

Receiver System - Centimeter Regime

Roger D. Norrod

August 12, 2003

Abstract

The receiver front-end of a radio telescope is generally considered to encompass components that amplify, filter, and frequency convert signals, provided by the antenna, to a level and frequency range appropriate for detection. This presentation will discuss critical parts of the centimeter wave radio astronomy front-end, and factors impacting the design and performance.

The feed efficiently converts propagating electromagnetic fields near a reflector antenna's focal point to a guided wave in coax or waveguide. Some types of feeds inherently detect and separate polarizations; other types require an orthomode transducer to deliver orthogonal polarizations to separate channels. Low-noise amplifiers, usually cryogenically cooled, amplify the signal and set the receiver noise level, and are followed by filters, mixers, and additional amplification. All the passive and active components add electrical noise to the signal, and models used during receiver design will be presented, explaining why loss and noise introduced in the early stages of the receiver are critical. The linear operating range of active components is limited by their power handling capacity, and how these limitations are considered will be discussed. We will also discuss stability of the receiver, and practical means to achieve the required performance.

1 Introduction

Figure 1 shows a very simplified block diagram of a typical heterodyne receiver front-end. (The term *front-end* generally excludes the part of a receiver that involves data acquisition, the *backend*). A heterodyne receiver involves at least one frequency conversion, or mix, and is by far the most common type of front-end for radio astronomy in the centimeter regime. This paper will provide an introduction to the major components in such a front-end and the system aspects typically faced by a designer.

2 Feeds

The feed is placed at or near a focal point of a reflector. The angle subtended by the reflector as seen by the feed, Figure 2, strongly influences the types of

feeds that may be used, and the details of their design. Generally, smaller angles require larger feeds in units of wavelengths.

The feed interfaces in an efficient manner between EM fields at the focal point and a guided transmission line. Compromises involve bandwidth, efficiency, size, polarization purity, beam shape, and other performance considerations. The gain required and practical considerations dictate the type of designs that may be considered.

Dipoles in a cavity or with a combination of reflectors are common at long wavelengths and at prime focus where their low gain is not a problem, and their small size in wavelengths is an advantage. Corrugated horns typically provide high gain and superior performance in most applications. Good circular symmetry of the beam, good polarization performance, and high efficiency are achieved. [1, 2]

3 Orthomode Transducer and Polarizer

The Orthomode Transducer, Figure 3 provides dual polarization reception, necessary in order to gather all the photons. OMTs come in many forms, depending on the wavelength and bandwidth needed [3]. In general, they have a square or circular waveguide port that connects to the feed horn and supports orthogonal linear polarizations, and two output ports separating the polarizations into two channels. By adding to the OMT input or output a phase shifter that delays one polarization by 90 degrees with respect to the other, circular polarization may be received or transmitted, Figure 4. It is difficult to design a broadband phase shifter, and this component currently limits the bandwidth of many receivers.

4 Thermal Noise in Receivers

Thermal noise, aka Johnson noise, arises from random motion of free electrons in a conductor, due to thermal agitation [4, 5, 6]. Figure 5 shows an electrical resistance at physical temperature T , connected to a band-limiting filter. The open-circuit rms voltage that will be measured at the output terminals is given in the figure. The quantities h and k are Planck's and Boltzmann's constants respectively, and f is frequency. If $hf/kT \ll 1$, the integral reduces to B , the form commonly used in microwave system analysis. However, the inequality needs to be checked if $f/T > 1$ GHz/Kelvin. