Basic Equipment Design Tutorial

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1 ABSTRACT

The aim of this tutorial is to describe the typical system requirements for the design of RF equipment and to be able to understand each design drivers/parameter that may be specified. Each design driver/parameter is then explained with worked examples given where applicable, verified using an ADS schematic/simulation. This tutorial is based on a lecture given by a now retired engineer, D.E Ellis MBE – a.k.a Jean-Luc Picard!

2 INTRODUCTION

For any given piece of RF/Microwave hardware there has to be a specification, which defines the electrical, mechanical and environmental performance of the equipment. **Table 1** highlights the typical requirements of an equipment.

Specification	Requirement
Equipment Definition	Eg Interfaces, Connectors
Performance Characteristics	Electrical, RF/Microwave performance
Physical Characteristics	Mechanical requirements eg Mass, maximum size
Operability	Reliabilty, Environmental conditions
Design & Construction	Mechanical & Electrical design constraints, eg EMC, bonding, Mechanical Stress

Table 1 Typical requirements given in an equipment specification.

3 DESIGN DRIVERS

The typical key design drivers are listed in **Table 2**. Whether all or some of the drivers are defined depends on the equipment, for example and amplifier would not have a frequency conversion specification but a PLL synthesizer would. A more detailed description of each section of a typical specification in relation to the design drivers will now be given.

3.1 DESCRIPTION

This section of the specification, gives a brief descriptive overview of the equipment, including any redundancy issues.

Basic Functionality
Redundancy Needs
How Much gain
O/P Power and/or Linearity
Noise Figure
Frequency Conversions
Spurious Signals
Mass
DC Power
Temperature
Vibration
Radiation, Life-Time Drift
Schedule
Cost

Table 2 Typical design drivers.

3.2 EQUIPMENT DIAGRAMS

The equipment diagrams give a rough overview of the of the equipment broken down into functional blocks together with the approximate maximum allowed dimensions and mass.

Functional Block diagram –Show the breakdown of the equipment into it's constituent parts eg amplifiers, filters, mixers, oscillators etc.

Initial Interface-Control drawings – Diagram showing the positions of the RF, DC and control connectors. Also shown are the maximum dimensions & mass allowed for the completed equipment.

Connector/pin assignments – Referring to the connectors described in the Initial Interface-Control drawing, each connector pin is designated with it's function.

Ground and return line diagram – For most equipments there will be some form of analogue and/or digital control, and therefore careful consideration of the signal returns need to be considered eg chassis returns and/or cable returns.

3.3 INTERFACE DEFINITIONS

There are normally several interfaces on RF equipments, to allow connection to other equipments and to provide electrical power to active equipments. The interface definitions can be sub-divided into several categories:

3.3.1 Electrical Interfaces

A summary of the key electrical interfaces, are given below:

- Primary D.C power
- Secondary D.C power
- Command
- Telemetry
- RF
- Signal

Clearly, all of these definitions directly impact on the equipment design both source and load definitions. It is probably true that a large proportion of equipment hardware queries are related to these interfaces, so it is worth making sure that these are **clearly defined** and **understood** by all parties!

3.3.2 Mechanical Interfaces

In any RF system there may be many equipments that are connected together and therefore there will be a particular connector standard used for the cabling. In many cases it will be important to have connectors in particular positions, so that for example, when using RF waveguide, these can be manufactured to a particular shape. The position and type of connectors used are defined on an Interface Control Drawing (ICD). Also included on this drawing may be the mass, center of gravity, fixing positions and approximate dimensions. Note that this drawing will obviously evolve as the design progresses.

3.3.3 Thermal Interfaces

The equipment will be required to operate over a particular temperature range eg –20ºC to 75ºC for space hardware applications, and with this in mind consideration has to made with regard to the thermal dissipation of the unit ie will the internal components over heat if there is a poor thermal path through the equipment to the rest of the system. The Thermal requirements to consider are the following:

- Temperature Ranges
- Total power dissipation
- Thermal finish ie black paint to dissipate heat
- Thermal flux density (W/cm^2)

Also a requirement for space hardware is for the equipment to survive cold storage & cold switch on (typically as low as -40ºC).
With such extremes of temperature this will effect the choice of components, processes and the thermal compatability between materials (ie will different thermal expansions of the materials used cause failures?

3.3.4 Performance Characteristics

This section is the most important in influencing the functional electrical design of the equipment. It does not follow that there is a rigid format because the requirements will vary from equipment to equipment. The most typical performance characteristics are shown below:

- Frequency range
- Gain and gain/phase ripple
- Out of band response/rejection
- Amplitude linearity
- Output Power
- ALC Output power
- Noise Figure
- Return Loss
- Spurious signals
- Conducted susceptibility (EMC)
- DC Power

4 PERFORMANCE CHARACTERISTICS

As section 3.3.4 is the most important in influencing the functional electrical design of the equipment, each parameter will be discussed in turn.

4.1 FREQUENCY RANGE

The frequency range of an equipment usually influences the equipment size, technology and component type. At low frequencies there is a tendency to use lumped components on fibre-glass PCB's, bipolar devices for amplifiers and/or oscillators (due to their lower flicker noise) and discrete package parts.
At higher microwave frequencies circuits tend to built on Alumina substrates thin-film (one layer) or thick-film (multi-layer). Alumina (a ceramic material) has considerably less RF Loss than that of fibre-glass PCB, (Eg FR4) at frequencies above approximately 1-2GHz The active components used are GaAs – HEMP Fet's (as they have much greater Ft's) and /or GaAs MMIC die. Leaded components are not suitable for high frequencies (due to the added inductance of the component leads) therefore, surface mount components are thus widely used.
A more detailed discussion of technologies is discussed later on.

4.2 GAIN AND GAIN/PHASE RIPPLE

The gain specification largely influences the active component count and device type. At higher frequencies devices have less gain and so more may be needed to meet a particular gain specification, which will invaria-

bly increase the DC power consumption. And possibly increase the size of the equipment.
Special consideration should be made to the following conditions:

- Gain Stability
- Maximum tolerable gain block size.
- Conducted susceptibility (Supply line feed-back)
- Mis-match effects.
- Multipath effects.

4.3 DEFINITION OF GAIN RIPPLE

Unless there is an operational need to separate the ripple components, then use the following specification format:

The gain variation with frequency between f1 and f2 shall not exceed "X" dB peak-to-peak.

If any more complicated definition is used, then test confusion can easily arise (It is difficult to separate slope and ripple unambiguously).

However, in certain situations for example systems using large attenuations between gain blocks, there may be both gain slope and gain ripple. In such situations the definition is defined as shown in **Figure 1**.

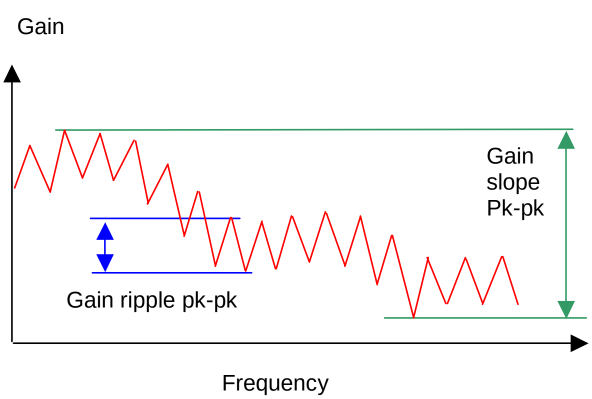


Figure 1 Definition of gain ripple and gain slope where gain ripple is measurable, normally gain slope and gain ripple are used interchangeably.

4.4 MAXIMUM TOLERABLE GAIN BLOCK SIZE

There are no "well defined" gain limits, but unwanted feedback effects can induce gain & phase ripple together with gain instability. Gain ripple due to feedback can best be explained with an example shown in **Figure 2**.

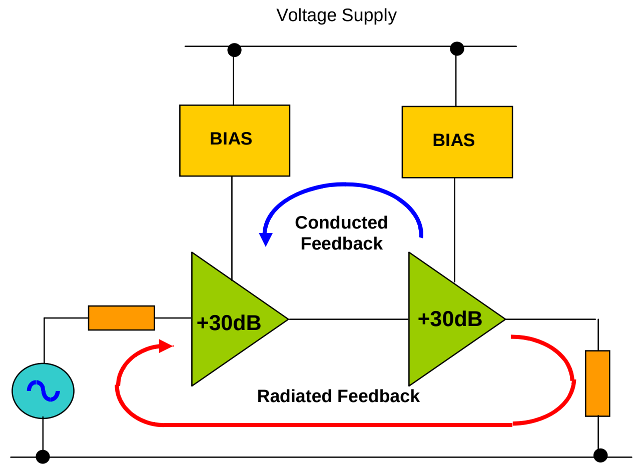


Figure 2 Feedback paths that can degrade the gain ripple of a system.

From **Figure 2** we can see that the overall gain of the system is 60dB, however with such high gain there we be signal leakage though the DC bias networks and onto the supply ie '**conductive**' feedback. In addition signals may leak back to the input due to cavity resonance effects and reflections of walls/lids and this is known as the radiated feedback path.
For our example we will assume that the overall feedback attenuation is 70dB, thus giving us a loop gain of 60-70 = **-10dB**. This figure although probably not large enough to give feedback oscillations with increase the gain & phase ripples by introducing the loop gain phasor to the main path as shown in **Figure 3**.

Our resulting combined phasor can lie anywhere on the circle thus giving a maximum potential phase ripple of ±17.5º.

Our gain uncertainty will be **20log(1±0.316)** = +2.4dB to –3.3dB. This may well show up as a lid effect.

And the potential gain ripple will be (2.4-(-3.3) = 5.7dB peak-to-peak. However, this depends on the loop phase change within the system bandwidth.

To minimize these potential problems due to radiated isolation, we can be add screening walls between gain blocks.

NOTE! With a perfectly screened box you will get potentially 100dB of isolation, however in the real world where substrates are assembled into a cavity etc, etc a more typical isolation figure of 70dB is possible. To eliminate lid effects, firstly ensure that some form of cavity resonance simulation/prediction is performed and secondly, if necessary radio absorptive material (RAM) can be used to greatly reduce lid effects (Ensure that RAM material selected, is suitable for the frequency of the equipment).

To reduce the effects of conducted feedback we can add attenuation between the gain blocks and ensure that there is adequate RF filtering (by using filter-cons/capacitive feedthru's) as part of the DC bias networks.

$$-10\text{dB} \Rightarrow \text{Convert to power}$$
$$= 10^{\frac{-10}{20}} = 0.316$$

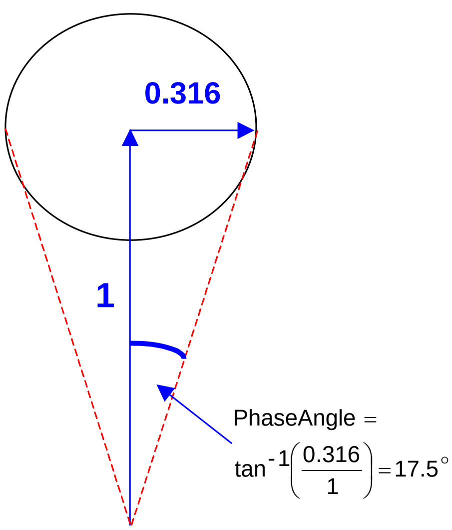


Figure 3 Resulting signal phasor due to the introduction of the feedback loop gain phasor of magnitude –10dB (which equates to a power of 0.316)

The previous example can be run on ADS [2] using the S-parameter schematic shown in **Figure 4**. In this simulation a electrical delay of 1uS has been added to give the circuit some frequency dependence. This is effectively giving us the phase error, to cause the resulting phasor to move around the circle of **Figure 3** with frequency.

The resulting simulation plot is shown in **Figure 5**. The resulting phase plot would consist of a 'saw-tooth' response between ±180 degrees. If however, we remove the 1uS electrical delay (with an electrical delay of –1uS at the output termination) then we will flatten out the phase response to reveal the phase ripple as shown in **Figure 6**.

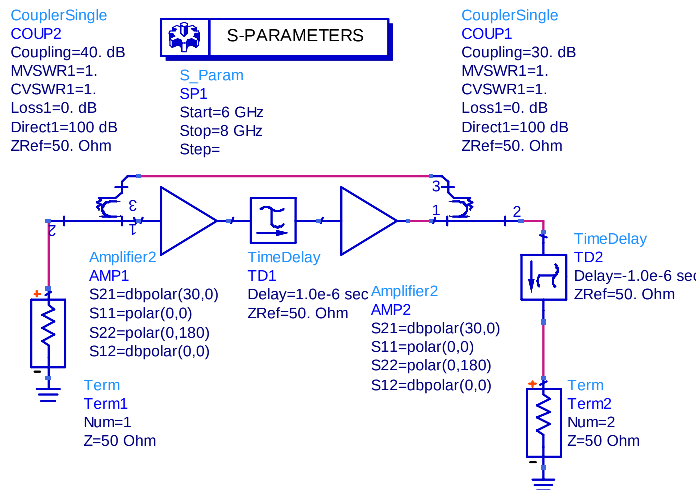


Figure 4 ADS S-parameter schematic of the example amplifier chain shown in Figure 2. The signal is fed back from the output of the second amplifier, to the first amplifier input using two couplers of total coupling factor of 70dB. The electrical delay is to give the circuit some frequency dependance and is arbitrarily set at 1uS.

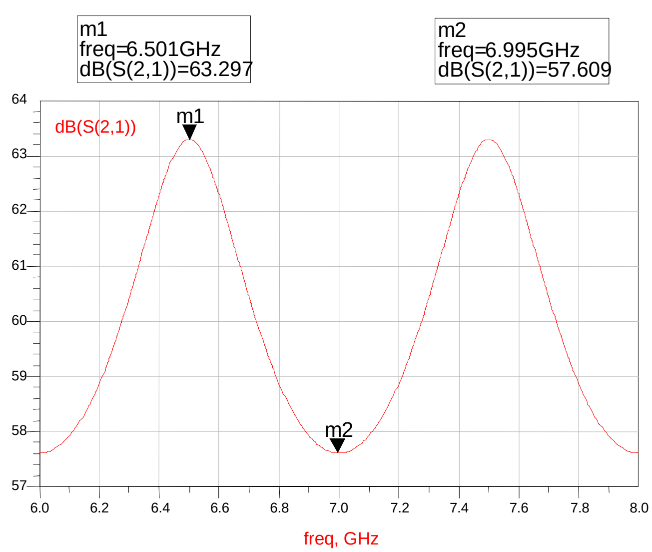


Figure 5 Resulting gain plot from the ADS S-parameter simulation shown in Figure 4, showing the

predicted 5.7dB peak-to-peak gain ripple response due to the leaked back signal.

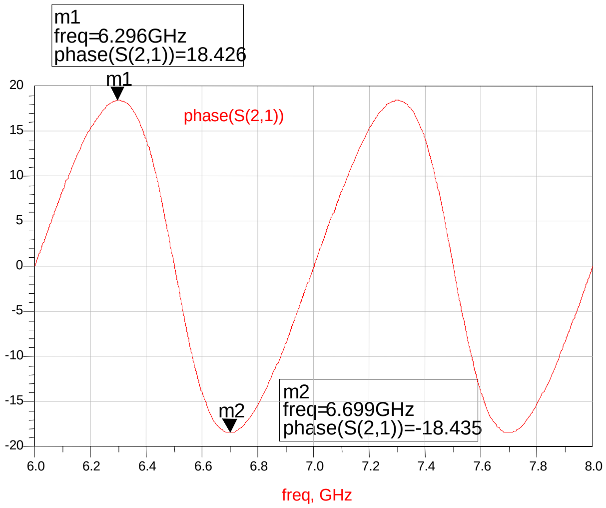


Figure 6 Resulting phase vs frequency plot from the ADS S-parameter simulation shown in Figure 4, showing the predicted ~35 degree peak-to-peak phase ripple response due to the leaked back signal.

4.5 MAXIMUM TOLERABLE ATTENUATION BLOCK SIZE

When using large attenuators it is possible for the signal to leak to the output by a feedback radiation path that effectively limits the attenuation possible. To illustrate this effect, an example is shown in **Figure 7**.

Using the same hand calculations of section 4.4, the resulting phasor will cause the potential phase ripple to be **20 degrees peak-to-peak,** with a gain uncertainty of **+1.4dB to –1.7dB**.

If the attenuator is variable, then the phase will change depending on the value of the attenuator, which may be undesirable in phase sensitive systems.

The gain uncertainty may well limit the practical attenuation per stage. Hence large RF systems requiring a high attenuation range may require more stages or use a variable gain configuration.

Finally, the passband slope will increase with increase attenuation.

To simulate the effect of feedforward radiation the ADS simulation of **Figure 4** can be modified to replace the amplifiers with attenuators and reversing the couplers as shown in **Figure 8**.

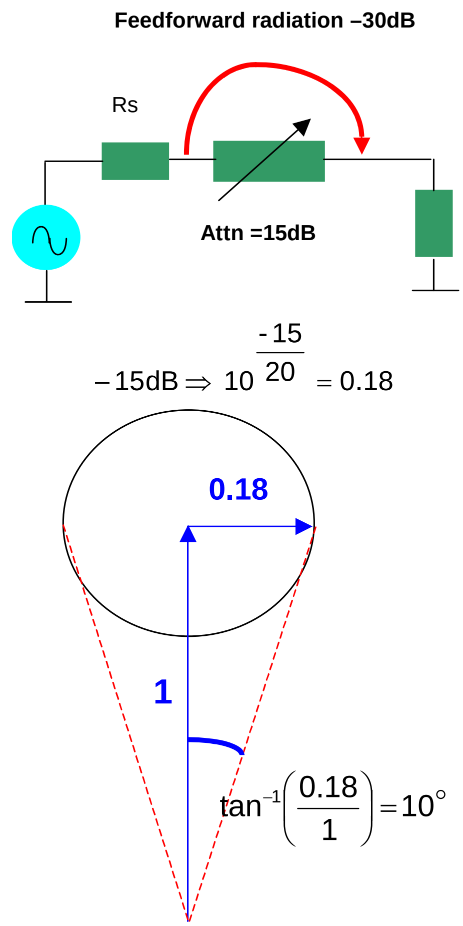


Figure 7 Resulting phasor from feedforward radiation across the attenuator that will increase the gain/phase ripple.

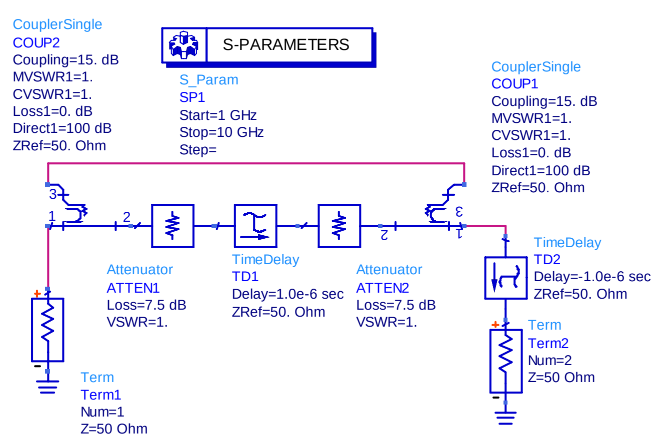


Figure 8 ADS schematic to simulate the effects of feed-forward radiation in attenuator blocks.

4.6 MISMATCH EFFECTS

In ideal 50-ohm functional blocks all the applied power is absorbed in the block and no signal is reflected back to the source. Unfortunately, the functional blocks have finite return losses and thus will reflect some of the incident power back to the source. The sources will also have a finite return loss and will reflect and received signal from the load. These ‘reflected’ waves will travel backwards and forwards between the source and load interacting with each other over the passband, to form peaks and troughs in the frequency response ie standing waves.

The amplifier arrangement shown in is used as an example is shown in **Figure 9**. Here we have two amplifiers connected by a length of coaxial cable. The return losses (output of first amplifier & input of second amplifier) are 16dB.

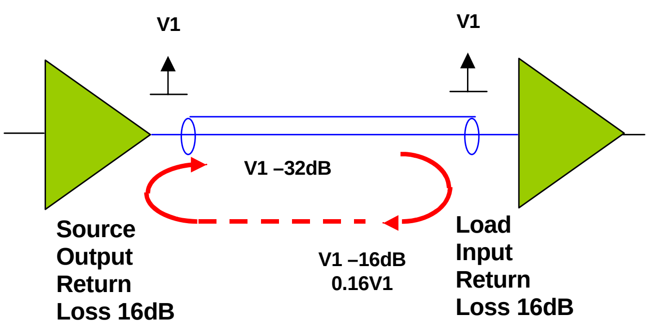


Figure 9 Example circuit to show the effects of finite return losses on the gain uncertainty or gain/phase ripple.

If we assume negligible transmission loss in the coaxial cable then:

The reflected signal back from Amplifier 2 will be

$V_1 - 16\text{dB}$

(if v1 = 0dBm then reflected signal will be –16dBm.

Now this signal will get reflected back (ie double reflected) by amplifier 1 reduced by another 16dB ie

$V_1 - 16\text{dB}-16\text{dB}$

(if v1 = 0dBm then doubled reflected signal will be –32dBm.

Using our phasor calculation results in an additional vector of magnitude:

$$-32\text{dB} \Rightarrow 10^{\frac{-32}{20}} = 0.025$$

..resulting in a gain uncertainty of:

$$20\log(1 \pm 0.025) = \pm 0.21\text{dB}$$

with a potential gain ripple of 2 x 0.21dB = 0.42dB pk-pk.

Phase ripple will be

$$\pm \tan^{-1}\left(\frac{0.025}{1}\right) = \pm 1.43^\circ = 2.86^\circ \text{pk} - \text{pk}$$

Note: The full gain and phase ripple is only encountered if the loop phase shift encompasses the in-phase and anti-phase condition, within the required system bandwidth. This phase shift can come from the interface source and load as well as the interface line length.

This example can now be verified by running the S-parameter simulation shown in **Figure 10** with the corresponding gain & phase ripple plots shown in

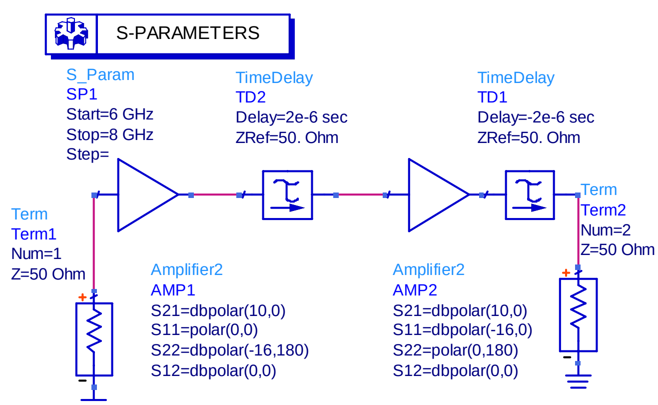


Figure 10 S-Parameter ADS schematic to simulate effect of finite return losses. The first amplifier has been set up to have an output return loss of –16dB, with the second amplifier input return loss also set to –16dB. The delay of 2us is inserted to represent coaxial cables, micro-strip etc joining the two amplifiers. The second delay of –2us takes out the electrical delay to leave the phase pk-pk response after simulation.

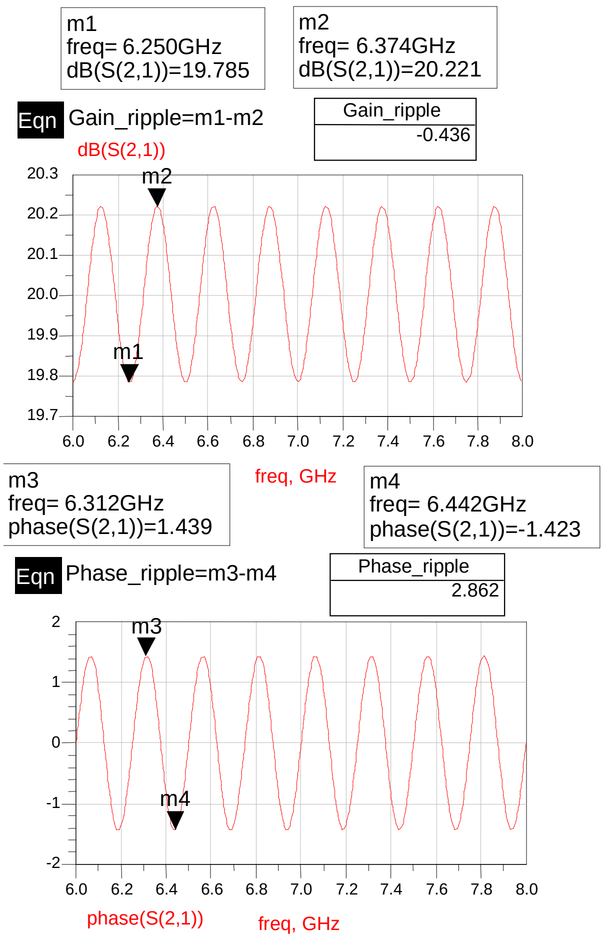


Figure 11 Resulting gain & phase ripple plots from the ADS simulation shown in Figure 10. The plots show agreement with the hand calculations presented in the example.

