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COURSE ON BASIC TELECOMMUNICATIONS SCIENCE

9 January - 3 February 1989

Notes on Lectures 2 - 5

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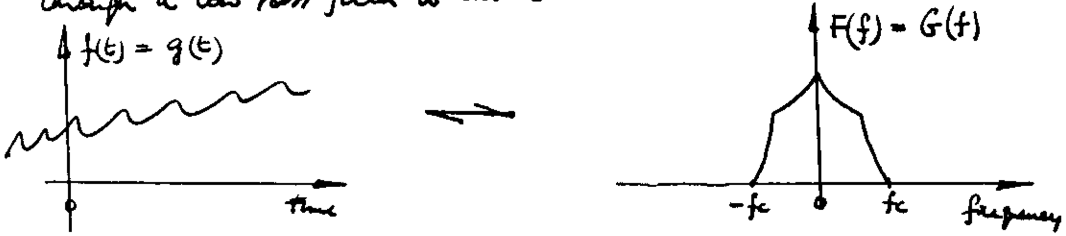
SYSTEMS I — TELECOMMUNICATION SYSTEMS — ISDN

All implementation of ISDN are based on the use of Pulse Code Modulation (P.C.M.) so we need to first introduce the ideas of pulse modulation.

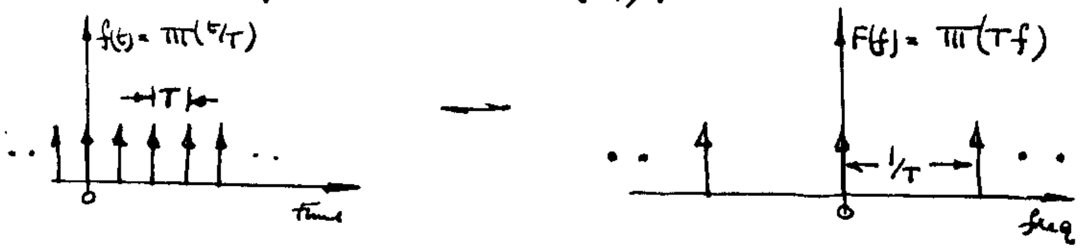
In fact if you think about it pulse communication has been around for a lot longer than analogue (voice)

All modern pulse modulation systems are based on the Nyquist sampling theorem.

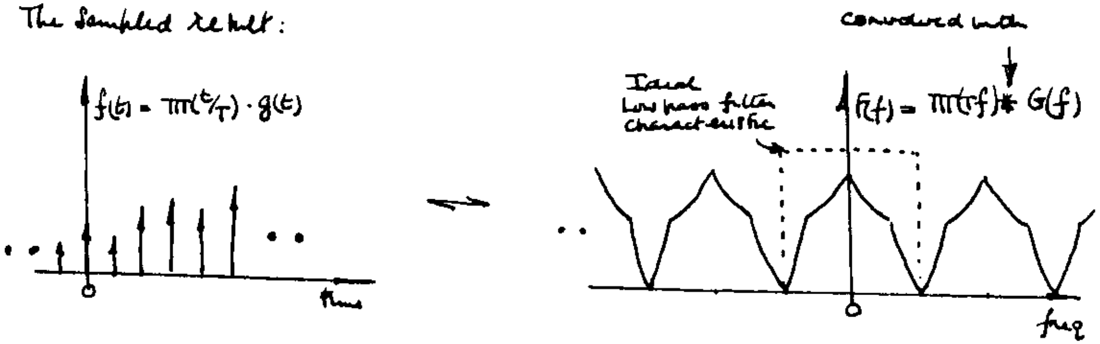
First consider a time signal $g(t)$ which we specify to be bandlimited — if necessary we would pass the signal through a low pass filter to ensure that this is so.



We can then sample this with our $Ш(t/T)$ function



The sampled result:



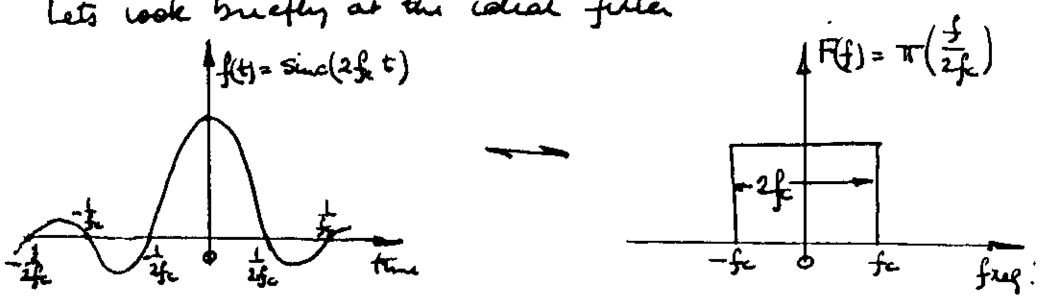
The case shown is for critical sampling & it is then clear that the replicated spectra just touch one another ie there is no overlapping. So if we then passed this sampled data through an ideal low pass filter then the original waveform would be perfectly retrieved.

This shows that $\frac{1}{T} = 2f_c$ is the criterion for critical sampling

So that a signal must be sampled at (at least!) twice its highest frequency component.

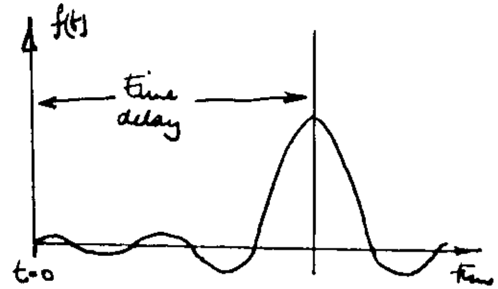
If we had sampled slower than this it is clear that the spectra would overlap (known as aliasing). Once it has occurred you can not retrieve the original information!

Lets look briefly at the ideal filter



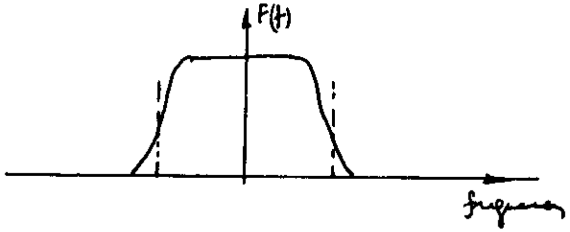
Now the time signal is identically the impulse response ie it is the response of the filter to an impulse occurring at $t=0$

As it stands it is clearly unacceptable as it is predicting the impulse! So we must arrange:

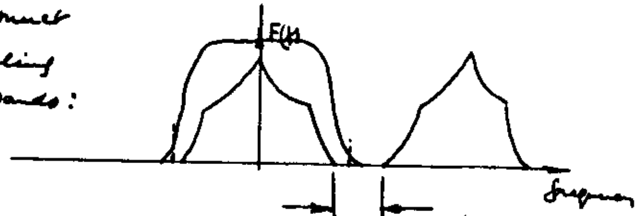


the time delay means linear phase in frequency. nothing before $t=0$ means rounded shoulders to our ideal filter.

i.e realistic filter:



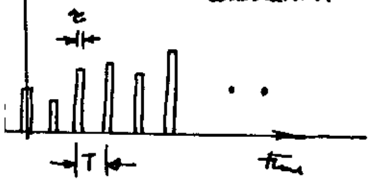
So in practice we must increase our sampling rate to allow guard bands:



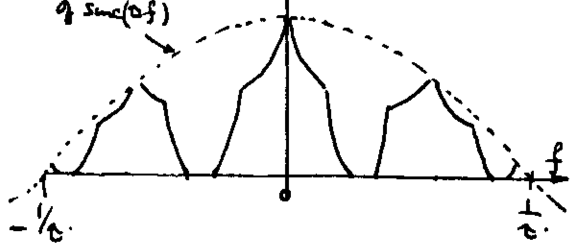
Also in a more practical situation we would not expect the sampling impulses to be infinitely short but rather to be finite in time. i.e.

$f(t) = \Pi(t/T)\cdot g(t) * \Pi(t/\tau)$

convolution!



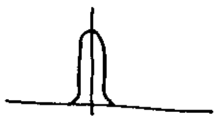
$F(f) = \Pi(Tf) * G(f)\cdot \mathrm{sinc}(\tau f)$



The original information is still available.

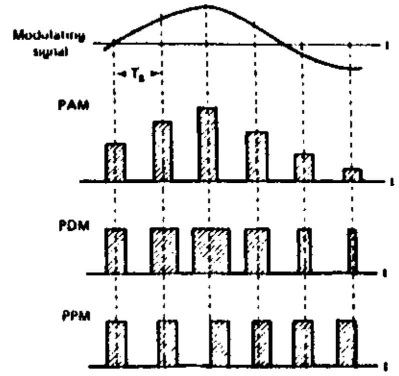
Even these perfect rectangular pulses are unreal in practice

rather:



So the basic possibilities for pulse modulation can be summarised as:

Where the sampling rate is always such as to satisfy the Nyquist criterion

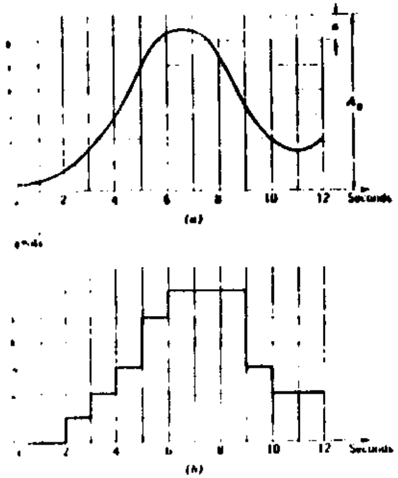


Types of pulse modulation.

All of these pulse systems are extremely sensitive to noise & it was for this reason that PCM was developed. With PCM each of these samples is represented by a binary code. Typical is 8 bits which allows 256 quantization levels.

PCM does ~~also~~ introduce Quantization error which is illustrated in the following.

For a given application it is a matter of choosing sufficient bits. 8 bits covers most things.



Quantization and sampling. (a) Given signal. (b) Quantized and sampled version.

Because we now send 8 pulses instead of the original single sample PCM does require 8 x Bandwidth (in this case)

PCM will be more fully discussed later.

"ISDN" stands for Integrated Services Digital Network.

The term integrated implies that the system should be able to cope with both analogue & digital signals.

The intent is to standardize a public end-to-end digital network which will provide a very wide range of user services.

All over the world telephone systems are being converted their exchanges & trunk systems to digital.

This is driven by:
- reducing costs
- ever increasing ability (speed)
- even better reliability.

Almost all Public Switched Telephone Networks (PSTN) are based on analogue switching & analogue transmission

Analogue telephone channel 300 – 3400 Hz.

Up till recently typical trunks have used Frequency Division Multiplexing (FDM) typically allowing 4 kHz for each telephone channel.

FDM trunks have been typically 12 MHz.

But now the Same trunk can cope with 120 bits/sec PCM. & quite immune from noise as we shall see in later lecture.

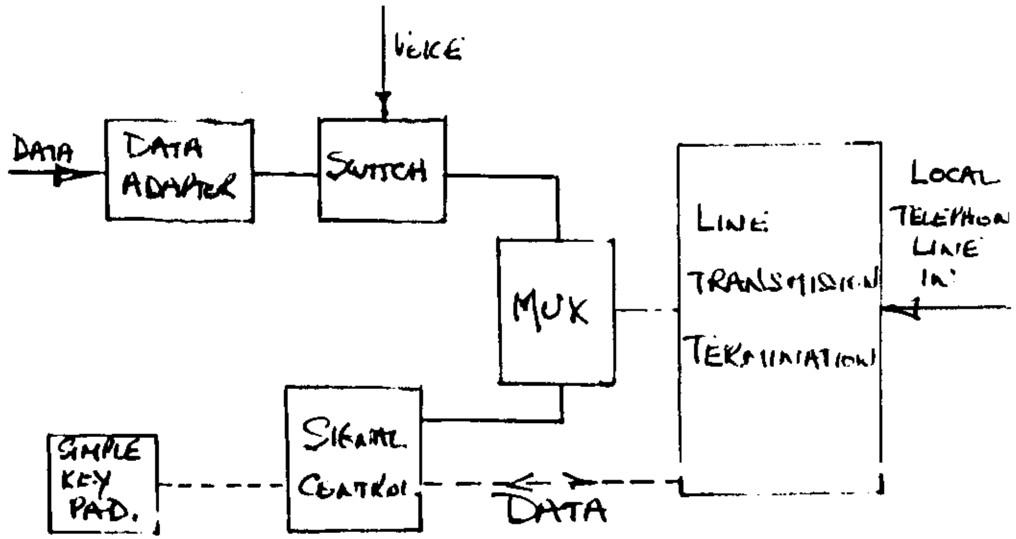
From each house to a typical exchange may be 4 kilometers simple copper wires

with normal 4kHz analogue signal there is loss of typically 15dB.

Because of the tremendous investment in the wiring system the plan is to use the same copper wires for PCM signals up to 200kbits/sec. Although 60 dB of loss is expected the PCM system will still cope!

Thus at each house a digital interface will have to be progressively introduced.

