Point to Point

26GHz RECEIVER

Parametric investigation

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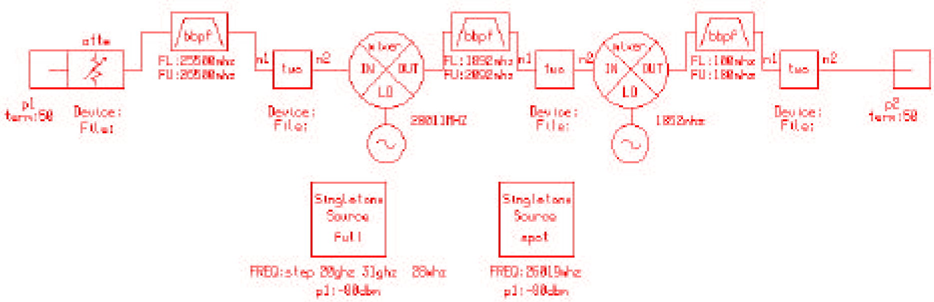
Simulating the receiver

This *theoretical* receiver analysis is based on a channel 17 or 26019MHz response. We chose this channel because it is midband. We assume various bandwidths for the modulation and demodulation, but it can be assumed that this analysis is appropriate to all the classes of data we aim to transmit and receive. Also this receiver design is characteristic and typical for all of the channels to be considered within the available ~420MHz wide band.

The block diagram of the receiver is shown below. The antenna and waveguide feeder is omitted for clarity. Each element has its own modelled characteristics. It is essential to note that at this stage of the project the data used for these modules is assumed and not measured.

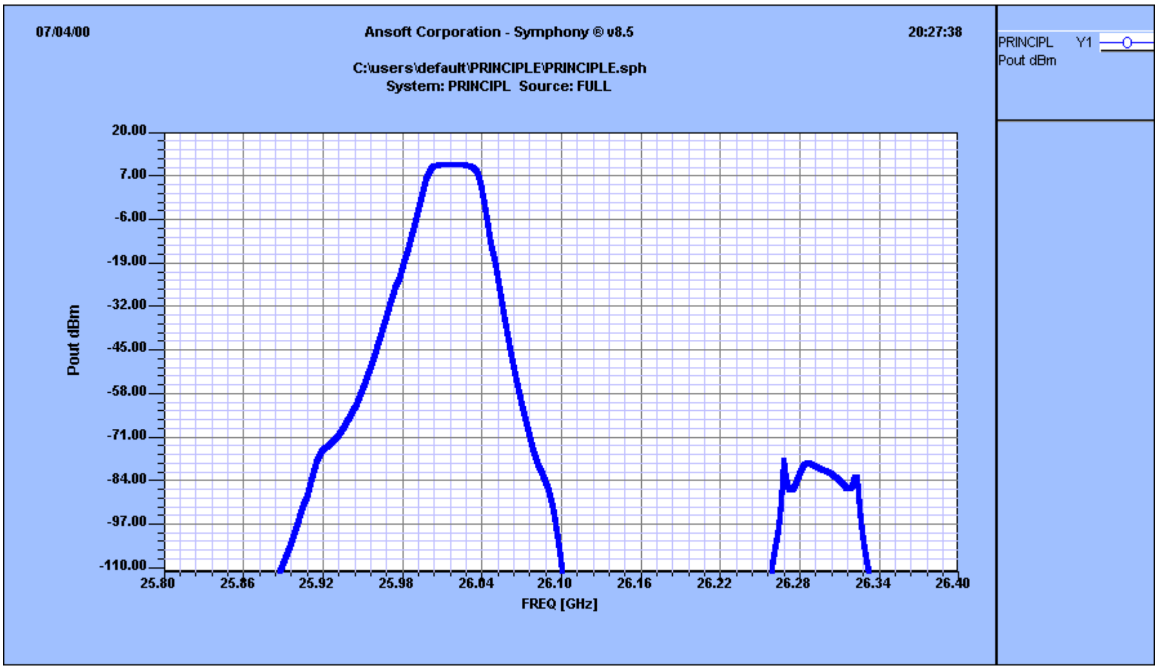


To clarify the effects the input circulator is replaced in this schematic with an equivalent attenuator of 0.3dB, which is representative of a commercially available unit operating at 26GHz. This may readily and conveniently be combined with the attenuation introduced by the input TE_{10} feeder from the waveguide to the antenna. (Although we have not included it in this analysis). Also the output duplexer is omitted since it will not effect this analysis.

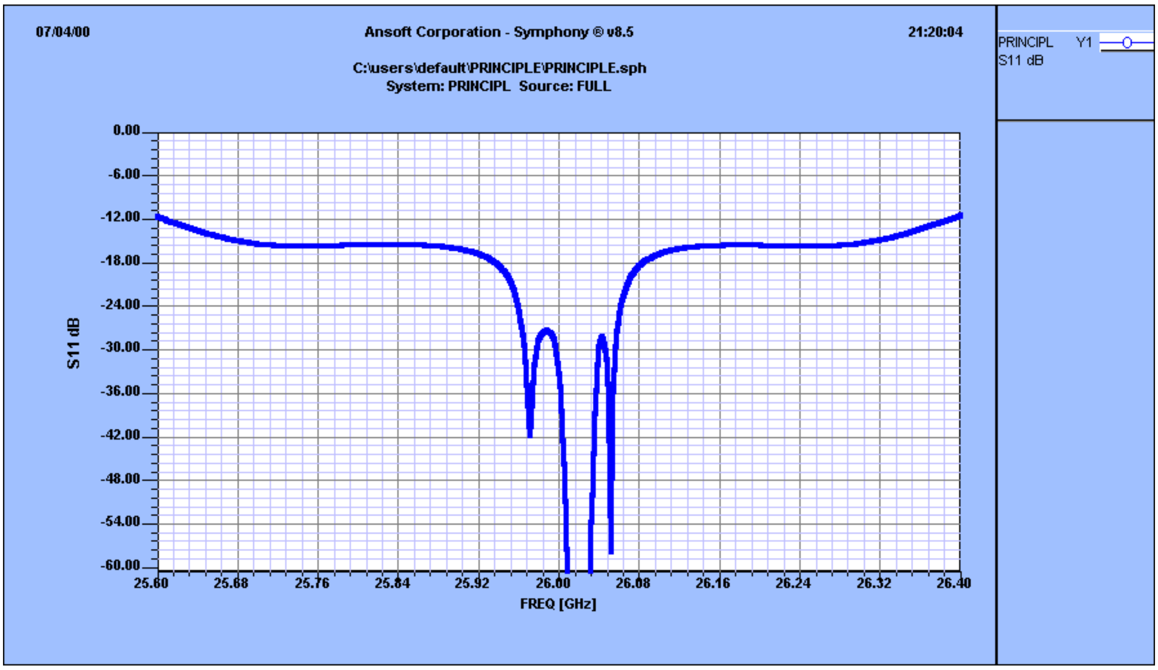
It is important to note that the responses of the receiver rely, almost entirely, on the filters employed.

The filters depend on their performance on basic limitations, the choice of filter technology I.e. waveguide, stripline, microstripline, lumped, and the quality of the materials and construction employed. In this analysis we initially ignore the details of the filters and look, parametrically, at the receiver design. CAD is excellent for parametric investigations.

Below see the frequency response of the receiver if we assume a 6 –pole Butterworth type bandpass filters with 3dB bandwidths, as indicated in the schematic diagram. The image resulting from the 140MHz 2nd IF which occurs around 26.28GHz is suppressed by more than 90dB as a consequence of all the filters preceeding it. The adjacent channel response however is due almost entirely to the 140MHz centre passband filter reponse chosen in the simulation.



The input return loss (RL) of the input filter (S11) at 26GHz is simulated to be greater than 24dB over 100MHz and >10dB (2:1 VSWR) over 1GHz. The lobing evident below 20dB RL is due to phasing and is an artefact of the filters and is unimportant <20dB. It should be noted we have used Butterworth filters. A sharper roll off, like that provided for example, by a Chebyshev response, will produce more and sharper return loss lobes. However we want to avoid sharp responses or any element which will introduce phase and amplitude shift across the *wanted* signal band since such effects will result in degraded bit error rate performance (see later).

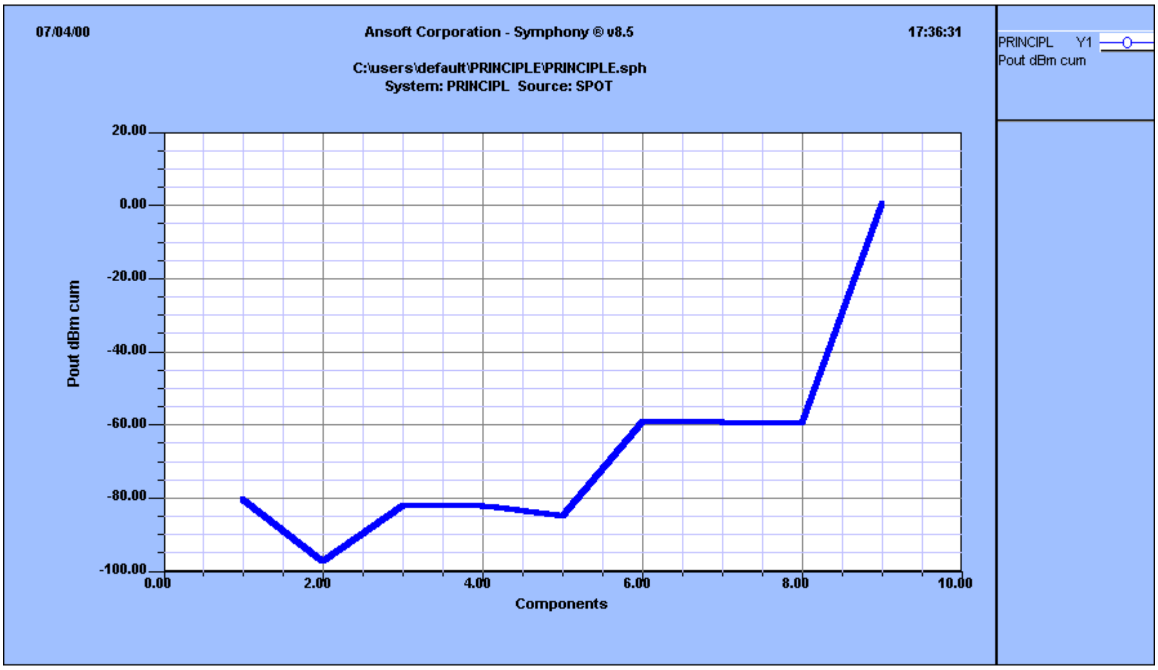


The noise figure, gain and noise distribution of the receiver depend almost entirely on the front-end amplifier and the losses in the input (image) filter and circulator (together with the input waveguide losses) according to the Friis formulation:

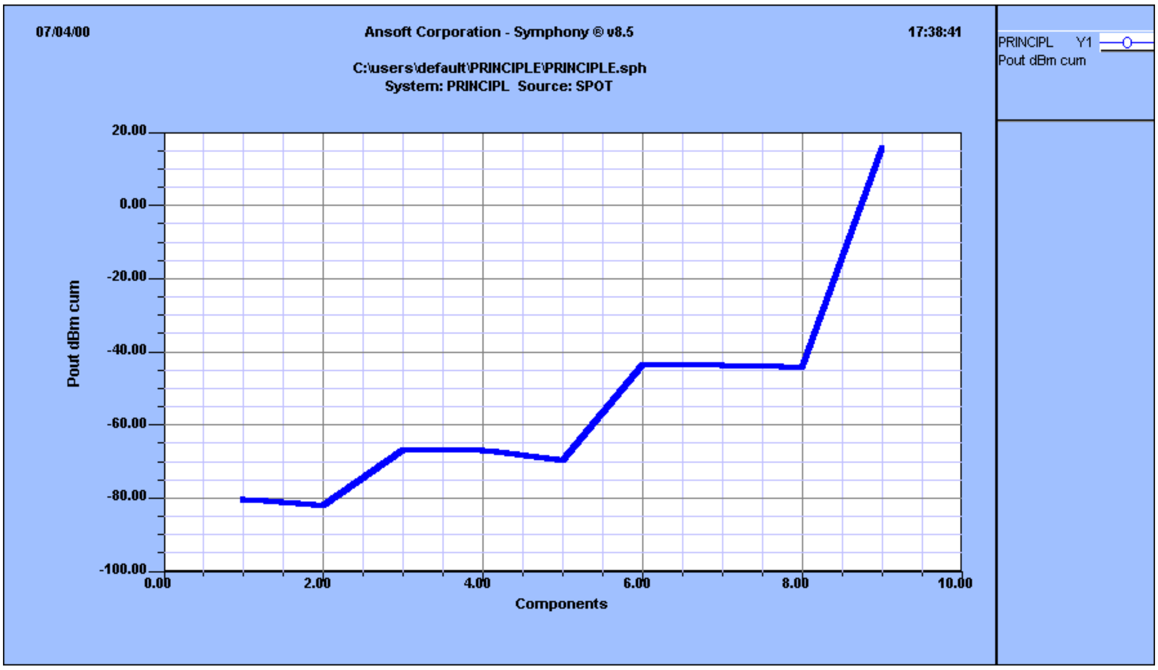
$$F_{Total} = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \text{-------------} + \frac{F_N - 1}{G_1 G_2 \text{----} G_N\text{-}1}$$

I.e. F1 and G1, where G1 is actually the loss of the circulator and input waveguide [both modelled as an attenuators].

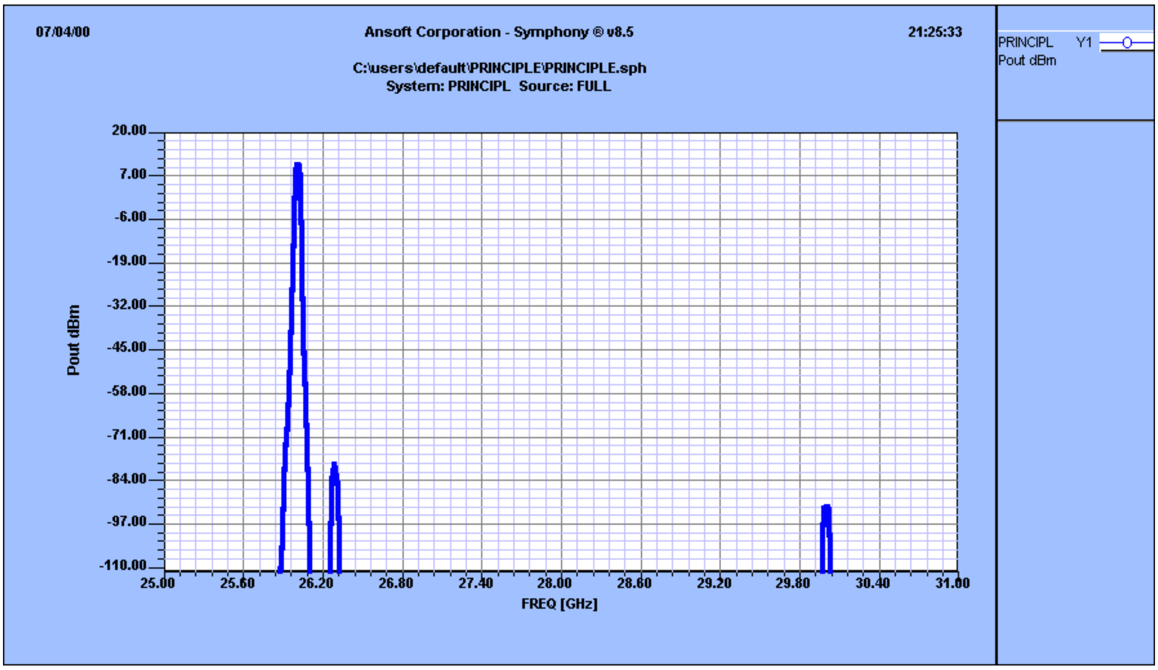
The resultant cumulative noise in the receiver chain is shown below.



Now let us suppose we have the choice of filter topologies. That is a microstripline filter or a waveguide type operating at 26GHz. The latter has a larger realisable Q. For example suppose by this means we are able to increase the Q of the input bandpass filter elements to 500 (from the 50 value assumed above). If we do this then the insertion loss decreases and hence the gain distribution and sensitivity improve as indicated in the repeated cumulative noise analysis (using these new data) below.



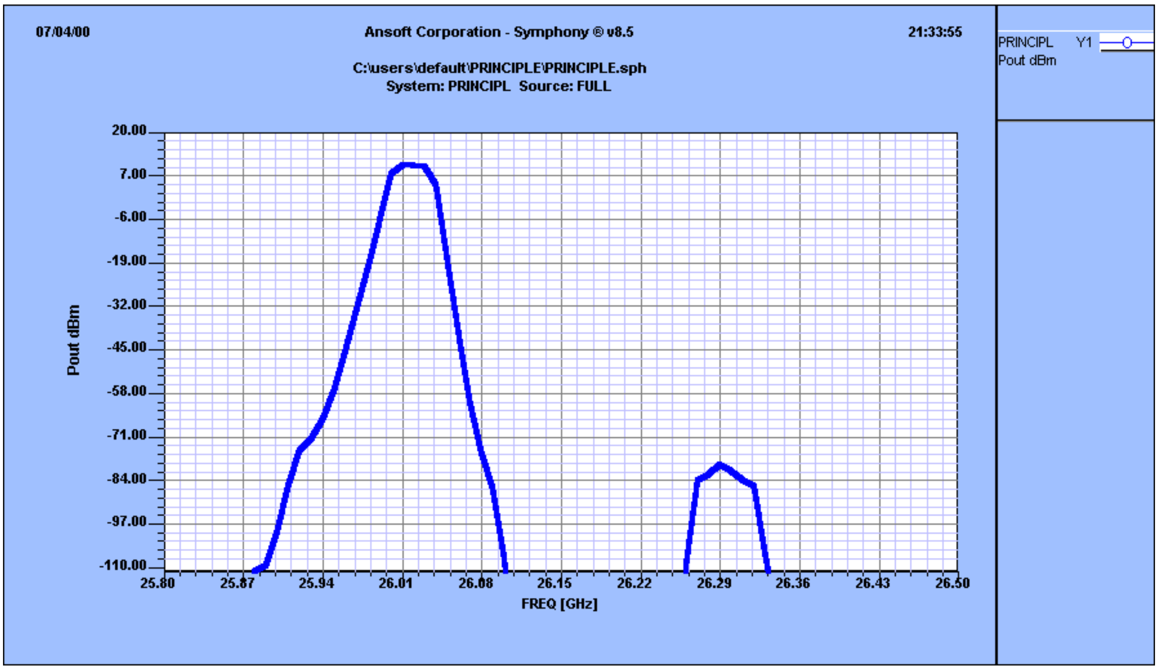
As a result, it can be seen that, the passband response and image rejection are also now improved because of the decreased losses in this filter.



Clearly it is critical how good the Q of these filter elements is a 26GHz and we must strive for the best possible Q, Silver-plating conductors if necessary.

The filter topology, the number of poles and type of construction all contribute to the resultant response. A waveguide filter offers the best performance at 26GHz. But this technology will be more costly than a printed stripline version, especially if produced in quantity, because of the accurate metalwork required for the waveguide version. So economics, as always, comes into consideration. For example if the manufacturer has invested in CNC machines then the waveguide versions may not be more expensive. In other words the OEM manufacturers methods govern their costs.

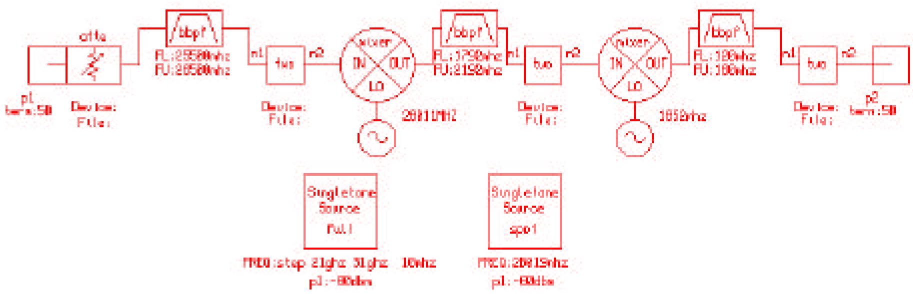
The image response at the 1st IF ~29.5GHz (i.e. 1992 +/- 28.011MHz 1st Local Oscillator) is suppressed by over 100dB using this input filter. A very good performance and certainly greater than will be the actual reality - due to imperfections and unavoidable leakage. Other unwanted responses are suppressed by >90dB again an excellent characteristic but based on practical experience again overoptimistic.



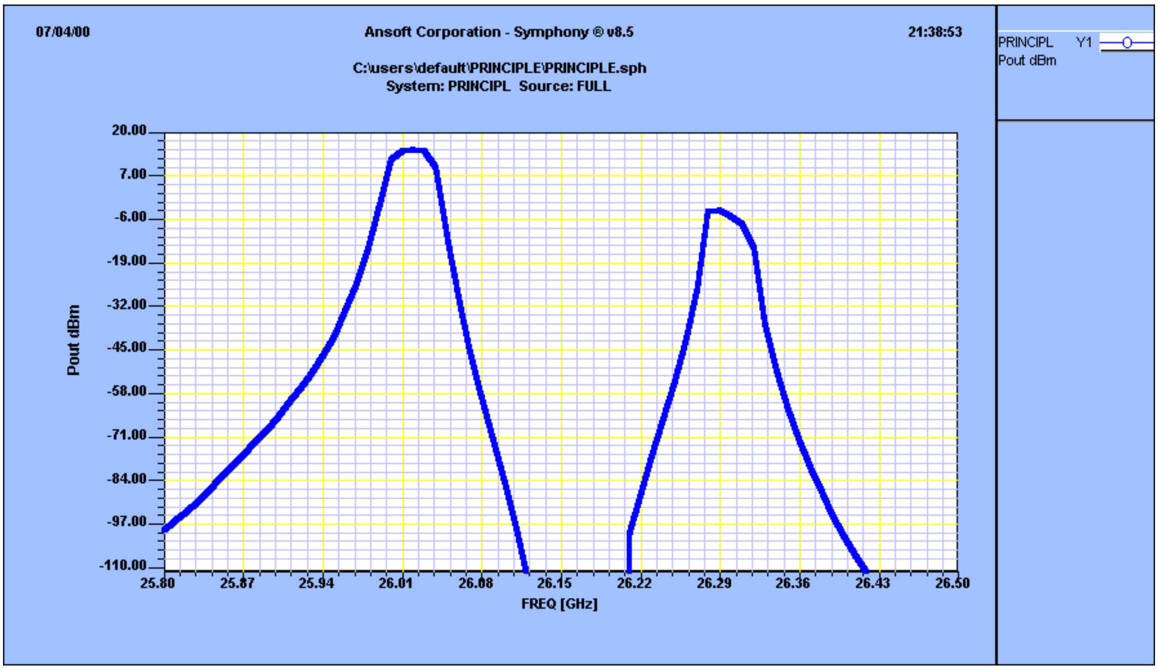
Once we have some measurements of the actual performance we can alter the CAD parameters to more closely model the reality. However although this will ultimately be useful, especially for future designs, it is not the point with our present CAD investigations. Here we are exploring the parameters of the receiver to see which parts are the most critical.

The overall response of the receiver essentially depends on the filters. In particular for the third filter operating at 140MHz. The ‘flat top’ i.e. the passband, of the left hand wanted response is primarily a result of this filter.

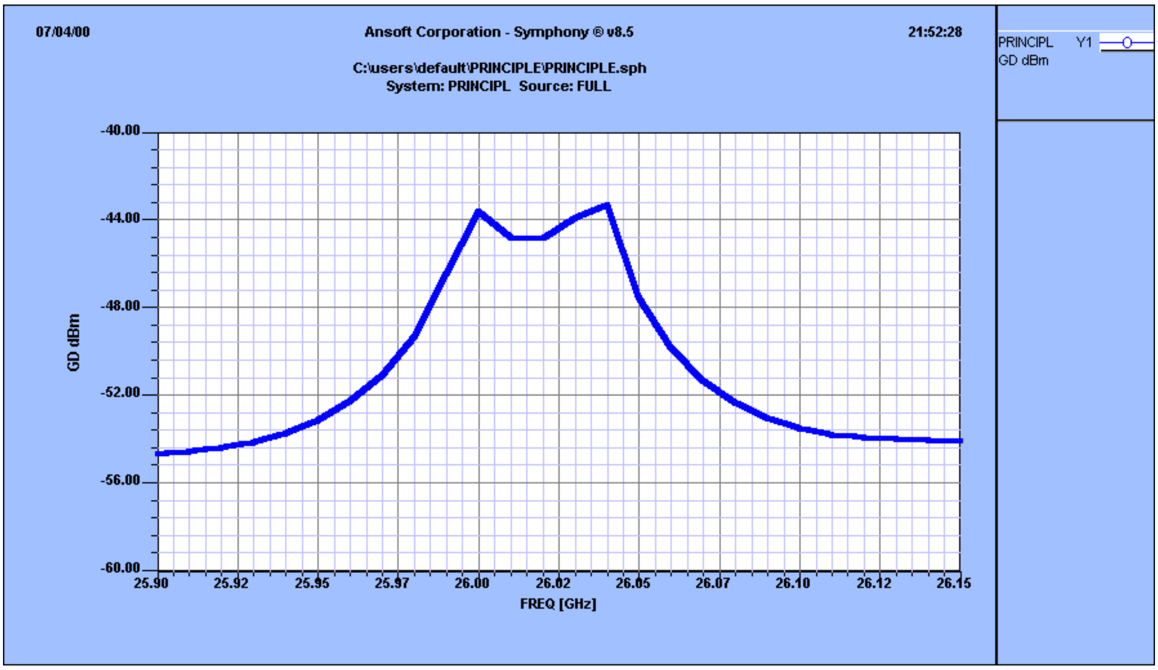
If we were to increase the bandwidth of the second bandpass filter to 400MHz, (1792 – 2192MHz) to allow for all possible channels,



Then the adjacent channel rejection suffers and the image response decreases to <10dB as indicated below.



Because we are using QPSK modulation we have to be concerned with phase and amplitude distortions *within* the wanted passband. We can readily assess the levels of distortion using a group delay analysis.



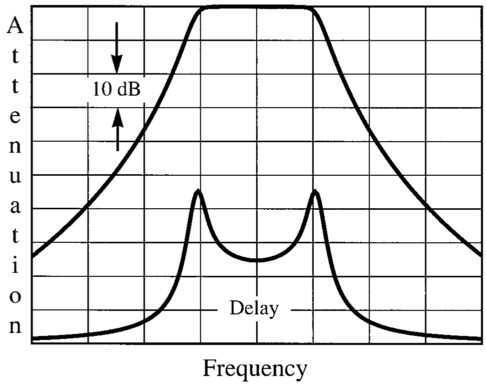
We note that, using the Butterworth filter topology and the 3dB bandwidths assumed that we have a minimal amount of delay distortion across the modulated bandwidth. This is a good result. On the other hand a Butterworth topology does not offer the best image or adjacent channel filtering because the achievable shape factor is poorer than with say a Chebyshev topology. Thus we need to compromise. In fact we may be able to employ Chebyshev characteristics provided the 3dB points are well outside the wanted passband. To illustrate this point consider below a comparison of Butterworth, Chebyshev and Elliptic type filters. Note especially the group delay characteristics. Designs must avoid placing regions where the phase or amplitude changes rapidly in the wanted passband. All these filter simulations are for 6-pole filters.

Approximation Types

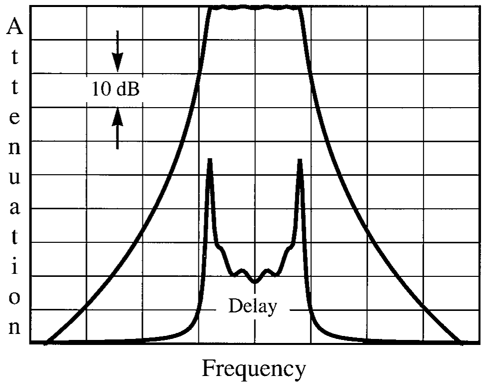
Depending upon the required specifications, an approximation type will be chosen to meet those requirements. The following are the most common types used.

Types used for control of attenuation only

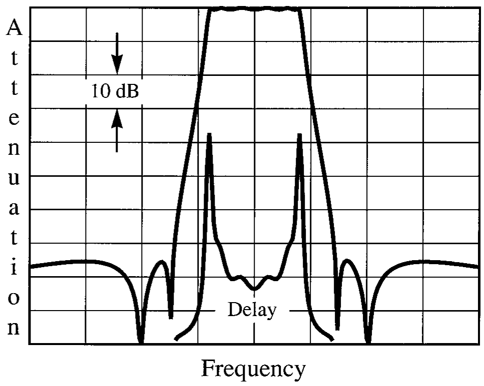
Butterworth — Produces a very smooth, flat passband with a fair rate of roll-off. This approximation produces easily realized networks.



Chebyshev — The response is similar to the Butterworth but with ripple within the passband and an improved roll-off rate. Networks obtained by this approximation are the most easily realized.



Cauer or Elliptic Function — The passband ripple is similar to the Chebyshev but with greatly improved stopband selectivity due to the addition of finite attenuation peaks. The network complexity is increased over the Butterworth or Chebyshev but it still yields practical realizations over nearly the entire operating region.



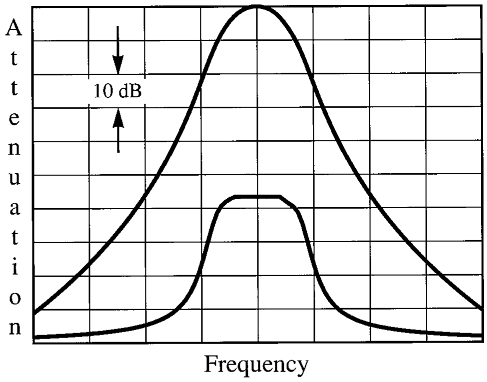
All of the responses shown here are for six pole networks with identical design bandwidths.

A Bessel filter offers the best performance for digital systems but at the expense of relatively poor adjacent channel attenuation. The Gaussian response is better in this respect but suffers some passband ripple.

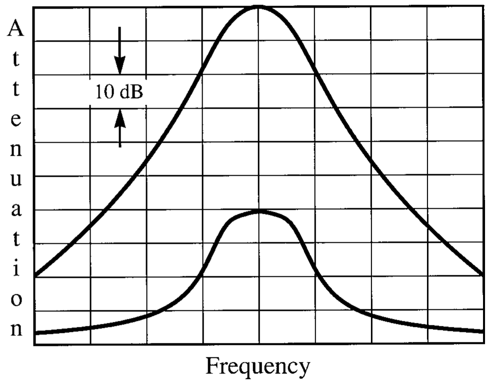
Approximation Types

Types used for phase and delay control

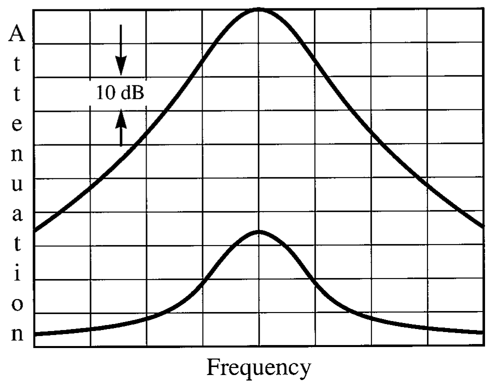
Bessel or Linear Phase — This approximation is the Butterworth of delay control. It produces filters with a flat delay around center frequency. The more poles used, the wider the flat region extends. The roll-off rate is poor. There are some realization restrictions and so these designs cannot be obtained over the entire operating region.



Gaussian — Very similar to the Bessel except that the delay has a slight “hump” at center frequency and the rate of roll-off is slower. Because of the delay response, the ringing characteristics are better than the Bessel. Realization restrictions also apply to these filters.



Synchronously Tuned — These filters have the same advantages and disadvantages as the Bessel and Gaussian except that the ringing response is the best of all design types and the roll-off is even slower than the Gaussian. As with the other two types, some realization restrictions apply.



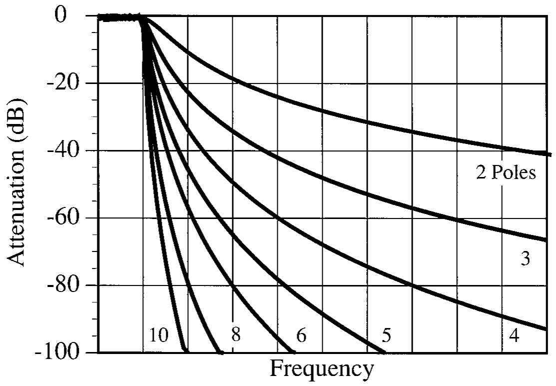
Responses shown are for 6 pole networks with identical 3 dB design bandwidths.

The table below indicates the parameters we can expect for Butterworth and Chebyshev 'n' pole filters.

Shape Factors and Practical Limits

***Theoretical* Shape Factors (60/3 dB) for Monotonic Responses**

Number of Poles	2	3	4	5	6	8	10
Filter Type							
Butterworth	32	10.0	5.6	4.0	3.2	2.4	2.0
Chebyshev 0.1 dB	29	8.5	4.4	3.0	2.3	1.7	1.4
Chebyshev 0.5 dB	27	7.7	4.0	2.7	2.2	1.6	1.4



Attenuation Characteristics for Varying Numbers of Poles

***Practical* Shape Factors (60/3 dB) for Monotonic Response**

Number of Poles	2	3	4	5	6	8	10
Filter Type							
Butterworth	32	10.0	5.8	4.2	3.3	2.5	2.2
Chebyshev 0.1 dB	30	9.0	4.7	3.1	2.4	1.8	1.5
Chebyshev 0.5 dB	29	8.0	4.2	2.8	2.3	1.7	1.5

Practical Limits

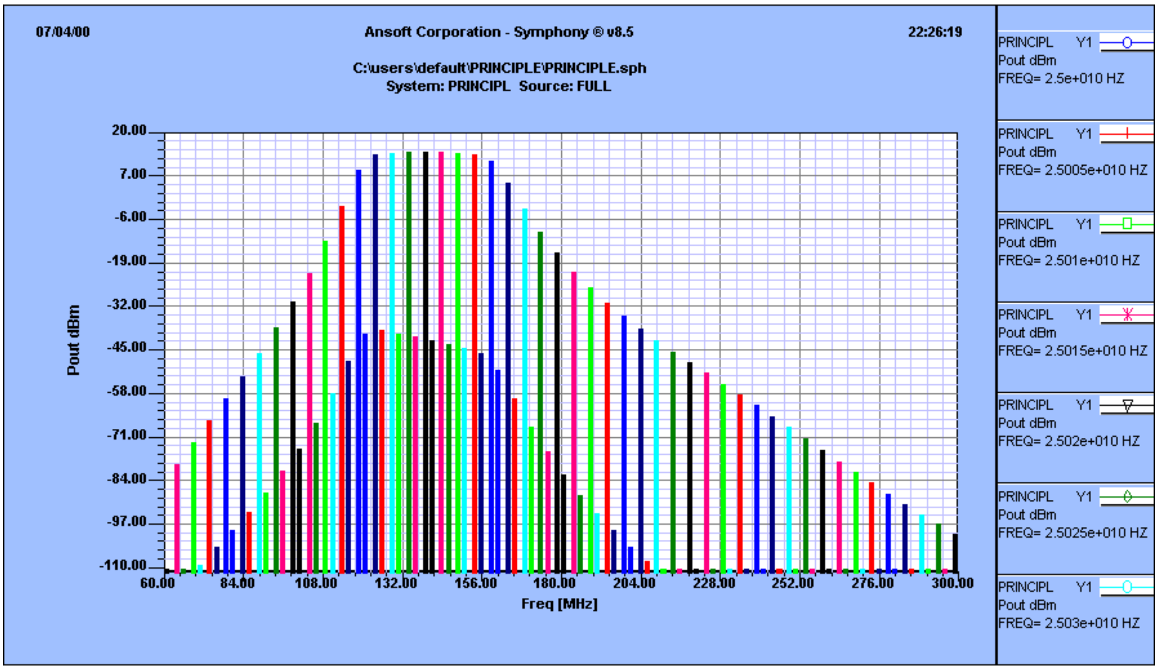
	Typical	**Best**	
Shape Factor	1.5 — 30	1.075	
Passband Flatness	1.0	0.10	dB
Narrowest Bandwidth	0.01	.001	%
Widest Bandwidth	2.5	10	%
Maximum Attenuation	90	100	dB
Phase Linearity	±5	±2	°
Phase Matching	±5 (Quad)	±3 (All)	°
Temperature Range	-20 to +70	-45 to +105	°C
Shock	15	1500	g's
Vibration	10g Sine 10-2000 Hz	45+ Grms Random	
Aging	<10ppM (10 years)	5ppM (15 Years)	

Mixer products

Mixers always produce wanted and unwanted responses. We can judge whether our design has problems by looking at the amplitude of the unwanted responses. For example for our receiver there is no exact harmonic relationship between the two local oscillators.

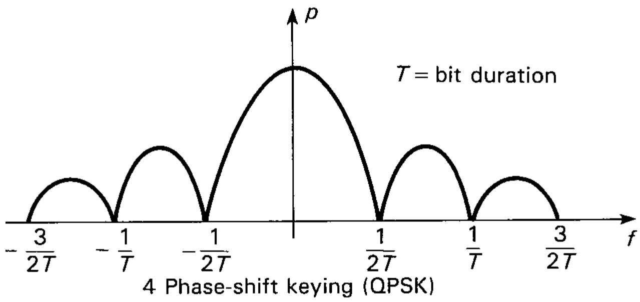
Of course we are using coherent generators (because the CAD simulator is unable to calculate a large number of spurs). However the transmitted signal has a bandwidth and is not coherent. For example the 14th harmonic of the 1852MHz 2nd local oscillator produces an output at 25928Mhz, which is in band. I.e. 26011-25928 = 83MHz. The question is 'what level does this harmonic have at the signal frequency'? We can in truth only answer this question when the system is constructed. It depends on the performance of each module and of the construction. The reverse isolation of each module and the leakage in the hardware, which is largely a function of the experience of the designers, govern the leakage. (We might expect -6dB / octave for an oscillator harmonic output, so 14 x 6 = 84db, so this should not be a problem. However poor layout and decoupling might make it so. Of course any unwanted in-band carrier, above the thermal noise, will degrade performance and reduce the receiver's sensitivity).

Theoretically calculated levels of unwanted products, centred on 140MHz final IF and produced at 10MHz intervals, are shown below. It can be seen that unwanted levels are at > –50dBr and for a QPSK modulation scheme this should not produce noticeable degradation in S/N.



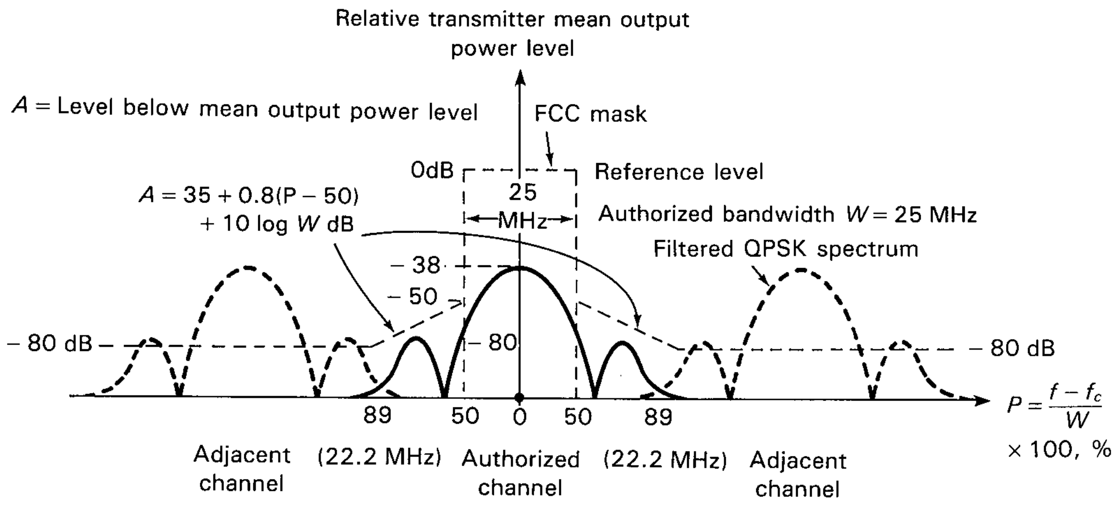
Signal filtering

The spectra of QPSK involve transmitting sidebands as illustrated below.



The sidebands need to be controlled so that adjacent channels will not suffer interference. At the same time we need to receive all the energy being transmitted.

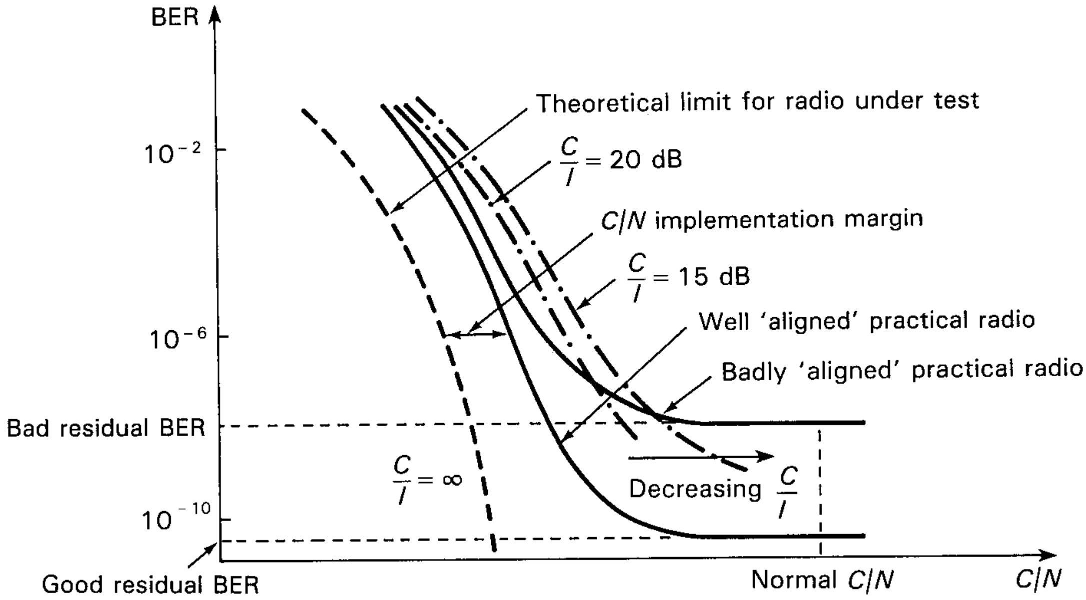
The bandpass filter thus requires a compromise. This is illustrated below including the location of the high and low adjacent channels, which can be seen overlap the sidebands. The required FCC mask (similar top CCIR) is described mathematically in the figure.



Thus the signal filter at 140MHz needs to be matched to the transmitted waveform and to have linear phase and amplitude across the bandwidth.

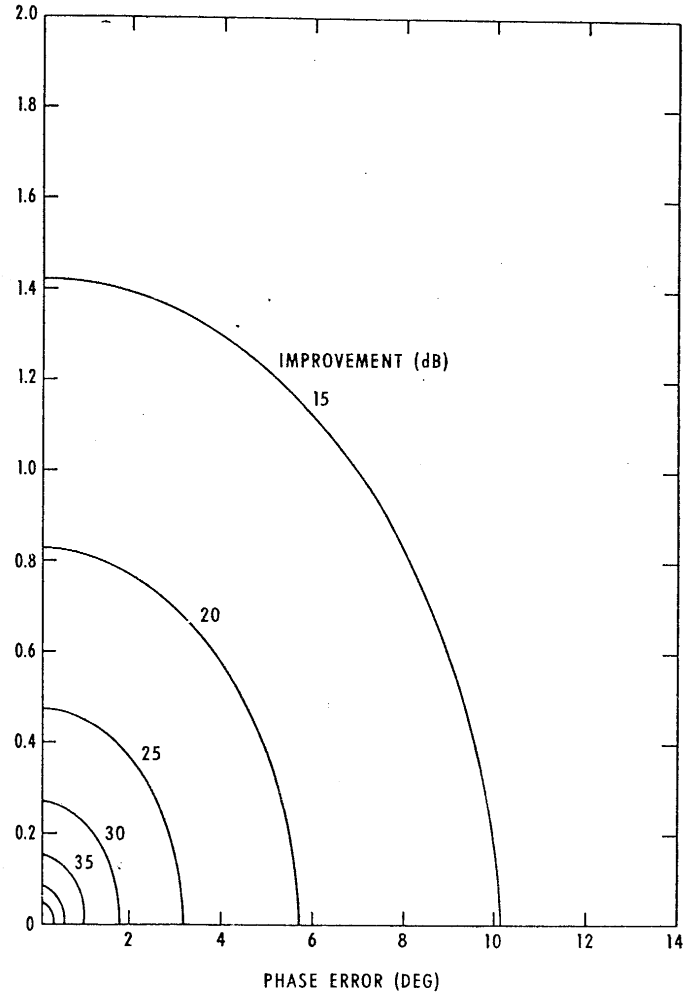
Allowable non linearity for QPSK

The use of digital modulation places stringent requirements on the amplitude and phase linearity required in a transmitter and a receiver. This is illustrated in the graph below.



Poor phase and amplitude linearity results in a raised residual bit error rate (BER). Receiver and transmitter imperfections result in a required increase of C/N for a given BER.

For example the C/N ratio relative to the phase and amplitude error are shown in the graph below. E.g. for a 15dB C/N we require >1.4dB amplitude error and 10 degree phase error. More realistically we require >30dB and as can be seen from the graph this requires errors in phase <1.8 degrees and in amplitude <1.8dB.



We need to consider the bit error rate (BER) probability for various digital modulation schemes. These are shown in the following graph.

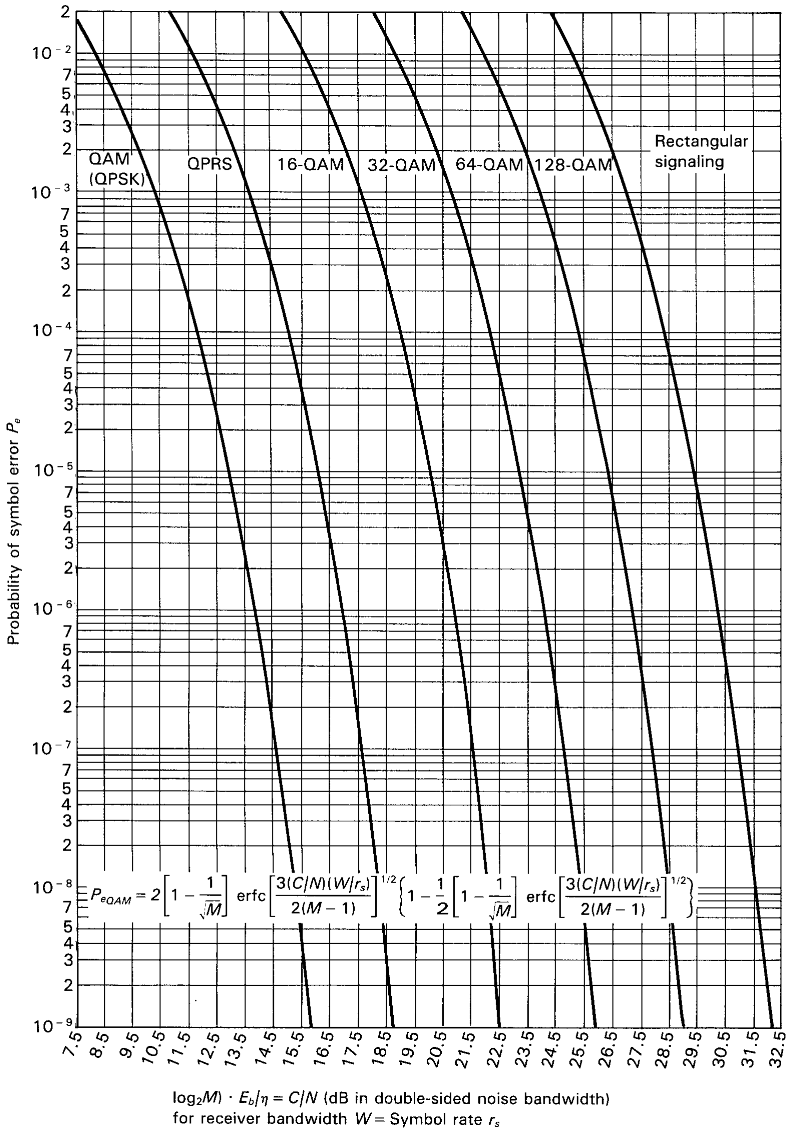


Figure 6.28 Probability of error versus double-sided noise C/N for QAM and QPRS systems

The Airstar system will use QPSK for E1 and E2 but for E3 at 34Mbps we need to use the spectrally more efficient 16-QAM. As can be seen from the graph QPSK and 16-QAM require 14dB and 21dB respectively for a 10^{-6} BER. These are theoretical values and in practice require greater margins to achieve the required BER.

A comparison between various modulation schemes is shown in the table below. Note that spectral shaping on transmit, for example using root raised cosine filters, allows improved spectral efficiency. The inverse process, however, needs to be employed on the receiver.

Modulator efficiency	Modulation scheme	Efficiency	Ideal	C/N	Section	Signaling
0.62	MSK	1.8	2	10.5	6.4.6	Raised cosine($\alpha = 0.25$)
0.59	Offset QPSK/SQPSK	1.8	2	13.5	6.3.7	Raised cosine($\alpha = 0.25$)
0.59	QPSK	1.8	2	13.8	6.3.3	Raised cosine($\alpha = 0.25$)
0.59	8–PSK	2.7	3	19.0	6.3.3.3	Raised cosine($\alpha = 0.25$)
0.59	16–SQAM/16–offset–QASK	3.6	4	20.9	6.5.5	Raised cosine($\alpha = 0.25$)
0.52	MSK	2.1	2	12.8	6.4.6	Partial Response
0.64	Offset QPSK/SQPSK	2.1	2	15.8	6.6.1	Partial Response
0.64	QPSK	2.1	2	15.9	6.6.1	Partial Response
0.64	16–SQAM	4.2	4	22.8	6.6.1	Partial Response