).

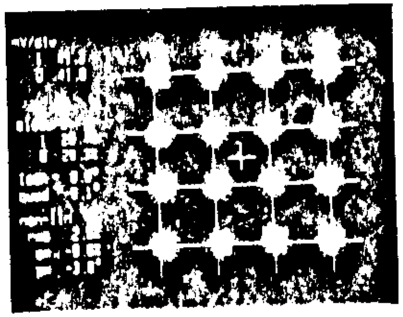
16 QAM RADIO NORMAL OPERATION



This is the picture you would like to see!

The 16 states are correctly placed on the rectangular grid and the cluster size is small indicating low thermal noise, interference and ISI. The constellation measurements can be seen on the left-hand side of the display and indicate an eye closure of 12 - 14% and negligible angular displacements.

16 QAM RADIO DEGRADED C/N

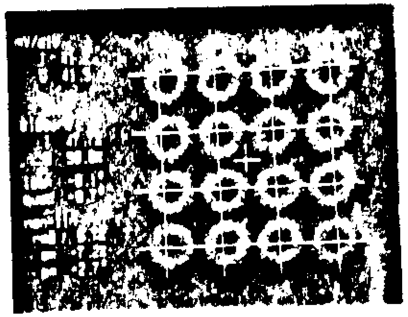


C/N = 15dB

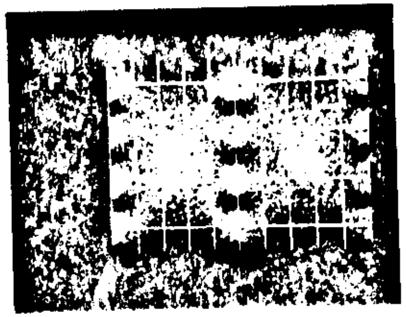
C/N = 15dB

This slide shows the effect of a poor C/N ratio (15 dB) in the radio, and as expected the constellation states are enlarged, therefore creating eye closure and a higher BER. In this slide the r.m.s. eye closure is around 27%. Note that in this and the last slide, the same information is available from the eye-diagram since there are no geometric distortions present in the constellation.

16 QAM RADIO WITH SINUSOIDAL INTERFERENCE



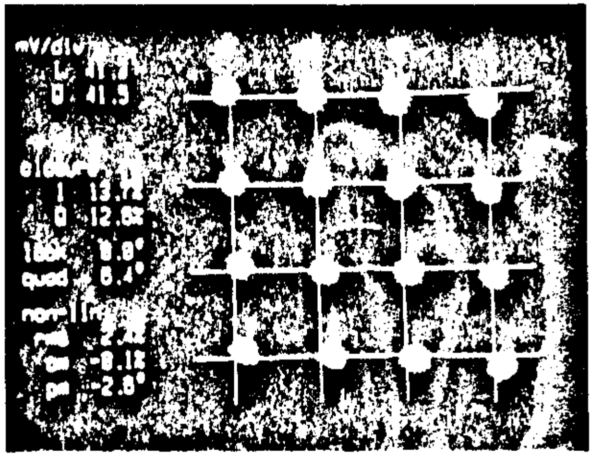
C/N = 15dB



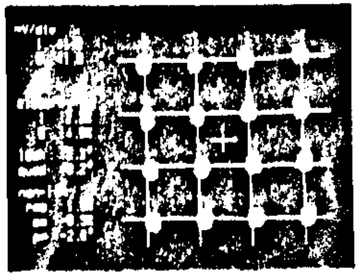
C/N = 15dB

In this case the radio is being affected by a sinusoidal interfering tone as might be the case when interference is being received from the strong carrier component of an FM radio. The sinusoidal interferer generates a "doughnut" shape due to the rotating vector of the interference signal being frequency offset from the radio carrier.

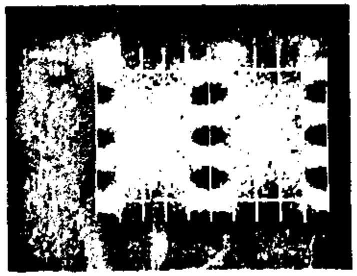
16 QAM RADIO: QUAD WITHOUT LOCK



16 QAM RADIO WITH CARRIER LOCK ERROR



2.3° Carrier Lock Error



2.3° Carrier Lock Error

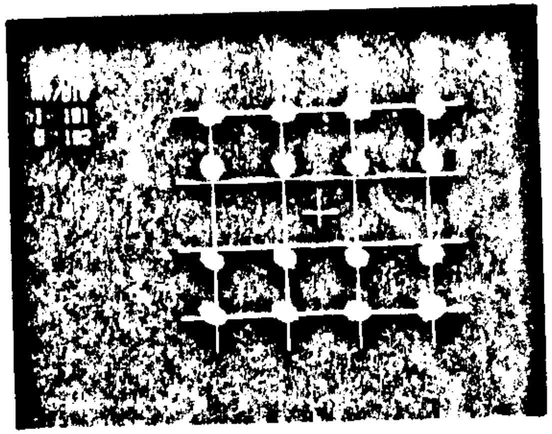
The next two slides show some modulator/demodulator impairments. This is the kind of picture you would see if the I and Q carriers were not exactly 90 degrees to each other. Such a fault would produce eye-closure on one eye-diagram but not on the other, provided no other fault were present. Often, however, this situation may be combined with a phase lock error shown in the next slide.

In this case the internal calculations indicate a quad angle error of 6.4 degrees (the negative angle indicates clockwise rotation).

The rotation in this picture is due to a phase-lock error in the demodulator carrier-recovery loop. Notice the eye-diagram only shows eye-closure as might occur in the presence of noise.

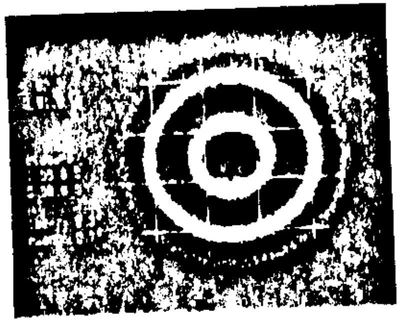
The constellation measurement indicates a rotation or lock error of 2.3 degrees, a serious impairment for 16 QAM as we saw earlier (page 71).

16 QAM RADIO: AMPLITUDE IMBALANCE

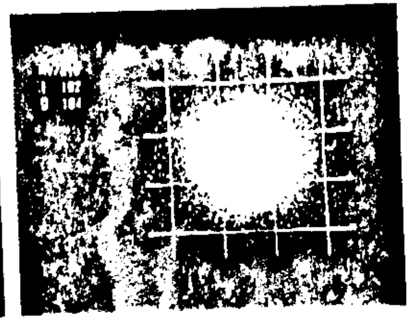


This constellation shows an example of another impairment which can occur in the modulator and demodulator. In this case the amplitude levels on the quadrature carrier are not correctly set. This might be due to a non-linear modulator or inaccurate D to A converter.

16 QAM RADIO: CARRIER RECOVERY LOOP OUT-OF-LOCK PLUS MEASUREMENTS



16 QAM RADIO: NO SIGNAL



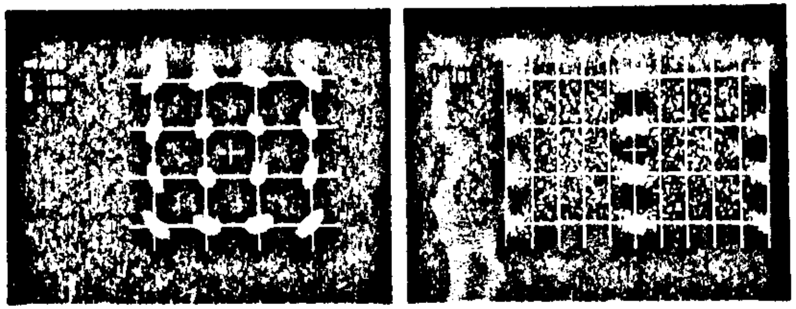
These two constellation pictures illustrate complete failure.

(1) When the radio demodulator is out-of-lock, the lack of phase coherence creates circles on the display.

(2) The lack of any input signal produces one large cluster of random points.

(Note that the eye diagram would show both of these as complete eye closure.)

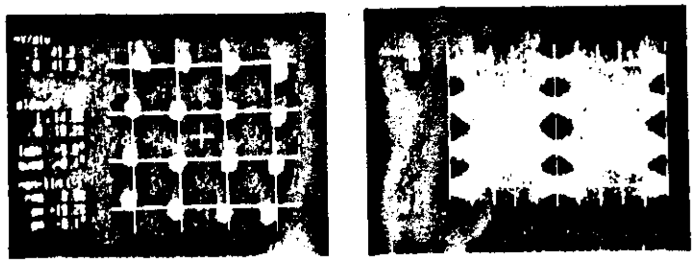
16 QAM RADIO: PHASE JITTER



Phase jitter on a carrier supply (local oscillator) or recovered carrier will cause a spreading of the constellation points in a circular fashion as shown in this slide. The eye diagram does not reveal this fault.

CONSTELLATION AND EYE DIAGRAM FOR 3dB OVERDRIVE OF TWT

CONSTELLATION EYE



Looking now at transmitter power amplifier non-linearities, here we can see the effect of a 3dB rise in output power. Notice how the constellation is twisted and rotated (AM/PM) and compressed at the corners where there is maximum power (AM/AM).

The constellation is a very powerful method of analyzing non-linearity, and optimizing performance either using the internal "non-linearity" measurement or with increased precision through an external computer.

Comment:

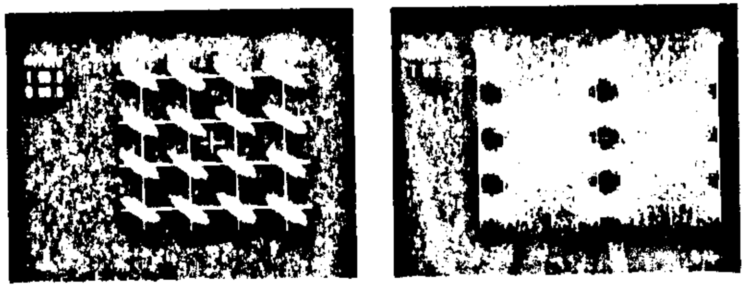
Constellation measurements are probably the best way of analyzing non-linearity effects for two reasons. Firstly, the test signal is the true digital radio modulation so the non-linearity is exercised exactly as it would be in normal operation - sinusoidal test signals, for example, do not have the same dynamic behaviour. Secondly, as this slide shows, both amplitude distortion (AM/AM) and phase distortion (AM/PM) effects can be measured. These effects are quantified in the HP 3709B non-linearity measurement.

For more information see:

"Troubleshooting Digital Radios using Constellation Analysis" by Tom Crawford and Geoff Waters (Publication No 5954-2037)

"Technique simplifies AM-to-PM Measurements" by I. Matthews, Microwave and RF, August 1984 pp 108-112