This is called a NETWORK TERMINATING EQUIPMENT (NTE) which has the basic form:

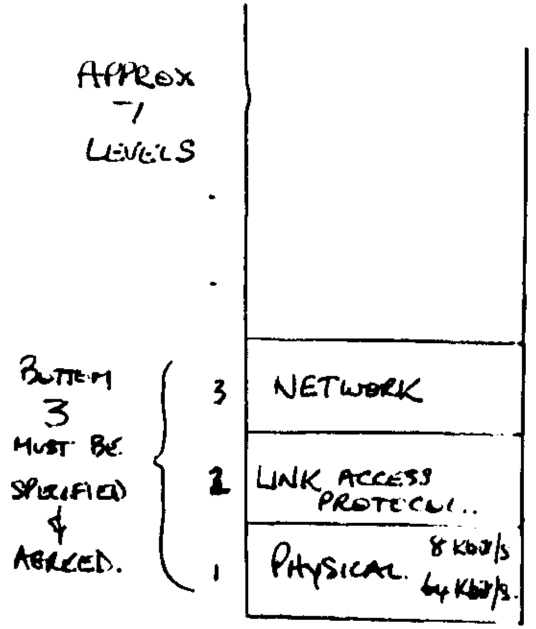


There has been difficulty in agreeing a world standard for this equipment.

ub File now has been used the Specification by the "International Standards Organization (ISO)"
REFERENCE:

ISO – TC 97/SC16/537

"DATA PROCESSING — OPEN SYSTEMS INTERCONNECTION — BASIC REFERENCE MODEL"



· THIS IS THE VITAL AREA CONTAINING
- FLAGS
- ADDRESSES
- CONTROL & ACKNOWLEDGE
- INFO FIELDS
- CYCLIC REDUND

RECENT CCITT RECOMMENDATION I420 CALLS FOR 2 × 64 Kbit/s
+ 16 Kbit/s overhead
144 kbit/s.
+ MORE OVERHEAD FOR EXTERNAL FLEXIBILITY.
→ 192 kbits/s.

It is hoped that full agreement will be reached on this specification in the near future.

The broad concept of the Complete ISDN system is summarised:

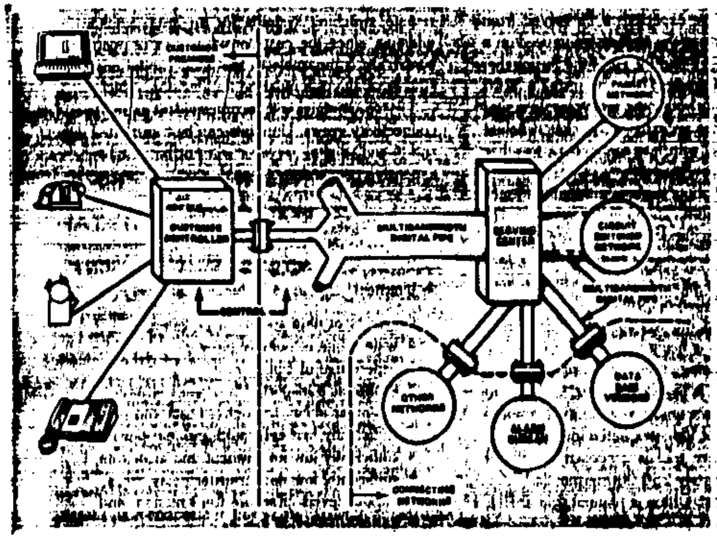


Fig. 1. ISDN Local access.

The Customer will control the bandwidth he uses (to the degree possible). Multibandwidth digital paths are already operating in excess of 600 MBits/sec. & no limit is presently envisaged for this.
The typical arrangement for a large connection is expected to be:

Totals 2048 Mbits/sec.
- 30 × 64 k bits/sec. (Customer controlled)
- 1 × 64 kbits/sec (Signalling).
- 1 × 64 kbits/sec (framing & alarms).

The Spread of Services that is expected into the future is summarised by:

Three generations of telecommunications services

Present services	New services	Possible future services
Telephony	Digitized telephony	Videotelephony
Telex, teletex	Textfax	Fast facsimile
Low-speed data	Videoconferencing	Bulk document transfer
Mobile telephony	Audiographic teleconferencing	High-speed data
Low-speed facsimile	Electronic mail	High quality videotex
	Wider availability of mobile telephony	On-line graphic design
	Higher resolution videotex	Remote printing and publishing
		Dynamic computer load-sharing
		Burst-mode host-to-host transfer

Source: Based on Malcolm Ross, "Investment Opportunities in Telecommunications Markets in Europe", paper presented to the 1983 Symposium of the European Venture Capital Association, Amsterdam.

Probably the only difficulty that is still existing in this area is the problem of ensuring absolute SECURITY OF telecommunications!

Some idea of the growth that can be expected generally in the digital communications area can be obtained by considering the National plan for Telecommunications in Italy 1985 – 1994 which projects the following Terminal growth:

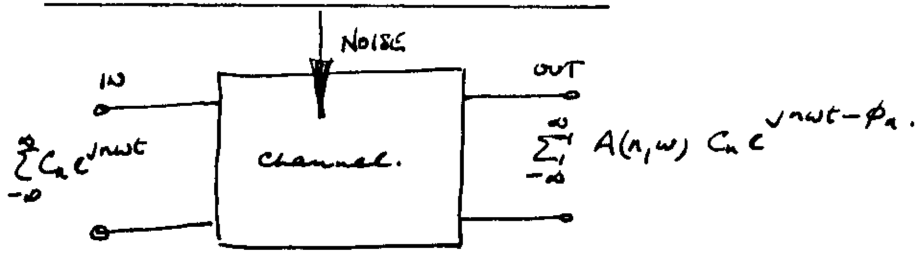
	1984	1994
Telex subscribers	64 000	120 000
Data transmission terminals	137 000	522 000
Facsimile terminals	8 000	40 000
Teletex terminals	700	140 000
Videotex terminals	2 000	250 000
Value-added services : terminals	–	250 000

Similar figures can be expected worldwide!

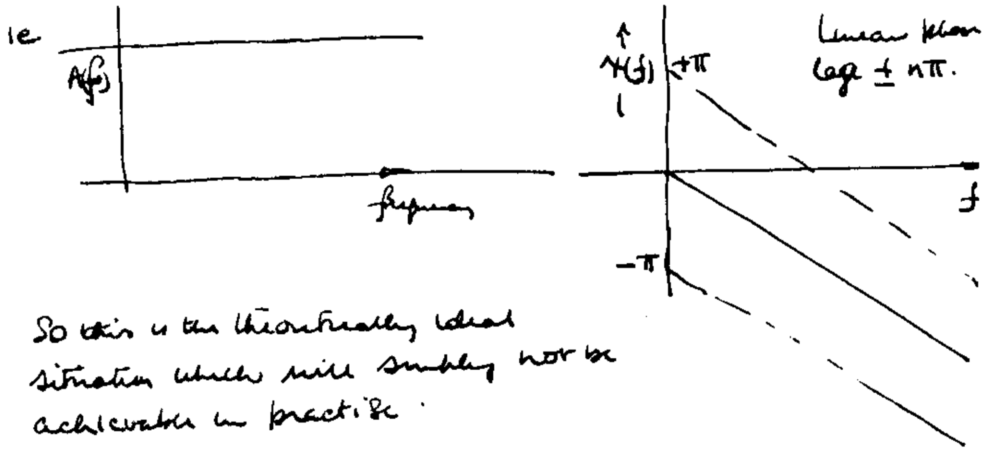
References:

1. 'Telephone Nets go digital' by I. Dorros. IEEE Spectrum April '83.
2. 'ISDN' by I. Dorros. IEEE Communications Magazine March '81
3. 'Future developments in Telecommunications' by Noel Martin, Prentice Hall (Library 621.39).
4. 'Communication Theory – The Digital Revolution', D.A. Bell, Wireless World, June '76

BASEBAND PULSE TRANSMISSION

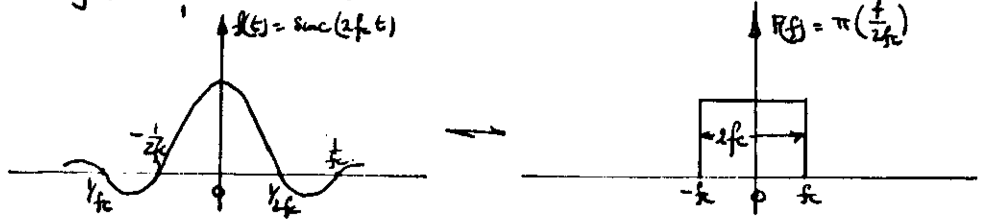


We can treat any communication channel essentially as a two port network. And we know that if the channel is to be linear then all frequency components have to be amplified or attenuated equally & we can introduce a phase lag to each component provided it is proportional to the component frequency & this results in a time delay.



So this is the theoretically ideal situation which will surely not be achievable in practice.

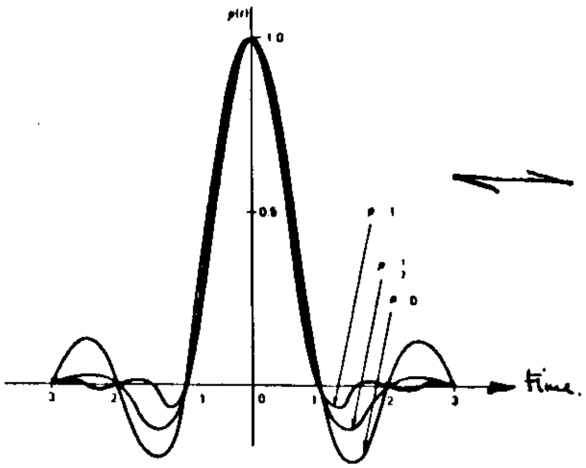
However it is good to start off considering an ideal system & then coming back to what we can achieve in practice



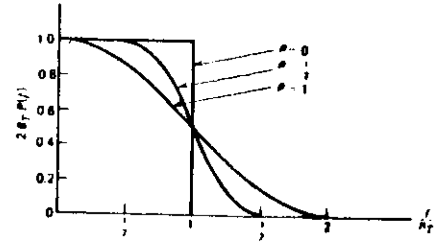
So that if we could generate a sinc shaped pulse there is no doubt that it would pass through our low pass filter with no difficulty.

It should also be clear that if we were to use a pulse repetition rate such that the pulse spacing was $\frac{1}{2f_c}$ then all pulses would contribute nothing at all to one particular pulse. Note here the effect of 2 & 3 on pulse #1.

There is obviously considerable merit in retaining this sinc function since it means that the (ISI) INTER SYMBOL INTERFERENCE is ideally zero.
In practice we try to achieve this result by using a cosine roll off to our ideal filter characteristic:



(b)

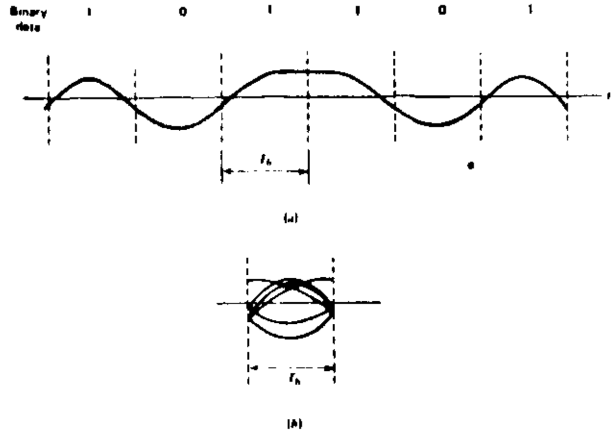


(a)

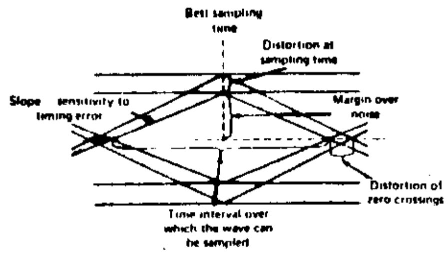
The rather more practical filter shape retains the ability for zero ISI but at the expense of increased bandwidth.

EYE PATTERN

A very useful measure of the Inter Symbol Interference (ISI) in a channel of digital data is provided by applying the received (simple binary shown here) digital signal to the vertical input of an oscilloscope & a sawtooth wave with the same frequency as the symbol rate to the horizontal input.
The following summarises the result & its interpretation:



(a) Distorted binary wave (b) Eye pattern

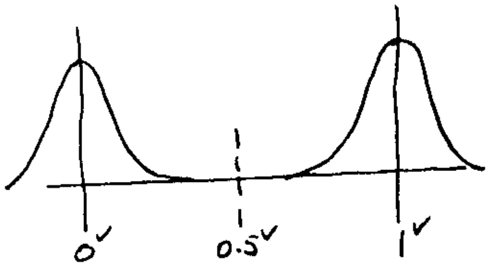


Interpretation of the eye pattern

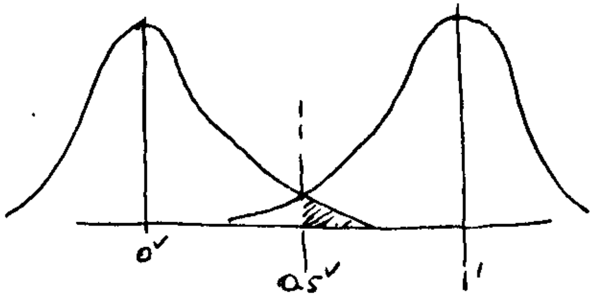
The effect of noise on a (simple binary) digital system.