4.7 MULTI-PATH EFFECTS

The Multi-path effect is best illustrated by the following example:

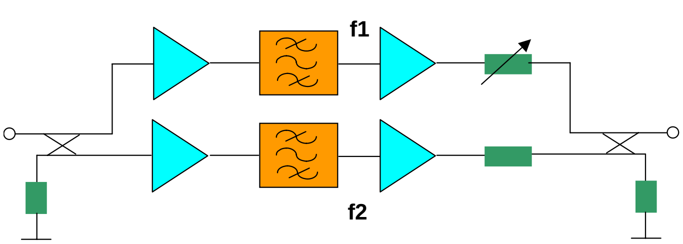


Figure 12 Schematic to illustrate the multi-path effect.

In this system the input signal is split between two (usually more than two) paths, each having it's own band-pass filter set to different frequencies as the system is channelised as per a typical satellite transponder input section.

If we pass a signal f1 through the system it will pass straight through path 1 as this chain filter is set to f1. Signal f1 will also path through chain 2 but this has a filter at a different frequency and will attenuate the signal by it's stop-band attenuation. Therefore, f1 will appear at the output twice – one from chain 1 and an attenuated signal from chain2. If we assume a stop-band filter attenuation of say 50dB we will have the frequency response shown in **Figure 13**.

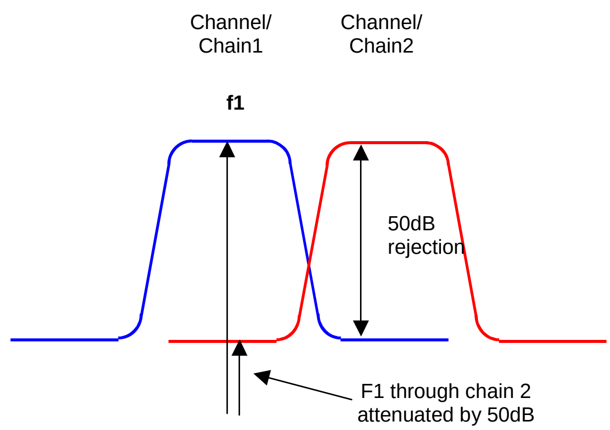


Figure 13 Frequency response of the two-channel system shown in Figure 12. The signal f1 is centred on channel/chain 1 and will pass through to the output largely unaffected by the chain 1 channel filter. However, f1 passing through chain 2 will hit the stop-band of chain 2 channel filter and will attenuate the signal by 50dBc (assuming that this is the stop-band rejection for the filter). This f1 attenuated signal together with the larger unaffected f1 signal will combine at the output of the system resulting in increased gain & phase ripple.

Assuming the output attenuators are set the same then the gain ripple will be:

$$-50\text{dB} \Rightarrow 10^{\frac{-50}{20}} = 3.162E^{-3}$$

..resulting in a gain uncertainty of:

$$20\log\left(1 \pm 3.162E^{-3}\right) = \pm 0.027\text{dB}$$

with a potential gain ripple of 2 x 0.027dB = 0.055dB pk-pk.

Similarly the Phase ripple will be:

$$\pm\tan^{-1}\left(\frac{3.162E^{-3}}{1}\right) = \pm 0.181^{\circ} = 0.36^{\circ}\,\text{pk}-\text{pk}$$

If however the chain 1 attenuator was increased by 20dB (So that the signal f1 was reduced by 20dB) then the gain and phase ripple will increase:

$$-50\text{dB} \Rightarrow 10^{\frac{-30}{20}} = 0.0316$$

..resulting in a gain uncertainty of:

$$20\log(1 \pm 0.0316) \; = \; \pm 0.27\text{dB}$$

with a potential gain ripple of 2 x 0.27dB = 0.55dB pk-pk.

And the new Phase ripple will be:

$$\pm\tan^{-1}\left(\frac{0.0316}{1}\right) = \pm 1.81^{\circ} = 3.62^{\circ}\,\text{pk}-\text{pk}$$

The ADS schematic of the multi-path effect is shown in Figure 14. The gain ripple plot of the circuit is shown in Figure 15. The attenuator ATTEN1 was then set to 20dB and the simulation re-run resulting in the gain ripple plot of Figure 9

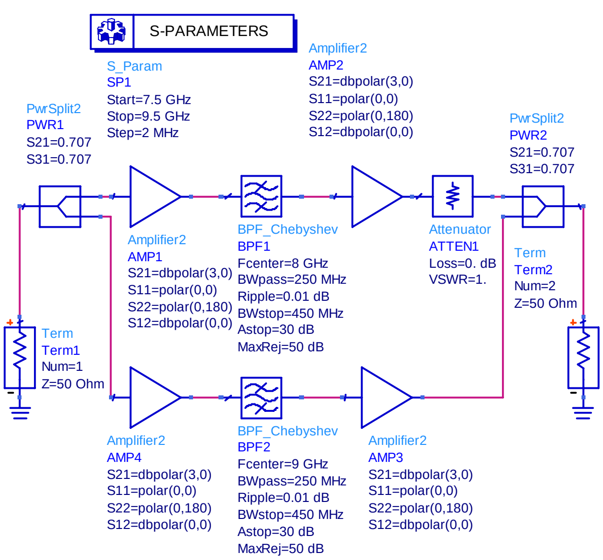


Figure 14 ADS schematic to verify the multi-path hand calculations performed in the example. The filters have been set with a maximum attenuation of 50dB (Max rej).

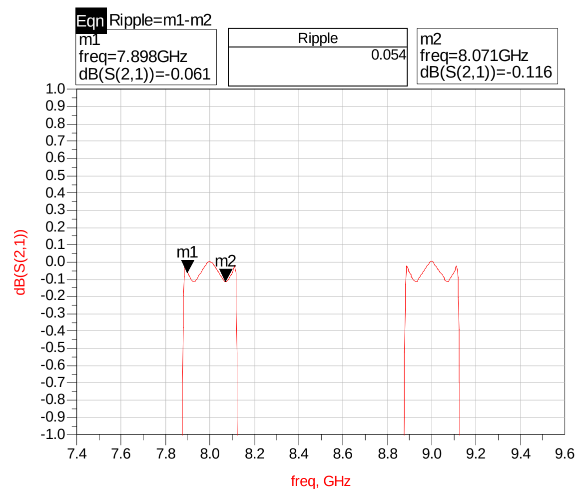


Figure 15 Simulation result of the multi-path ADS schematic of Figure 14, verifying a gain ripple of ~0.055pk-pk.

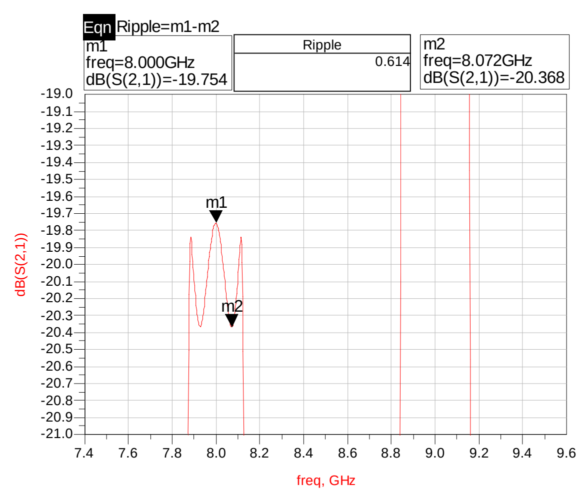


Figure 16 Simulation result of the multi-path ADS schematic of Figure 14, with attenuator ATTEN1 set to 20dB in the wanted signal path. The plot verifies that the attenuator will degrade the gain ripple to ~0.55pk-pk.

4.8 OUT-OF-BAND RESPONSE/REJECTION

This design parameter is dominated by the system filters. However, it is important to ensure that amplifier designs do not give rise to gain peaks at lower or higher frequencies. (This is important for stability considerations). The positioning of the filter within the gain chain has to be a best compromise considering the following:

- Rejection of unwanted input signals at the front end to avoid the need for excessive power handling capability of subsequent stages.
- Out-of-Band noise generated in the post filter stages. There may be a practical limit to the post filter system output noise density.