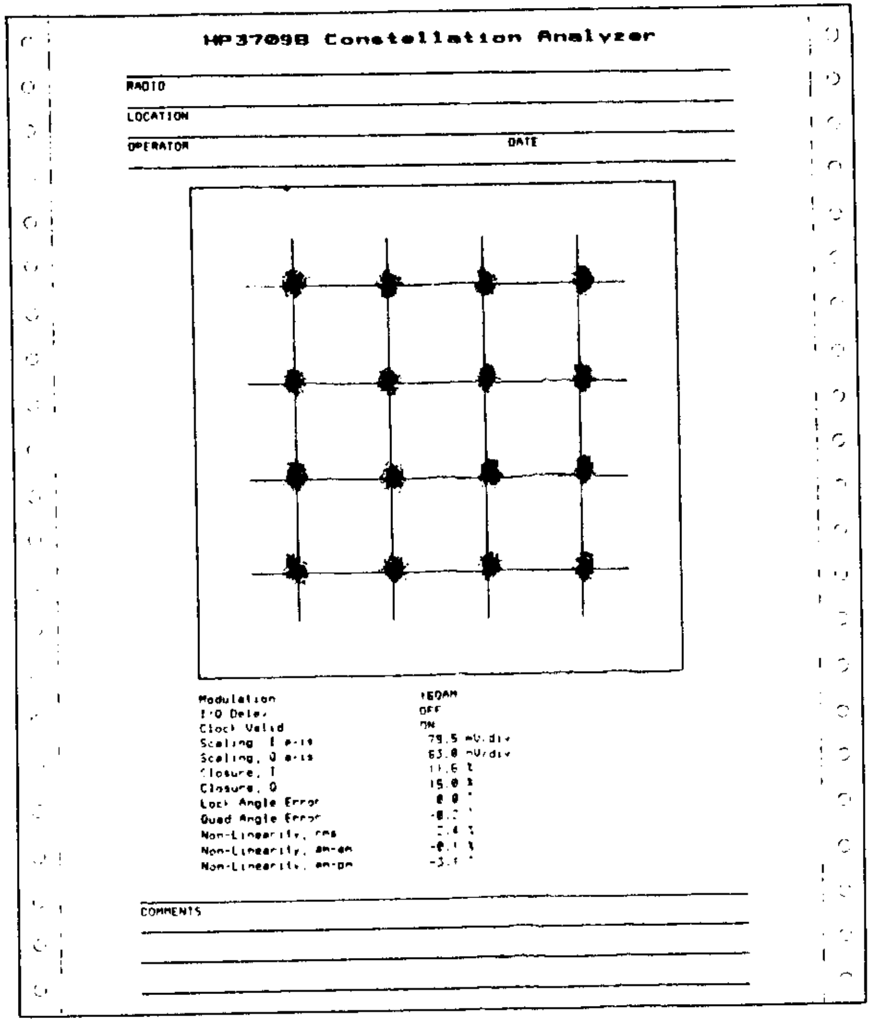
CONSTELLATION AND EYE DIAGRAM FOR AMPLITUDE SLOPE ACROSS THE CHANNEL



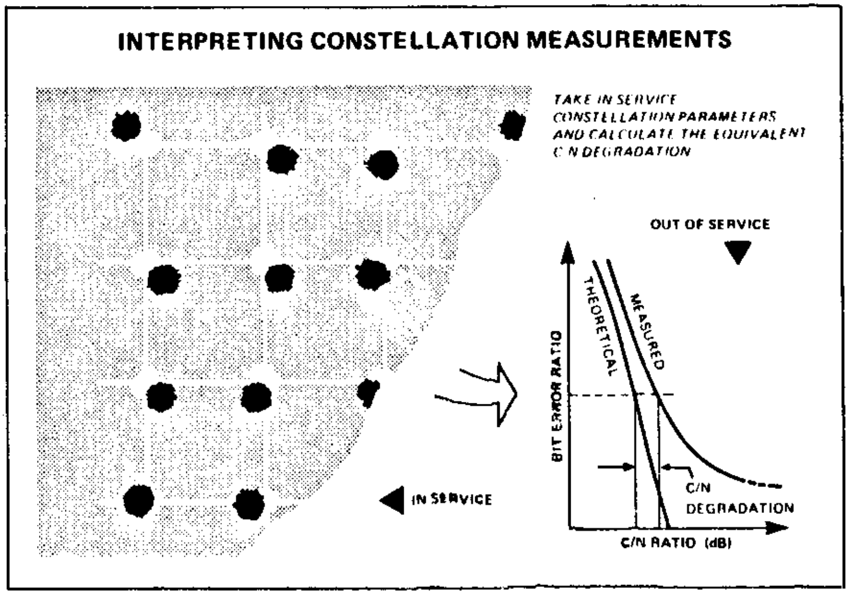
Lastly, this slide shows the effect of an amplitude slope as might occur during a multipath fade. The 45 degree oval shape of the clusters shows that there is crosstalk between the I and Q signal which is the dominant effect of assymetrical distortion such as amplitude slope. You may notice that the point of maximum eye-opening on the eye-diagram is shifted from the nominal position at the centre of the screen (i.e. the sampling position for normal operation). In order to obtain the constellation pictures the I and Q timing controls need to be slightly adjusted to explore the correlation between I and Q channels.

The characteristic eliptical and diamond shaped clusters can also be seen with time-varying multipath activity. The variable nature of this phenomina would be distinguishable from static transmission impairments in filters, antennas etc.

This series of slides shows the diagnostic power of the constellation measurement for in-service tests.



This shows a typical printout on the HP 2225A Thinkjet printer connected to the HP 3709B. This provides a simple way of documenting radio performance and generating a catalog of impairments to help in troubleshooting a particular radio system. A similar printout can be obtained from the HP 8980A Vector Analyzer.



In the last few slides we have used the constellation analyzer for visual interpretation of impairments, however, the constellation can provide valuable analytical information if it is processed numerically.