With analogue systems noise is generally a difficult thing to cope with.
This is not the case for digital systems where it is much easier to draw the dividing line between acceptable & unacceptable performance.
Consider a simple binary system which has levels of 0 & 1 volt & a decision level midway at 0.5 volts.

So:

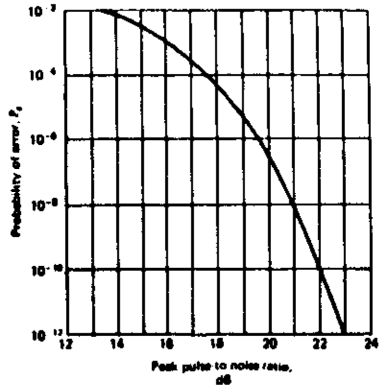


We erect Gaussian probability density functions for the amount of noise in the system & assess the effect at 0.5v
For this case the probability of error is quite negligible.



With increased noise the probability of a '1' occurring when we transmit a '0' is given by the shaded area under the curve. Because the curves are identical & symmetrical the probability of a '0' occurring when a '1' is transmitted is the same.
These can be read from tables & typical values are shown here.

Peak pulse-to-noise ratio γ	Probability of error P_e	This is about one error every
13.3 dB	10^{-2}	$10^{[illegible]}$ second
17.4	10^{-4}	$10^{[illegible]}$ second
19.6	10^{-6}	10 seconds
21.0	10^{-8}	20 minutes
22.0	10^{-10}	1 day
23.0	10^{-12}	3 months



Probability of error in a PCM

This result is summarised here. Experience has shown that with a signal/noise of 21 dB the system is essentially noise free.

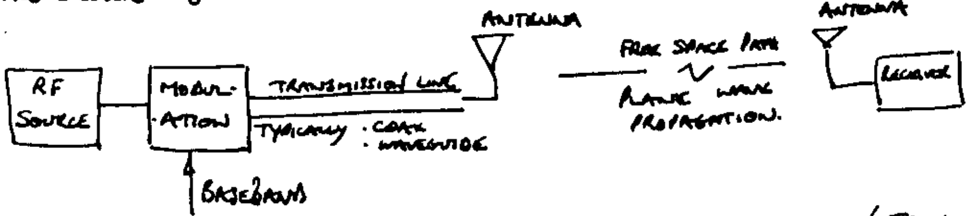
This is why PCM is so widely used — because a comparable analogue channel would require a signal/noise in excess of 70 dB. In fact we get the remarkable result that by changing to PCM (say with 8 bits) we require 8 times the bandwidth but because of the massively better noise performance we actually win!

The only drawback which exists with PCM is the effects of quantising error but for any situation it is simply a matter of choosing sufficient quantising levels to minimise the deleterious effects.
For example the use of 8 bits (256 quantising levels) gives entirely satisfactory performance for high quality audio & even television!

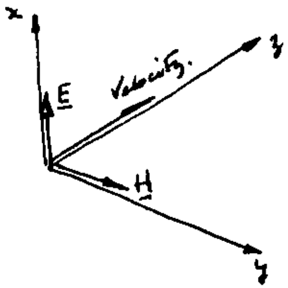
TRANSMISSION MEDIA — GUIDED WAVES.

This is our first distinct departure from signals and we now begin to consider fields & waves.

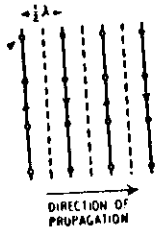
The broad system concept we now have is:



So we need to have a good idea of what a plane wave (≡ TEM wave) looks like. Lines of E & H are transverse to the direction of propagation and always form a right handed system.



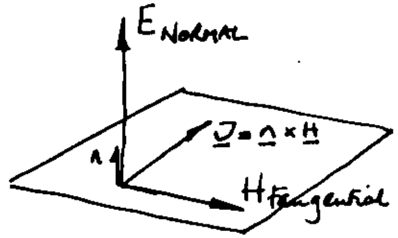
In free space the velocity of the wave is the same as the velocity of light $c \doteq 3 \times 10^8$ metres/second.



So we developed a picture of a slice of a travelling plane wave where the lines we see are magnetic field maxima (dotted in between are zeroes) & the electric field is correspondingly in and out of the plane of this page (heavy dots coming out)

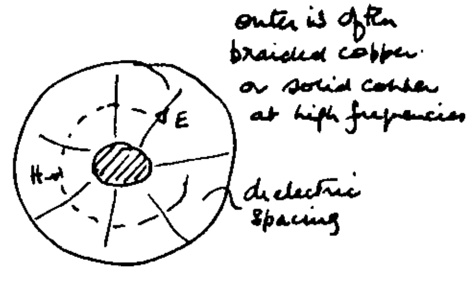
A lot of field theory analysis deals with the results of fields interacting with bodies. For our transmission media we shall consider only perfectly conducting situations and the boundary conditions then are particularly simple.

Boundary conditions at a perfect conductor:



The electric field must be normal at the conductor. The magnetic field must be tangential and changes smoothly into surface current given by $\underline{J} = \underline{n} \times \underline{H}$ where $\underline{n}$ is a unit normal vector to the surface.

Now in coaxial cable the cross-section has the form:



The 'plane-wave' now takes the form shown and it propagates down the line with the velocity of the wave in the dielectric (which may be air, dielectric spacers, or solid dielectric)

velocity $= \sqrt{\varepsilon\mu} = \sqrt{\varepsilon_r \varepsilon_0 \mu_0}$

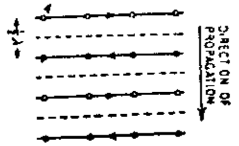
$= \sqrt{\varepsilon_r} \cdot$ velocity of light

Some typical coaxial performance:

O.D. of wire, 2a (in.)	I.D. of tube, 2b (in.)	Dielectric	Z_0 (ohms)	α (db/100 ft)		Peak voltage (kV)	Mean power (watts)		Relative velocity v_r
				30 Mc/s	300 Mc/s		30 Mc/s	300 Mc/s	
0·056	0·334	Polyethylene	72	1·05	3·8	12	640	130	0·67
0·022	0·133	Polyethylene	72	2·2	7·5	4·5	160	44	0·67
0·155	0·555	Partial air (disc spacers)	75	0·45	1·6	3·5	1,000	300	0·86
0·128	0·755	Partial air (disc spacers)	100	0·30	1·05	6	1,400	380	0·96

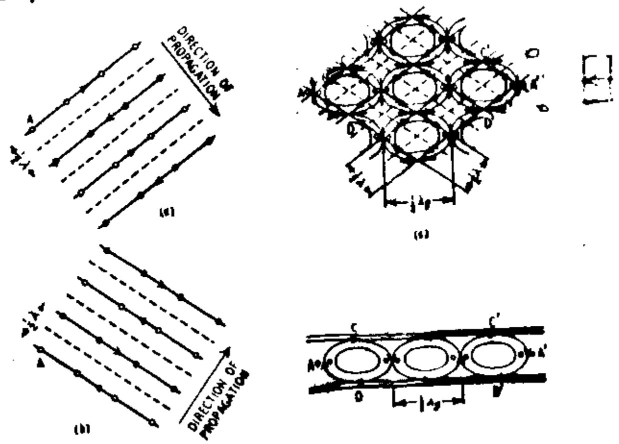
Ideas of characteristic impedance and the effects of mismatch on transmission lines (which includes waveguides) is included in a separate article.