4.9 AMPLITUDE LINEARITY/OUTPUT POWER [4]

As an RF system is driven harder from an RF source, then the increased RF power will gradually change the gain characteristic from linear operation (for small signals) to gain compression and eventually limiting. This gain compression gives rise to non-linearity.

Many equipments, have to deliver a specified output power (with a defined linearity performance) and meet the overall DC to RF efficiency or specified DC power limit.

This however, is not often a major problem for a single carrier system where essentially, the gain reduces and the efficiency can increase.

However, for multi-carrier arrangements, the non-linearity gives rise to unwanted inter-modulation products (which act as interference signals to the system).

4.9.1 How can amplitude Linearity be defined?

An obvious method is to define *gain compression.* This is sometimes used as an indication of performance for single carrier conditions, but it's measurement and "unambigious" interpretation are difficult. A typical gain compression level often used for devices is **1dB**.

The *Two Carrier* performance is usually defined by specifying the level of the third order inter-modulation products (relative to each carrier), defined at a total output power.

Noise Power Ratio tests are sometimes used to establish a more "traffic realistic" performance evaluation.

Referring to **Figure 17**, a uniform noise density, with a narrow stop-band notch is applied. The corresponding output notch for a linear system would not contain any noise. Any Non-linearity's will result in noise products appearing in the output notch band.

This level relative to the un-notched frequency band is a measure of the non-linearity **(NPR).**

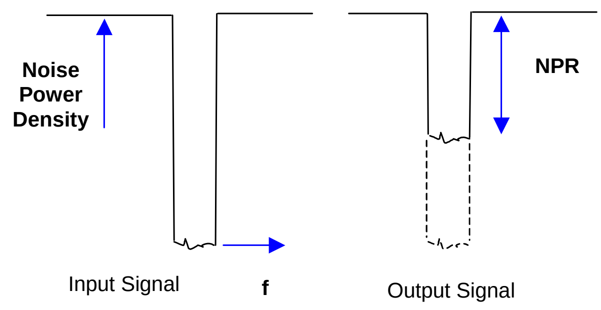


Figure 17 Illustration of the *Noise Power Ratio* test, sometimes used to establish a more "traffic realistic" performance evaluation, resulting in the measurement of non-linearity (NPR).

4.9.2 Third Order Intermodulation Product Characteristics (For 2 equal carriers)

When two equal RF carriers are supplied to an active device it's non-linearity's will give rise to intermodulation products separated by the difference of the two main carriers as shown in **Figure 18**.

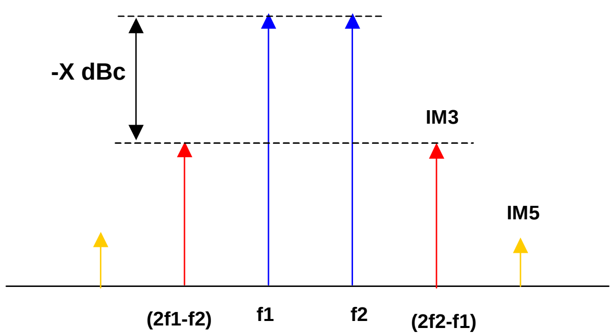


Figure 18 Resulting carrier spectrum when two equal signals cause non-linearity in an active device. –X is the third order modulation product relative to the carrier in dBc.

We can measure the gain compression and 3rd order intermodulation products on say an amplifier by firstly performing a Pin vs Pout measurement. This will determine the output power where the gain has compressed by 1dB (the normal definition of gain compression). An example of gain compression is shown by the red curve in **Figure 19**..

The intermodulation measurement requires two equal carriers to varied in power whilst measuring the difference in power between **one** of the fundamental carriers and **one** of the 3rd order intermodulation products.

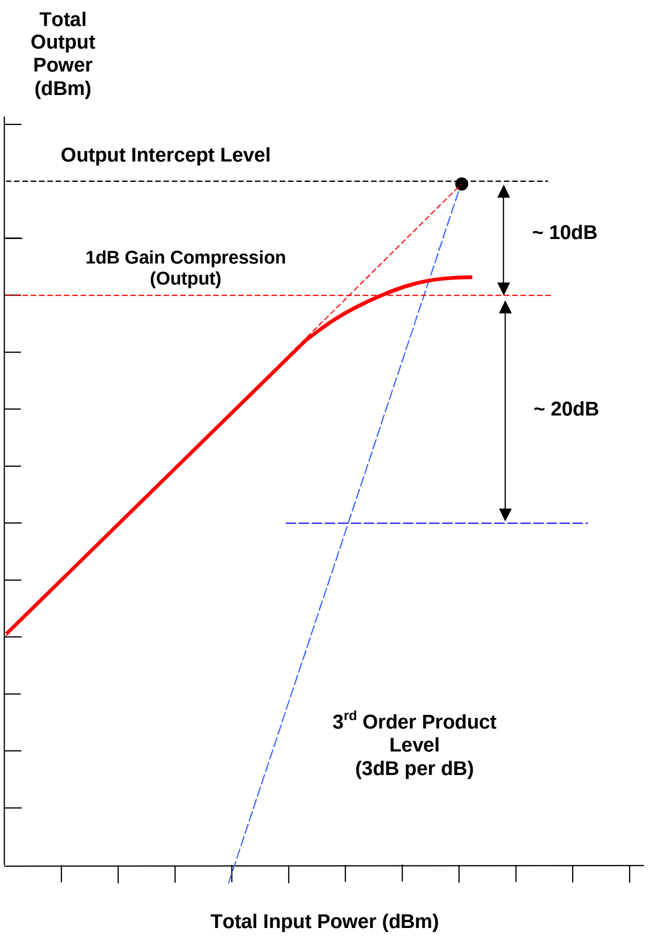


Figure 19 Plots of gain compression (red curve) and 3rd order intermodulation level (blue line) for a typical amplifier.

The third order intermodulation products will occur at (2.f1-f2) and (2.f2-f1). For a typical arrangement:

- The Output Intercept point is about 10dB above the 1dB gain compression point.
- The 3rd Order Intermodulation ratio (IM3) is about 20dBc at the gain compression output level.

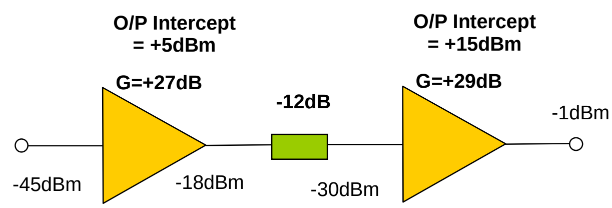
4.9.3 Linearity Assessment of Cascaded Stages

The first part of the assessment involves establishing the 3rd order Intermodulation charactersitics for each stage. This can be done by studying the data sheet, CAD simulation or by test if possible. For simple calculations use the approximation already outlined.
Analyse the cumulative gains and stage output powers. From each stage 3rd order IM characteristic, deduce the intermodulation ratio (dBc), using the formula:

-(O/P Intercept power level – (Stage O/P Power) x 2

Then vector add (ie convert dB's to powers, add then reconvert back to dB's again) the intermodulation components for each stage, to give an overall performance.

An an example system is shown in Figure 20.



IM3 = -[5-(-18)]x2 = -46dBc   IM3 = -[15-(-1)]x2 = -32dBc

Convert to linear power: Convert to linear power:

$-46dBc \Rightarrow 10^{\frac{-46}{20}}$ $-32dBc \Rightarrow 10^{\frac{-32}{20}}$

Add power vectors and convert back to dB's:

$$= 20\log\left[\left(10^{\frac{-46}{20}}\right)+\left(10^{\frac{-32}{20}}\right)\right] = -30.42dBc$$

Figure 20 Example system used to show how to calculate the overall 3rd Order Intermodulation ratio.

4.9.4 Automatic Level Controlled Output Power (ALC)

Some systems require the output power level to be maintained constant as the input signal level is increased.

This can be achieved by controlling an RF/Microwave attenuator by a feedback control loop as shown in Figure 21.