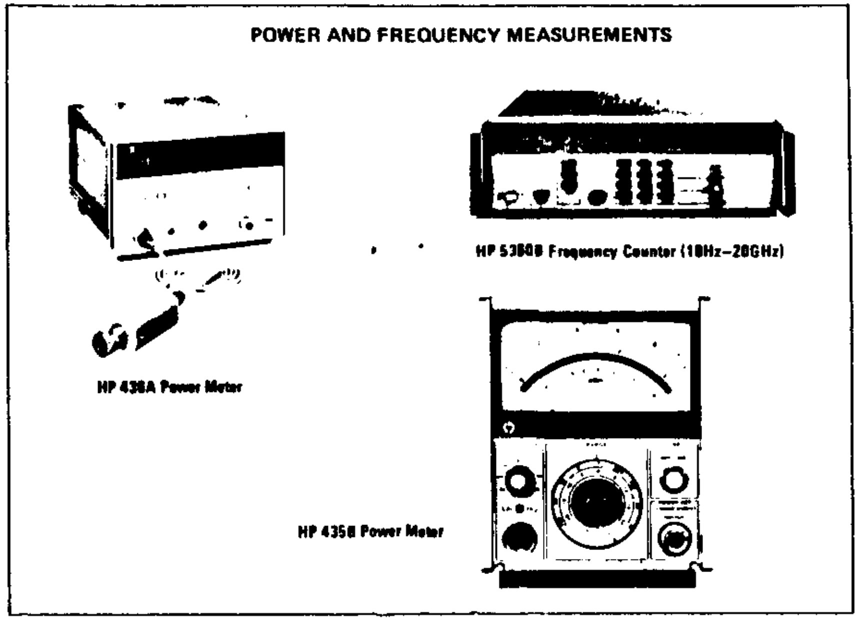
By taking the statistical constellation parameters for position and cluster shape and size it is possible to calculate the equivalent C/N degradation and residual BER. This requires the HP 3709B or HP 8980A to be connected to a computer to analyze results. The benefit is that an **in-service measurement** technique can provide test results that would normally require the radio to be taken out-of-service.

In summary, the constellation is perhaps the most useful of all the digital radio measurements, all the impairments are there just as the radio receiver sees them!

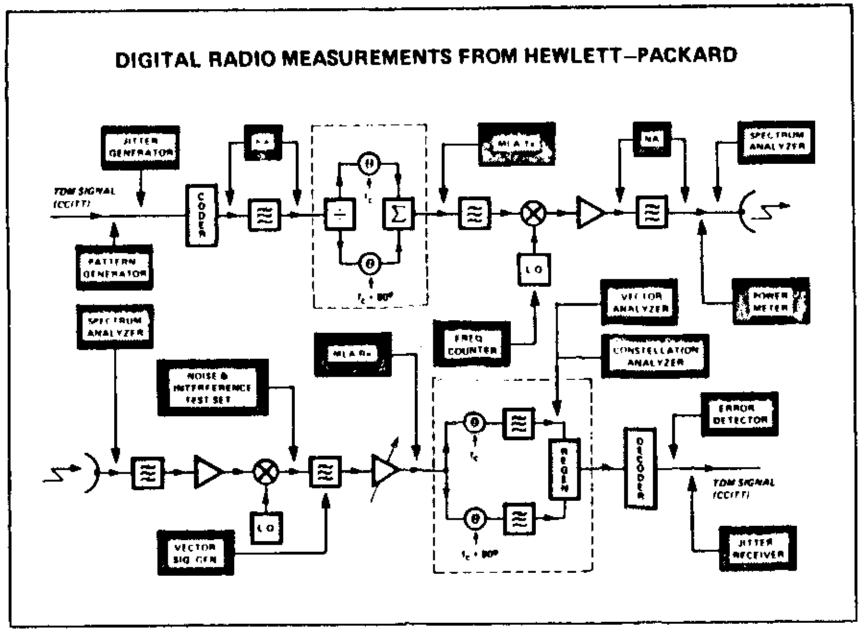
For more details on this technique, including sample programs see:

Product Note 3709-2 "Constellation Parameters" (Publication No 5954-9550).

Demonstrate the Constellation Analyzer and Vector Products



Finally, as in all microwave systems we need to measure power and frequency accurately. Hewlett-Packard produces a wide range of accurate power meters and frequency counters, some of which are illustrated here. See the appropriate data sheets for more details.