Now moving on to waveguides:



If we consider the plane wave shown here propagating down the page then you see that we can insert an infinite perfectly conducting sheet in the plane of the page and it will satisfy the boundary conditions perfectly – In fact the wave will not even realize that it is there! We could then introduce a second such sheet parallel to & spaced apart from the first & only retain that portion of the wave which exists in between the two sheets. This is the so called parallel plate transmission line. With air between the two sheets the wave will simply propagate with the velocity of light.

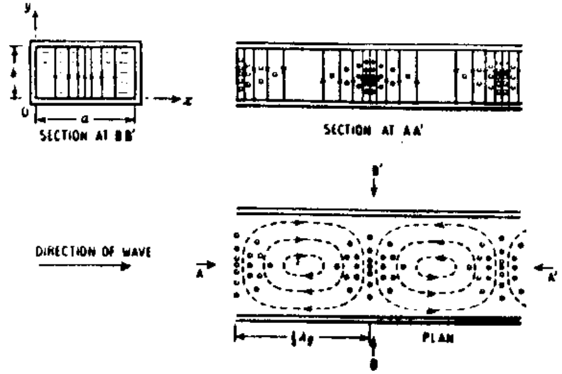
If we now added two such plane waves travelling simultaneously in a parallel plate transmission line:



When we add the two waves together we get loops formed & careful inspection will show that we can insert two further plates normal to the plane of the paper which encloses one set of loops. All boundary conditions are satisfied as the electric field is identically zero along these new walls.

This arrangement perfectly describes the so called "dominant mode" in rectangular waveguide.

In more detail this dominant mode looks like:

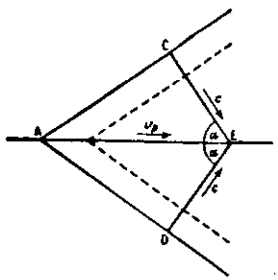


The component plane wave bounces at different angles at different frequencies:

Zigzag path of component waves in rectangular waveguide. (a) $\lambda = 0{\cdot}5\lambda_c$; (b) $\lambda = 0{\cdot}9\lambda_c$.

When α reaches 90° in this diagram the component waves bounce from side to side & do not progress down the guide at all. In this condition you can see that one half wave length of the free space wave exactly fits across the broad wall dimension. This defines the (free space) cut off wavelength as $\frac{\lambda}{2}\Big|_{\text{cutoff}} =$ broad wall dimension.

The actual field pattern in the waveguide does not travel with the velocity of light:

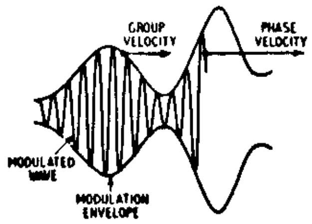


While the plane wavefront moves from C to E (at vel of light) the point A moves to E

So A moves at $V_{\text{phase}} = \frac{c}{\cos\alpha} > c$

The actual progress of the plane wave down the guide is

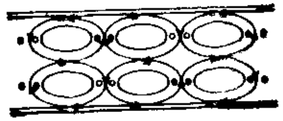
$V_{\text{group}} = c\cos\alpha < c$



. Phase and group velocities in a modulated carrier wave.

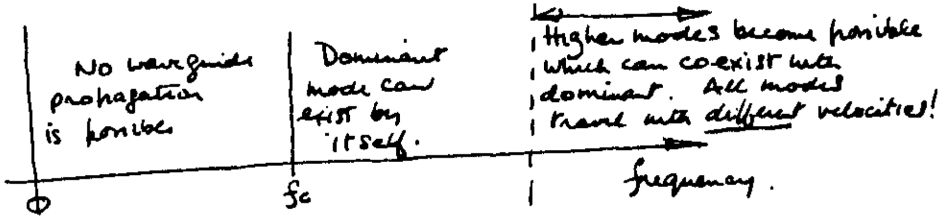
This means that in the waveguide transmission of information the 'carrier' appears to pass through the modulation.

We could equally well have enclosed two loops as:

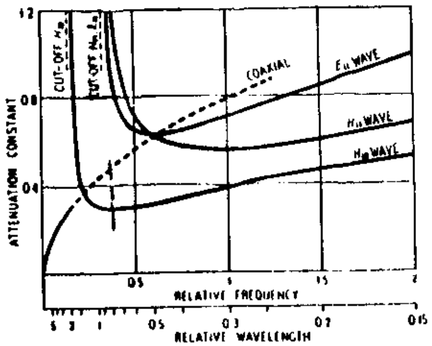


This is a higher mode of there are many types of possibilities. However dominant mode is the only one which exists by itself.

So we get a picture of the operation of waveguide as:



Typical figures for loss:



Attenuation constant of E_{11}, H_{11}, and H_{10} waves in rectangular copper waveguide with $a/b = 2$, and of fundamental mode in air-spaced coaxial line of optimum dimensions and equal cross-sectional area.

Multiply frequency scale by $10^{9}/a$ (cm) to get Mc/s

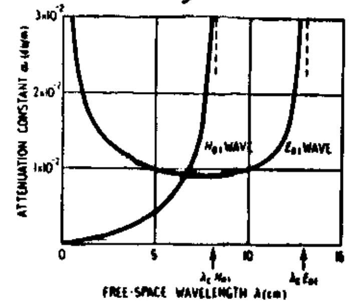
Multiply wavelength scale by a (cm) to get λ (cm)

Multiply attenuation scale by $a^{-3/2}$ (cm) to get db/m

Examples of waveguide sizes (internal) of performance.
All rectangular waveguides

Guide dimensions (in.)				Recommended[1] operating wave-length range (cm)	Attenuation[2] (db/m)	Power[3] rating (MW)
Internal section		External section				
Width	Height	Width	Height			
3·500	1·750	3·660	1·910	14·2 – 9·5	0·013	3·9
2·840	1·340	3·000	1·500	11·5 – 7·7	0·018	2·4
2·372	1·122	2·500	1·250	9·6 – 6·4	0·024	1·7
1·872	0·872	2·000	1·000	7·6 – 5·1	0·034	1·0
1·590	0·795	1·718	0·923	6·5 – 4·3	0·041	0·81
1·372	0·622	1·500	0·750	5·6 – 3·7	0·056	0·54
1·122	0·497	1·250	0·625	4·6 – 3·0	0·077	0·35
0·900	0·400	1·000	0·500	3·6 – 2·4	0·106	0·23

We also make use of circular waveguides and it is important to be aware that there is one particular mode in circular which can be arranged to have minimal loss:



There are examples of trunk routing which use this H01 mode in circular guide but is being rapidly overtaken (if not passed) by optical fibre performance.

Mismatches are considered in a separate note.

COMMUNICATION SYSTEMS — RADIO

The physical implementation of a communication link has many possibilities & the choice we have to make is often not a simple one and depends on many physical & sometimes even political circumstances!

The choice in fact ranges from transmission lines which were covered in the last lecture & a variety of radio links.

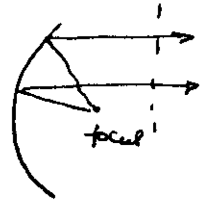
A typical link might include:

TELEPHONE LINES	—	NARROW BANDWIDTH.
V.L.F. RADIO	—	V. NARROW BANDWIDTH
H.F. RADIO	—	IONOSPHERIC PROBLEMS
V.H.F. LINKS	—	TERRAIN PROBLEMS
MICROWAVE LINKS	—	TERRAIN PROBLEMS
{ COAXIAL CABLE / OPTICAL FIBRE	—	COST
SATELLITE LINKS	—	COST.

For the remainder of this lecture we shall concentrate on microwave systems (≡ Microwave links & Satellite links).

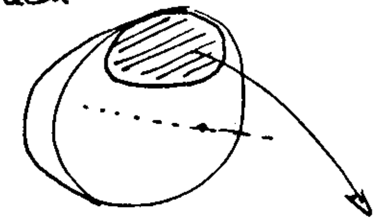
For this purpose we might class the frequency range of interest as 2 GHz — 100 GHz { Top limit is just present day practise.

Predominantly we use reflector type antennas most commonly based on the parabola which has a classical focussing ability

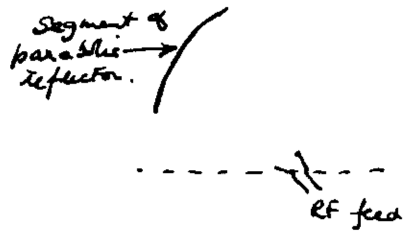


We can place a feed horn at the focus but then have the problem of feeding that horn to the ground

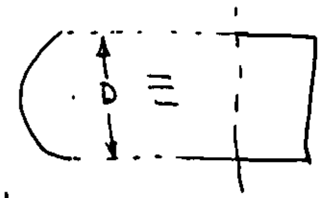
We can get around this problem by using a Cassegrain system with a subreflector which is typically hyperboloidal in shape. The RF feed is now conveniently at the back of the dish.
blockage may be 10%



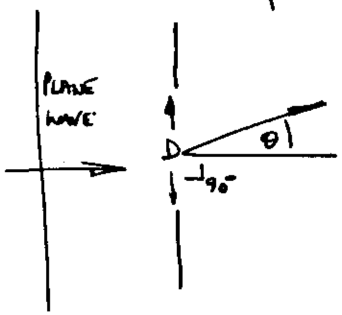
To overcome this we sometimes use just a segment of a full parabola in an offset feed arrangement. We need to be careful with cross-polarization quality with this arrangement.



A simple but extremely useful analysis of such reflector antennas can be carried out as follows:

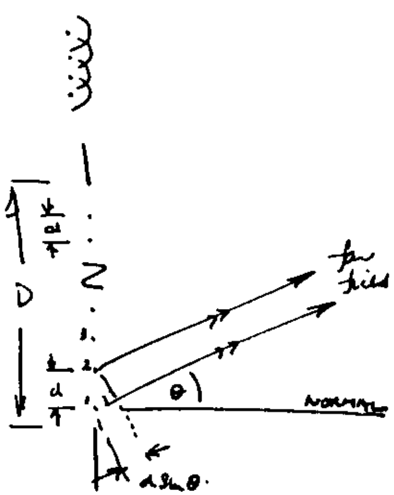


Assume the reflector antenna is equivalent to a uniformly illuminated aperture (D). Thru' our Fourier Transform we would straight away expect a Sinc function radiation patt.



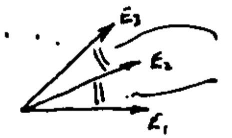
This is equivalent to an aperture being illuminated by a plane wave & calculate the result to the right. The aperture is assumed to have the same dimensions as the (paraboloid) reflector.

We then use Huygen's Principle which allows us to treat each point in the aperture as an individual source of a secondary spherical wave.



So we can say that our aperture D is equivalent to n metres such point sources spaced d apart.

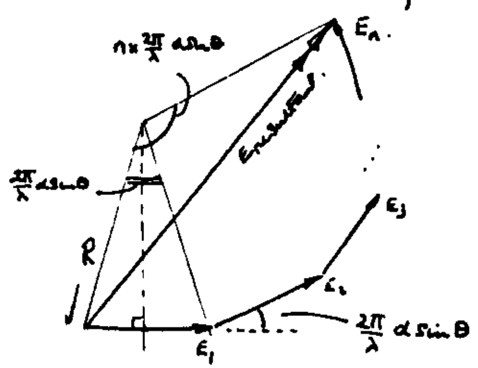
The contributions from point sources 1, 2, 3 .... would have the following form in the far field:



$\frac{2\pi}{\lambda} \cdot d \sin\theta$ radians

$|E_1| = |E_2| = |E_3| = \cdots$

which we can draw alternatively:



$$E_{Resultant} = 2R \sin\left(\frac{n \cdot \frac{2\pi}{\lambda} d \sin\theta}{2}\right)$$

$$\& \quad \sin\left(\frac{\frac{2\pi}{\lambda} d\sin\theta}{2}\right) = \frac{\frac{E_1}{2}}{R}$$

So.