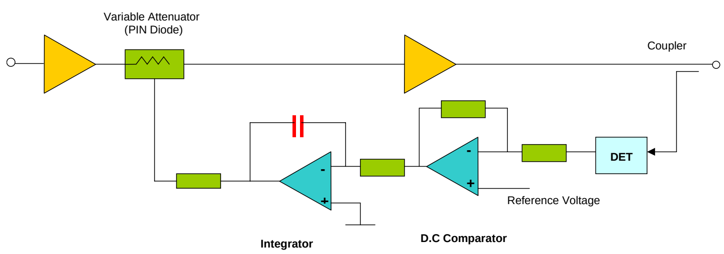


Figure 21 Typical ALC arrangement

ALC loops are sometimes poorly designed. The following points should therefore be considered:

- Ensure that the detector input RF power is sufficient to minimize the DC drift effects of the detector, comparator and integrator. The DC resistors tolerance & drift, reference & supply voltage variations etc can all be significant.
- Imperfect tracking of the detector diode and compensating diode over temperature, life and radiation environment will give errors. Life and Radiation drift per diode can be around 2mV to 5mV each. Hence the differential can be as high as 4mZ to 10mV.
- The detector characteristic changes between single carrier and multi-carrier operation when working at higher levels.
- Ensure loop stability by making the integrator time constant dominant.
- Ensure loop stability by making the integrator time constant dominant.
- Be sure to “current drive” the PIN diode attenuator if used.

- Check that the loop response time satisfies the customer needs.
- Beware of output transient overshoot until the loop takes control.

4.10 NOISE FIGURE [3][4]

Some of the most common definitions are given below:

Noise Factor (linear ratio F) = $\dfrac{\text{S/N In}}{\text{S/N Out}}$

Noise Figure = 10 $\log_{10}$ (Noise Factor)

Reference Temperature (To) = 290K

Equivalent Noise Temperature (Te) = (F-1)To

Attenuator Effective Noise Temperature = (X-1)To
–
Where linear attn X >1 and To is absolute temperature

Attenuator Noise Factor = $\dfrac{1+(x-1)T}{To}$

Noise Factor = $\dfrac{(Te+To)}{To}$

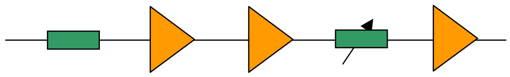
Overall Cascaded Noise Factor:

$$F_1+\frac{F_2-1}{G_1}+\frac{F_3-1}{G_1.G_2}+....etc$$

Overall Cascaded Noise Temperature:

$$T_1+\frac{T_1}{G_1}+\frac{T_2}{G_1.G_2}+....etc$$

An example RF chain using these equations is shown in **Figure 22**. The operating temperature is set at 50ºC or 323ºK

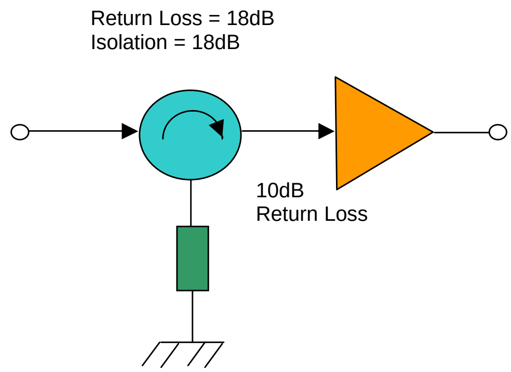


Gain/Loss (dB)	-2dB	+12dB	+13dB	-6dB	+20dB
Linear Gain	1/1.585	x15.849	x19.953	1/4	x 100
Noise Figure		2.5dB	4dB		7dB
Noise Factor		1.778	2.512		5.012
Rel Noise Temp	0.651To	0.778To	1.512To	3.341To	4.012To
Noise temp Refrred to I/P	0.651To	1.233To	0.151To	0.017To	0.08To

Overall Noise Temperature = 2.132 To

Noise Figure = 20 $\log_{10}$ (1 + 2.132) = 4.958dB

Figure 22 Example RF chain to show how cascaded noise temperature is calculated. The operating temperature is set at 50ºC or 323ºK. Note that this style of analysis readily identifies the major contributors to overall noise performance (Note that temperatures are referred to the input).



Relative reflected power –18dB and –28dB
= -15.6dB (worst case after vector multiplication)

Figure 23 Example to show how the input return loss will degrade as a result of a poor return loss added to the output of the isolator, with finite return losses and isolation. The signal from the isolator will be reflected at the amplifier ie –10dB then doubly reflected at the isolator ie –10-(-18) = -28dB.

4.11 RETURN LOSS

Beware that if the RF equipment is to use coaxial connector interfaces, ensure that interface return losses are not over-specified:

- **Typically for SMA connectors:**
  - VSWR is given by 1.04 + 0.009(GHz).
  - Stress relieved connections from the connector pin to the substrate will degrade the return loss by introducing a mismatch.
  - Connector path to substrate ground-plane will degrade the return loss by introducing a mismatch.
- **Typical coaxial isolator specifications:**
  - Return losses = 18dB
  - Isolation = 18dB

These two parameters combine (together with the equipment the interface return loss) as shown in the example of **Figure 23**.

Beware of the effect on return loss in redundant configurations, where one or more paths may be switched off, resulting in one or more poor return losses being presented. This is because if there is an amplifier at the front of the chain and it is switched off, it will have a very poor return loss.

4.12 SPURIOUS SIGNALS

Spurious signals assessment can be one of the most difficult parameters to evaluate. Summary of the types of spurious signal that can occur:

- Unwanted oscillations in active stages.
- Carrier related spurious
  - Signal Harmonics.
  - Intermodulation products (inc mixer products)
  - Conducted Susceptibilty
- Non-carrier related spurious
  - LO Leakage.
  - Cross-coupling from adjacent channel.
  - Signals derived from transponder input interference.
  - Radiated susceptibility.

4.12.1 Unwanted Oscillations in active Stages

Amplifier instability continues to be a major design problem. Specifications usually define maximum allowable spurious levels:

BUT – <u>ANY</u> oscillation from inadequate stability margins is completely unacceptable!!
It must be strictly accounted for and must be an automatic consideration for all designs – even if there is considerable filter rejection located after it.

4.12.2 Spurious Harmonics

A common problem is the harmonic level of the output signal (A relative measurement in dBc).

The usual methods to improve this parameter are:

- Provide output Filtering
  - Beware of the extra insertion loss!
  - Beware of filter spurious passbands (that co-incide with harmonics)
- Make offending stage more linear
  - Minimise any stage output loss that may be affecting linearity.
  - Increase DC power (But is it available? And consider thermal effects)

Note – Harmonics from earlier stages may be a problem. In this case, re-examination of the gain/power distribution may be worthwhile.

4.12.3 Intermodulation Products [4]

Third order intermodulation products have been discussed in section 4.9.2, however higher order products can be a problem as the example in **Figure 24** shows. The amplifier will generate harmonics of the input frequency if driven into compression. These harmonics will 'mix' in the mixer with harmonics of the Local oscillator frequency, that are generated in the mixer as a large LO power is often required to sufficiently switch the mixer.

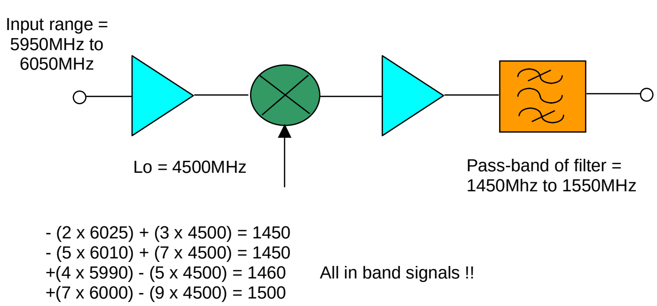


Figure 24 Example showing how intermodulation products in a down converter system can mix to produce products that fall within the pass-band and hence cannot be removed.

The way to deal with this problem is to first decide what performance is acceptable (with margin). Assess unwanted products by using ADS and/or by measuring early breadboards. Study the data sheets of mixers to determine their spurious products data.

The spurious data for a Mini-Circuits[3] SRA-220, 0.05 - 2000 MHz mixer is shown in **Figure 25**.

RF Order Harmonic		0	1	2	3	4	5	6	7	8	9	10
0	-	-	1	16	16	31	52	33	28	68	49	63
1	-	23	-	29	23	30	42	43	38	56	35	48
2	>95	61	59	59	61	66	71	70	72	64	>67	67
3	>94	>72	69	>74	63	>73	72	>71	>74	>74	>74	67
4	>96	75	>73	>73	>74	>74	>74	>74	>74	>74	>73	>75
5	>97	>74	>74	>73	>73	>72	75	>73	>72	>74	>74	>73
6	>90	>68	>74	>74	73	>73	>73	>75	>73	>73	>74	>74
7	>89	>67	>67	>74	>75	>75	74	>74	>74	>75	>73	>73
8	>90	>67	>69	>67	>73	>74	>74	>73	>73	>73	>74	>74
9	>89	>68	>68	>68	>67	>74	>74	>73	>74	>72	>74	>73
10	>91	>68	68	>68	>68	>68	>74	>74	>74	>71	>74	>74

Harmonic LO Order

Test conditions: RF = 999.1 MHz, Drive level = -13.99 dBm
LO = 969.01 MHz, Drive level = 10 dBm
IF = 30.095 MHz, Measured IF level = -21.81 dBm

Figure 25 typical spurious levels for a Mini-Circuits [5] SRA-220, 0.05 - 2000 MHz mixer. Levels of resulting spurious are given in dBc.