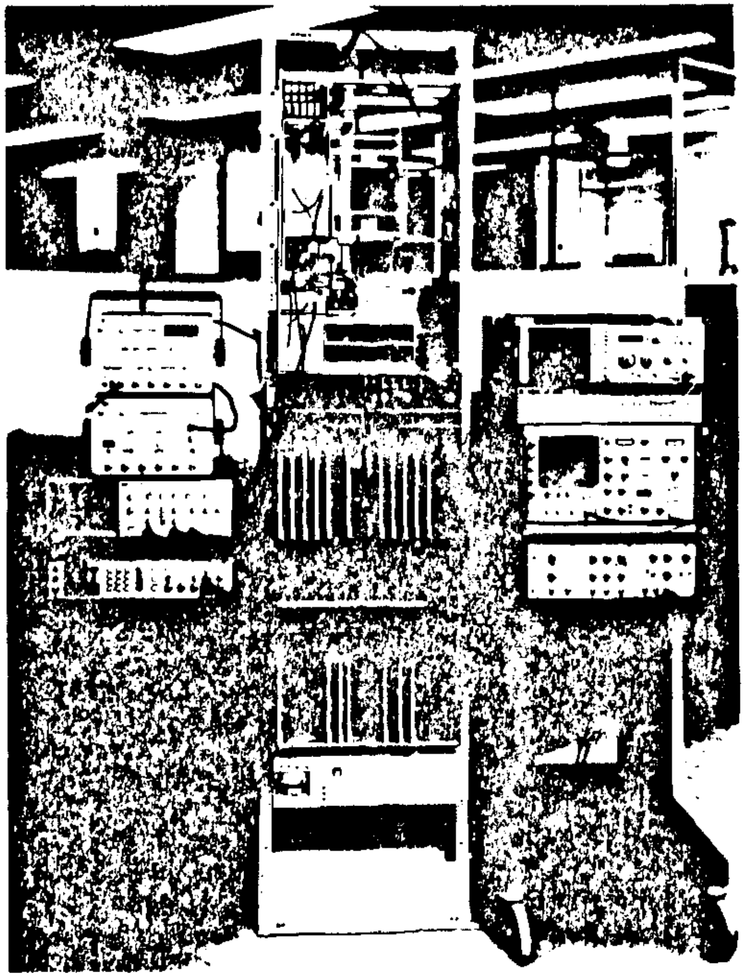
This slide shows the connection points for the test equipment we have just described. Exactly when and where these measurements will be made depends on the design philosophy of the manufacturer and the maintenance philosophy of the operating company. The next slide shows a typical application matrix rating the importance of various measurements in the key areas.

DIGITAL RADIO MEASUREMENT APPLICATION AREAS

MEASUREMENT TYPE	MAINTENANCE		REPAIRS/INSTALLATION COMMISSIONING	MANUFACTURE/DESIGN
	IN SERVICE	OUT OF SERVICE		
BER IN SERVICE MONITORING BUILT IN FOR PARITY AND CODE ERRORS	•••		•	
BER vs C/N AND C/I		••	•••	•••
MLA AND NETWORK ANALYZER TRANSMISSION MEASUREMENTS GROUP DELAY/IF RESPONSE/ RETURN LOSS		••	•••	•••
SPECTRUM ANALYSIS (RF/IF)	•	•••	•••	•••
POWER AND FREQUENCY MEASUREMENT	•	•••	•••	•••
CONSTELLATION ANALYSIS	•••	•••	•••	•••
MULTIPATH SIMULATION		•	•	•••
JITTER TESTS		•	••	•••
PHASE NOISE			•	•••
NOISE FIGURE			•	•••

••• Important Requirement
•• Occasional Requirement
• Rarely Measured

SOME OF THE RANGE OF HEWLETT-PACKARD TEST EQUIPMENT BEING USED TO CHECK A 16 QAM RADIO



That brings us to the end of the Digital Radio Seminar. This final slide shows some of the test equipment we have discussed being used to check a 16-QAM Microwave Radio.

Solution to Problem on Page 22

The first step is to calculate the symbol-rate of the radio, so that the theoretical minimum Nyquist bandwidth can be found.

16QAM has 16 states or symbols
therefore 4 bits/symbol (see page 13)

with a bit rate of 140Mb/s, the symbol rate = 140/4 =

35 Msymbols/s or 35 Mbaud.

Thus the minimum double-sided bandwidth (see page 20)

= 35 MHz.

If the practical raised-cosine filter has a factor = 0.5 then the system will have theoretically 50% extra bandwidth (see page 19)

= 52.5 MHz.

A practically designed radio would probably have a narrower bandwidth than this (e.g. 45 MHz) as the designer might decide to roll the filter off faster towards the cut-offs. This would not generate much ISI compared with a departure from the correct in-band response.

APPENDIX A - INFORMATION THEORY AND MODULATION SCHEMES

The question is sometimes asked, is there any limit to the amount of information that can be passed through a communication channel as more and more complex modulation schemes are used? The answer is in theory, no, provided we had a noiseless system, as we could encode the information on an infinite number of states without any uncertainty of the received state. Practical systems however are limited by noise, and it is possible to relate the maximum error-free information rate to the system signal to noise ratio through the Shannon-Hartley Law, which was derived from basic physical principles. The information rate in bits/second is expressed by the equation:

$$Cc = B\log_2 \left(1 + \frac{S}{N}\right)$$

where B = channel bandwidth in hertz

Expressed in bits/second/hertz,

$$\text{Bandwidth Efficiency} = \log_2 \left(1 + \frac{S}{N}\right)$$

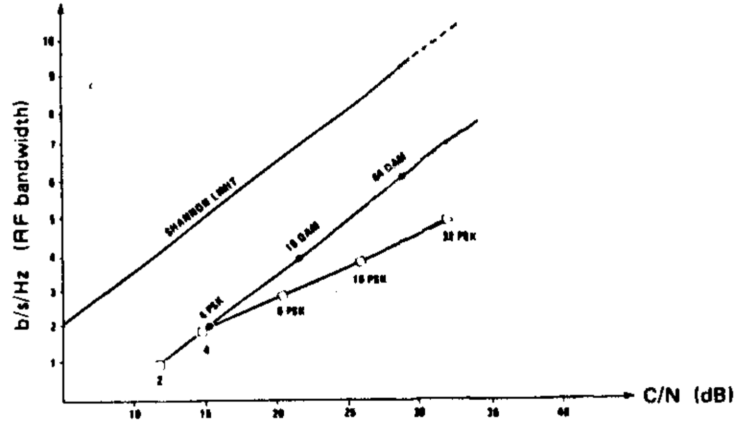


Fig. A1 Spectrum Efficiency in bits/second/Hz for different modulation schemes as a function of C/N ratio compared with the theoretical Shannon Limit.

This is plotted on Fig.A1 (together with some practical modulation schemes) and shows why the QAM schemes are preferred to PSK since they make better use of the available phase-state space. Other coding schemes and techniques such as correlation can be used to get closer to the Shannon limit.

For more information see:

"Principles of Communication - Systems, Modulation and Noise" by R.E. Ziemer and W.H. Tranter (Houghton Mifflin, 1976)

"Information and Communication Theory" by A.M. Rosie (Van Nostrand, 1973)

APPENDIX B - MICROWAVE ATTENUATION DUE TO RAINFALL

Fig. B1 shows the relationship between path attenuation per km and microwave frequency for varying rainfall rates in mm/hr.

Light rain in the 4-6 mm/hr range may last for several hours whereas heavy rain (50 mm/hr) will last for shorter duration.

In tropical or mountainous regions very heavy rain (150 mm/hr or more) can occur for short periods.

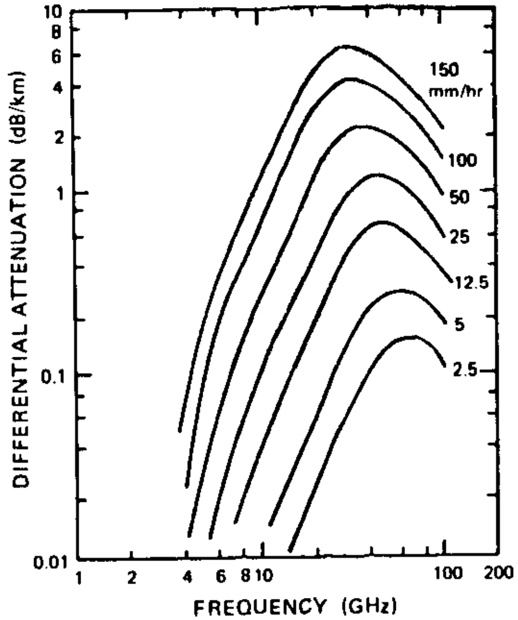


Fig. B1 Path Attenuation per km/Frequency for varying rainfall rates.

BIBLIOGRAPHY

From the first section of this seminar: Digital Radio - Theory and Implementation.

Feher, K.; **"Digital Communications - Microwave Applications"** (Prentice Hall Inc., 1981)

Lucky, R.W.; **"Functional Analysis Relating Delay Variation and Intersymbol Interference in Data Transmission"** (BSTJ, Sept. 1963)

Nyquist, H.; **"Certain Topics in Telegraph Transmission Theory"** (AIEE Trans., April 1928)

Rosie, A.M.; **"Information and Communication Theory"** (Van Nostrand, 1973)

Shannon, C.E.; **"Communication in the Presence of Noise"** (Proc. IRE vol. 37, Jan. 1949, pp 10 - 21)

Sunde, E.D.; **"Theoretical Fundamentals of Pulse Transmission"** (BSTJ, May 1954)

Ziemer, R.E. and Tranter, W.H.; **"Principles of Communication - Systems, Modulation and Noise"** (Houghton Mifflin, 1976.)

Weber, W.J.; **"Differential Encoding of Multiple Amplitude and Phase Shift Keying Systems"**, IEEE Transactions on Communications Vol COM-26 No 3, March 1978, pp 385-391.