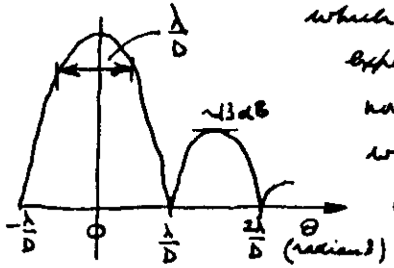
$$E_{Res} = \frac{E_1 \sin\left(\frac{n\pi d}{\lambda}\sin\theta\right)}{\sin\left(\frac{\pi d}{\lambda}\sin\theta\right)}$$

Now the maximum far field signal will occur normal to the aperture when all the contributions will be exactly in phase & $E_{max} = n E_1$

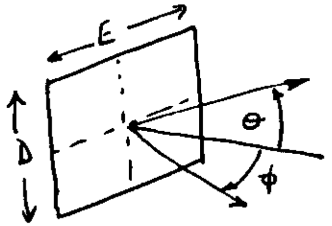
So we can write $\quad E_{res} = \dfrac{E_{max}}{n} \dfrac{\sin\left(\frac{n\pi d}{\lambda}\sin\theta\right)}{\sin\left(\frac{\pi d}{\lambda}\sin\theta\right)}$

If we now let the number n of elementary radiators making up the aperture D approach infinity & their spacing d approach zero in such a way that $n \cdot d = D$

we obtain finally $\quad E_{res} = E_{max} \dfrac{\sin\left(\frac{\pi D}{\lambda}\sin\theta\right)}{\frac{\pi D}{\lambda}\sin\theta}$.



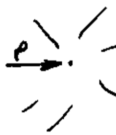
which has the 'sinc' function form which we expected. So the angle out from the normal to the first null is λ/D (radians!) and we can usefully approximate this to be the antenna beamwidth.



If the radiating aperture was rectangular $D \times E$ then the total pattern would be given by:

$$E_{res} = E_{MAX} \frac{\sin\left(\frac{\pi D}{\lambda}\sin\theta\right)}{\frac{\pi D}{\lambda}\sin\theta} \cdot \frac{\sin\left(\frac{\pi E}{\lambda}\sin\phi\right)}{\frac{\pi E}{\lambda}\sin\phi}$$

we can integrate this equation in spherical co-ordinates over the right hand half of the aperture space to get the total power radiated & we then find that



Isotropic antenna is a fictitious point source which radiates power uniformly in all directions.
So if we input power P watts then Power density at distance R = $\frac{P}{4\pi R^2}$ watts/□m.

$$\text{GAIN}\Big|_{\text{refered to isotropic antenna}} = \frac{4\pi}{\lambda^2}(a\,b).$$

$$= \frac{4\pi}{\lambda^2}(\text{effective aperture})$$

a general & very useful result.

If the aperture is circular diameter D then $G_i = \left(\frac{\pi D}{\lambda}\right)^2.$

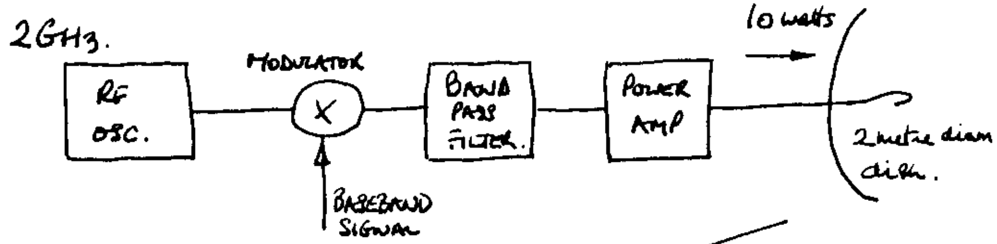
In practice the 13 dB sidelobes of the sinc function are unacceptable and we use a tapered aperture distribution — typically 10 dB down at the dish edge.
This has the effect of increasing the side lobes & reducing the gain.

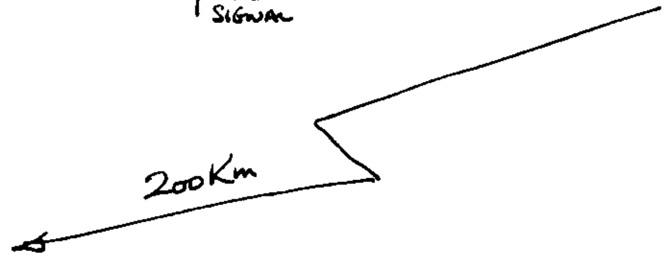
Then $\text{Gain}\big|_i \simeq (50-60\%)\left(\frac{\pi D}{\lambda}\right)^2.$

& beamwidth $\simeq 1.2 \times \frac{\lambda}{D}$ radians.

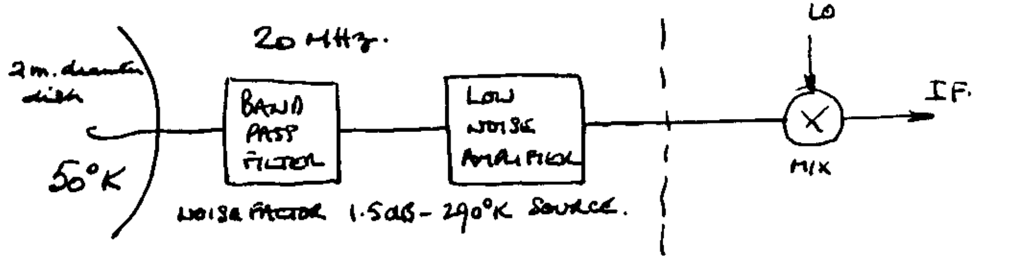
Consider then the example of a microwave link:

TRANSMITTER:





RECEIVER:



Dishes are 2 metres in diameter ∴ Gain $\simeq$ 29.8 dB.

10 watts into isotropic antenna

∴ Power density at 200 Km = $\frac{10}{4\pi \times (200{,}000)^2}$ watts/□m.

assuming antenna points correctly its gain must be added

Actual power density at receiver = 955 × (   ) watts/□m

= 1.9×10^{-8} watts/□m.

Effective area of receive antenna given by:

$\frac{4\pi}{\lambda^2}$(effective area) = 955 ∴ Effective area = 1.7 □ metres.

So received power = 3.23×10^{-8} watts.

Noise temperature of receiver = $290 (10^{0.15} - 1) = 119.6°K$.

Total noise in 20 MHz band $= k (119.6 + 50) \cdot 20\,MHz$.

$$= 1.38 \times 10^{-23} \times 169.6 \times 2 \times 10^{7}$$

$$= 4.68 \times 10^{-14} \text{ watts.}$$

So $\left.\dfrac{\text{Signal}}{\text{Noise}}\right|_{\text{complete link}} = 58.4\ dB.$

Satellite Communication — INTELSAT IV

12 × 36 MHz channels — 432 MHz.

in 36 MHz fit 750 Telephone circuits (4800 Hz allowed)

Uplink 5.925 – 6.425 GHz.

Downlink 3.7 – 4.2 GHz

Satellite spin stabilised – Stationary antennas

- 2 GLOBAL RECEIVE (LOW GAIN).
- 2 GLOBAL TRANSMIT
- 2 SPOT BEAM TRANSMIT.

Earth stations up to 100' diameter 60 dB antenna gain. 100°K. Temp.

No significant atmospheric effects

Heavy rain 1 – 2 dB path loss.

delay 45,000 miles ¼ sec.

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& many, many more!