When designing systems containing mixers here are a number features to be wary of:

- Mixers are usually specified (by the manufacturer) for 50-ohm broadband terminations. This usually gives the best linearity performance. So in reality actual performance will be worse!!
- An isolator at the LO input is not necessarily the best solution. They are often relatively narrow-band and give poor return loss at the higher order frequencies.
- A low value resistive attenuator can often improve performance (but be careful to ensure that the LO drive is still sufficient).
- Mixer/filter interfaces can be troublesome. Higher products will be reflected back to the mixer at some arbitrary phase, therefore, use a circuit with a broad-band match eg a diplexer.
- Watch out for "fortuitous" nulling at breadboard level that cannot be consistently repeated.

4.13 CONDUCTED SUSCEPTIBILTY

This is an EMC test where signals are injected onto the equipment power supply lines. These signals which, are typically in the range 20Hz to 50MHz with a level of 1V pk-pk, may appear on the RF carrier as a discrete spur (non-carrier related). Additionally if switched mode power supplies are used then it is possible that the switch frequency of the supply (typically a few hundred KHz and harmonics) can cause interference to the RF carrier.

Conducted Susceptibility can be improved in a number of ways:

(1) Provide local circuit ripple rejection

- Dilute the RF amplifier's non-linearity if possible.
- Current bias regulation with good low frequency decoupling for the reference.
- Wide bandwidth for current bias loop.
- Local DC voltage regulation.
- Secondary supply passive filters (eg filtercons or feed thru's). Note that filtercons may be specified in a 50-ohm system and may not give the performance stated by the manufacturer.

(2) Provide sufficient primary to seconday ripple rejection via the DC converter (around 50dB rejection is typical).

- Note that the rejection crossover between active rejection and filter rejection, at around 5KHz, can be a problem.
- There may be a need for extra rejection in the RF circuits at this frequency.
- Ensure that grounding is effective. (There are no general rules for this).

4.14 LO LEAKAGE

For normal operation of mixers at high-level local oscillator RF power is required to ensure mixer switching. As the RF level is high, there will be leakage from the LO port to the RF & IF ports.

A typical up-converter configuration, so show the effect of LO breakthrough is shown in **Figure 26**.

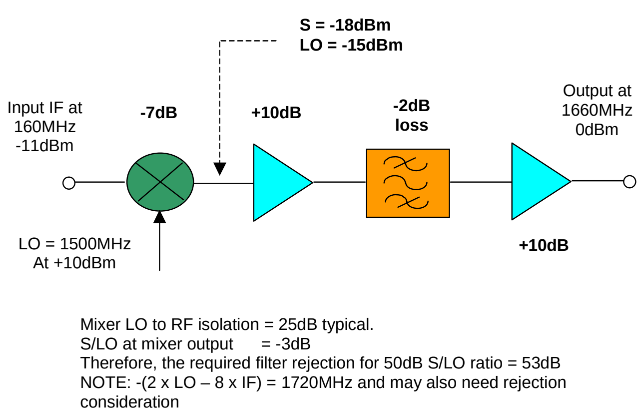


Figure 26 Typical up-converter configuration used to illustrate the effect of LO breakthrough.

We can verify this example by using ADS and a harmonic balance simulator (as we are using a mixer) as shown in **Figure 27**. The simulation was first run with the filter removed to determine the level of LO breakthrough and verify that it is 3dB higher than the wanted signal – this plot is shown in **Figure 28**. The second simulation was performed with a filter inserted with a stop-band rejection figure of 53dB. The resulting plot with the filter added is shown in **Figure 29** and shows that the LO breakthrough is now 50dB down on the wanted signal (ie S = +2dBm and LO = -48dBm).

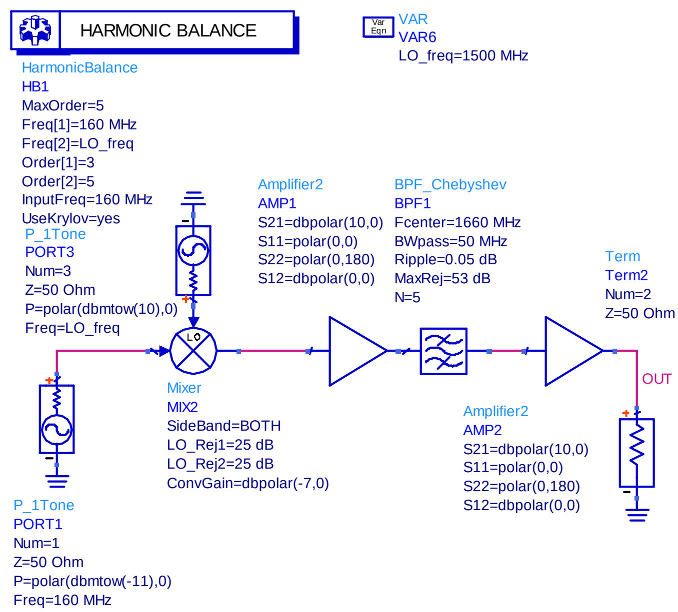


Figure 27 Harmonic balance simulator to examine the LO leakage from a typical mixer. The RF and LO frequencies have to be specified in the harmonic balance simulator. The mixer is set to up-convert AND down-convert – Side Band BOTH, with a LO_Rej2 (LO to Output Port) rejection of 25dB. Two simulations have been performed one without the filter to show the level of LO breakthrough and the second simulation with a filter added to give 50dB LO rejection.

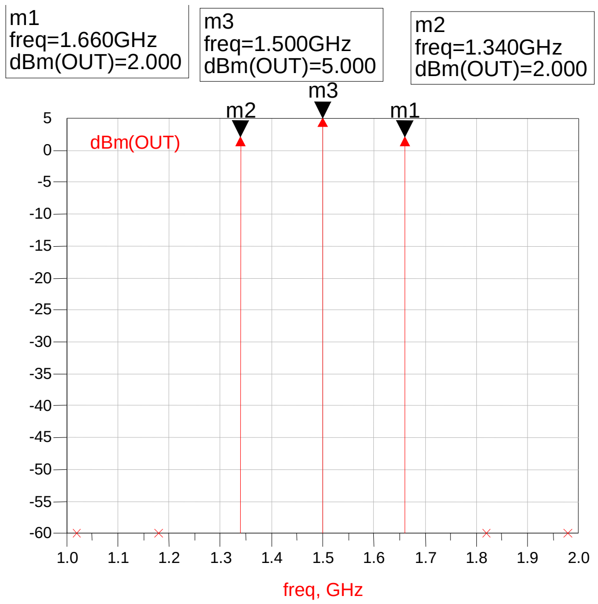


Figure 28 RF spectrum (dBm) of the simulation of Figure 27, with the LO rejection filter omitted, resulting in a S/LO ratio (Marker 3/Marker 1) of only 3dB, as per the example

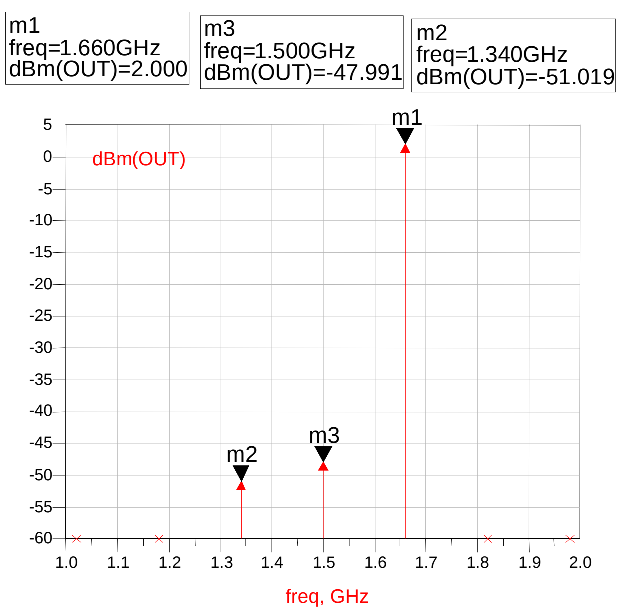


Figure 29 RF spectrum (dBm) of the simulation of Figure 27, with the LO rejection filter fitted and designed to have a stop-band rejection of 53dB, resulting in a S/LO ratio of 50dB, as per the example.