Feher, K. and Engineers of Hewlett-Packard; **"Telecommunications Measurements, Analysis, and Instrumentation"** Prentice-Hall Inc, 1987.

From the second section of this seminar: Digital Radio - Practical Problems and Measurement Solutions.

Borgne, Michel: **"Comparison of High-Level Modulation Schemes for High-Capacity Digital Radio Systems"** (IEEE Transactions on Communication, Vol. Comm-33 No. 5 May 1985, pp 442-449.)

Hart, G.; **"Performance of Digital Radio-Relay Systems"** (British Telecommunications Engineering, pp 201-209 Vol 3, Oct. 1984.)

Matthews, I.: **"Technique Simplifies AM-to-PM Measurements"** (Microwaves and RF, August 1984, pp108-112.)

Kincar, A.D.; Feher. K.: **"Performance of Multi-level Modulation Scheme in the Presence of Phase Noise"** (ICC'85 Proceedings Vol 1, pages 511-514. Published by IEEE)

Damosso, Dionis, Zenobio and Fabbin; **"Experimental Results on Digital Radio-Relay Quality compared with the new CCIR Objectives"**, European Radio Relay Conference, Munich, November 1986.

Biester and Glauner; **"Distortion Analysis of Digital Radio-Relay Systems using 4 - 256QAM"**, European Radio Relay Conference, Munich, November 1986.

Noguchi, Daido, and Nossek; **"Modulation Techniques for Microwave Digital Radio"**, IEEE Communications Magazine, Oct 1986, Vol 24, No 10, pp 21-30.

Rummler, Coutts and Liniger; **"Multipath Fading Channel Models for Microwave Digital Radio"**, IEEE Communications Magazine, November 1986, Vol 24, No 11, pp 30-41.

Wu and Achariyapaopan; **"Effects of Waveguide Echoes on Digital Radio Performance"**, Globecom 85, paper 47.5.

HEWLETT-PACKARD PUBLICATIONS :

RF + Microwave Phase Noise Measurement Seminar (Pub No 5953-8479).

HP Product Note 11729C-2 **"Phase Noise Characterization of Microwave Oscillators"** (Pub No 5953-6497).

McDowall, Ron; **"Error Performance Requirements for Future ISDNs"** (Hewlett-Packard Pub No 5954-7938).

Compston, Ian Kennedy;**"Methods for Estimating Residual BER in Digital Radio"** (Hewlett Packard Pub No 5954-7942).

Kaylor, Jim; Walker, Hugh: **"Practical Applications of Spectrum Analyzers for Digital Radio Measurements"** (Hewlett-Packard Pub No 5954-7944).

Walker, Hugh: **"A Review of Digital Radio Principles and Measurements"** (Hewlett-Packard Pub No 5954-7941).

McDowall, Ron; "Jitter measurements in the Integrated Digital Network" (Hewlett-Packard Pub No 5954-7939).

Crawford, Tom; Waters, Geoff: "Troubleshooting Digital Radios using Constellation Analysis" (Hewlett-Packard Pub No 5954-2037).

Product Note 3709-1 "Measurement Principles" (Pub No 5954-9546).

Product Note 3709-2 "Constellation Parameters" (Pub No 5954-9550).

Product Note 3708-1 : "Noise and Interference Effects in Microwave Radio Systems" (Pub No 5953-5481).

Product Note 3708-2 : "Using the HP3708A on Microwave Radio Systems" (Pub No 5953-5489).

Product Note 3708-3 : "Determination of Residual Bit Error Ratio in Digital Microwave Systems" (Publication No 5953-5490).

Product Note 3708-4 "HP 3708A Demonstration Guide" (Pub No 5954-9551).

HP 3785A/B Training Manual (Pub No 5953-6683).

CEPT Standards - HP 3764A Training Manual (Pub No 5953-5444).

N. American Standards - HP 3781/82B Training Manual (Pub No 5952-3290)

Product Note 3730-1 : "High-accuracy RF Measurements using HP 3730B" (Pub No 5953-5468)

Waters, Geoff; Young, Ivan; "Additive White Noise and Interference Testing of Digital Radio Systems with Noise of Finite Crest Factor" (Pub No 5954-2007).

Product Note 8780-1 "Introductory operating guide for the HP 8780A Vector Signal Generator" (Pub No 5954-6368).

Application Note 343-1 "Vector Modulation Measurements" (Pub No 5954-6365).

Decker, Terry; Temple, Bob; "Choosing a Phase Noise Measurement Technique" (Part No 1000-1118).

DIGITAL RADIO MAINTENANCE