4.15 CROSS-COUPLING FROM ADJACENT CHANNEL

Whenever channels in RF equipments are multiplexed as shown in the example of **Figure 30**, there is a potential for cross-coupling. In case of **Figure 30**, LO1 can leak to mixer 2 (at a level of around –30dBm) and mix with 1650Mhz to give in–band spurious.

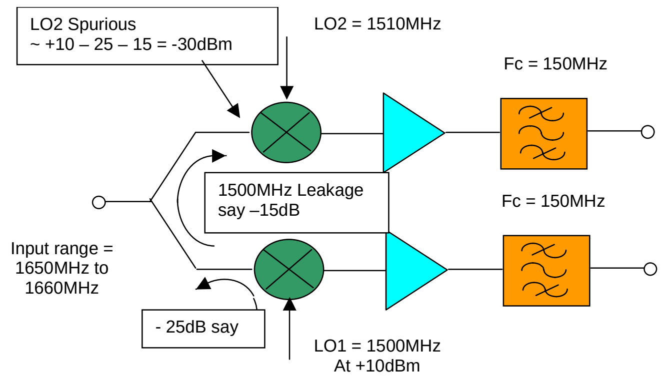


Figure 30 Multiplexed RF channels showing how the LO1 local oscillator can leak through to mixer 2 – at around –30dBm. This can then mix with the input signal at 1650Mhz to give in-band spurious.

Unfortunately, this circuit cannot be simulated in ADS as the LO2 breakthrough and RF input are mixing due to the non-linearity of the mixer, therefore, circuit measurement is usually required to evaluate mixer O/P level spurious.

4.16 TRANSPONDER INTERFERENCE

Relatively strong interference signals are sometimes specified at the equipment input. These signals may come from radars or nearby satellites for example). The solution to provide equipment immunity from these signals is to:

- Improve the input rejection filtering, assuming the interfering signal is not too close to the pass-band of the system.
- Improve the linearity of the input stages so that any interfering signal cannot produce spurious.

4.17 RADIATED SUSCEPTIBILITY

This EMC test involves radiating the powered up equipment and checking to see if the radiating signal has leaked into the RF equipment – to either appear on the RF output or generate internal spurious signals.
Some common RF leakage paths are:

- Co-axial connector – Check the torque settings, check for solder cracks and seal threads with electrically conductive sealant.
- Via screw-on covers – Reduce screw pitch, use EMC gaskets and conductively bond covers.
- Via module mounting – Check flatness, cleanliness, possibly use an EMC gasket.

If there is a problem, can gain from the leakage point to the equipment output be reduced?

4.18 D.C POWER

The DC power required for the equipment is usually a compromise between the following:

- The use of existing technology and/or hardware – This is the most cost effective option, but not necessarily best for minimum mass or DC power.
- Develop newer technologies to suit current needs – Using the latest devices/technology is expensive to develop and will effect the project schedule. This route may require more complex ASIC's etc.

4.19 TEMPERATURE CONSIDERATIONS

When the equipment is designed to operate over a large temperature range (eg space equipments operating between –25ºC to +80ºC), careful consideration is required on selecting the components used in the design:

- Are the components suitable for the temperature operation, including the manufacturing process?
- On high-reliability equipments (Eg space applications), certain parameters eg junction temperature, voltages, currents etc have to be derated from their maximum ratings.
- Do any components require special heat-sink arrangements? – Check the device junction-case thermal resistance/dissipation. Make adequate allowance for ALL thermal resistances (eg glues, contact area, materials used etc – If in doubt seek a thermal expert!).
- The equipment may need special circuit design, to meet it's specified performance over it's operating temperature range – Temperature compensation circuits (using extra feedback, RF Compensation, digital controlled compensation etc).
- Check the parts, processes and materials to be used.
- Does the thermal flux density require special attention? ie are there localized hot spots within the equipment?

4.20 VIBRATION

Most parts/equipments may need to be vibrated to ensure their survival eg for satellite launches, aircraft vibration etc.

- Identify the most susceptible parts to ensure that they are not made unnecessarily vulnerable. This may require vibration analysis of the equipment to find mechanical resonances etc. If this is a problem consider mounting the item elsewhere, flexible fixings etc.
- Design the equipment to meet the needs of the parts where possible – Don't pass the problem onto the supplier. Special parts will be more expensive (with the associated schedule and technical risks).

4.21 NATURAL RADIATION

This is mainly applicable to *space hardware* where there are high doses of radiation from cosmic rays and the Van Allen belts. Radiation doses are given in units of Grays where 1 Gray = 100rad, typically parts should be designed to survive levels of 100Krad.

- Assess the in-orbit dose that the electronic circuits will see (This analysis is normally performed by the radiation specialist). This analysis will result in a mission dose-depth curve (ie what metal plating will I require to ensure the circuits are at acceptable radiation dose levels).
- Analyse individual circuits for doses computed from the radiation analysis. – Select the most susceptible circuits. Obtain parts sensitivity. Check circuit performance as a result of device degradation due to radiation.
- Calculate extra shielding (if feasible) to achieve circuit performance compliance.
- Radiation analysis should form part of the worst case analysis (WCA).
- Initial design analysis must include some allowance for – Part radiation drifts, Part life drifts and temperature effects.
- Design circuits for good drift tolerance (including radiation effects) – especially for DC current gain (drops of 60% typical!), offset current/voltage and op-amp gains (DC gain can drop by 30dB at 1Kgray.

- Brehmsstrahlung level will limit practical screening, therefore select a 'harder' component and/or use a circuit that is more tolerant to drift.

5 SYSTEM APPROACH CHECKLIST

Finally here is a list of the main things to consider as part of the equipment design:

- Be aware of the Project Requirements, ensure that you fully understand what the equipment is to do and the definitions of it's parameters.
- Establish a functional block diagram (Include redundancy if necessary) – based upon the functional analysis.
- Check to see if the is a *single part failure* and if so is there a way of designing it out? Some customers may permit simple passive single part failures but they are best avoided.
- Examine closely to ensure that a failure does not jeopardize a redundant circuit – For example don't rely on a good input match from a failed amplifier!
- Incorporate existing hardware designs if possible in order to minimize cost and reduce risk.
- Divide the whole equipment into lower level assemblies.

  - Divide the technologies where sensible (For engineering disciplines).
  - Ease of testing (Provide intermediate test monitor points to minimize the number of sub-assemblies).
  - Minimize the number of different module or circuit designs.
  - Design the equipment for ease of manufacturing.

    - Produce simple short-form specifications for each functional block, in order to make the design more controllable.
    - Involve Mechanical & Production engineering at an early stage – Ensure that every body *buys into* the scheme, to avoid manufacturing and processing difficulties.
    - Be very cautious with new parts.

- Don't be influenced by salespersons over–enthusiastic claims.
- Do the offered engineering samples truly represent the production item performance?
- Avoid single sourcing if possible.
- Don't place unrealistic requirements on the part supplier as it can increase risk!

6 CONCLUSION

This tutorial described the typical parameters specified by customers for RF equipments in the form of the 'Equipment Specifications' that includes electrical, mechanical and environmental performance. Each design driver/parameter was described and illustrated with examples, verified by ADS simulations where applicable. Where possible solutions to potential design pitfalls were given, together with general guidance on ensuring a compliant design. It was noted however, there will be design trade-offs between certain key parameters (eg linearity and power consumption, noise figure and gain/attenuation etc.

Finally, a checklist of the main points to consider during the design process was given, which could form the basis of a design/peer review.

7 REFERENCES

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[2] Pekka Eskelinen, Introduction to RF Equipment and System Design, Artech House, 2004, ISBN 1-58053-665-4.

[3] M.M Radmanesh, Radio frequency and Microwave Electronics (Ilustrated), Prentice-Hall, 2001, ISBN 0-13-027958-7, p473.

[4] Devendra.K.Mistra, Radio Frequency & Microwave Communication Circuits: Analysis & Design, John Wiley & Sons, 2001, ISBN 0-471-41253-8, Chapter 2

[5] Mini-Circuits, P.O.Box 350166, Brooklyn, NY 11235 U.S.A, http://www.minicircuits.com.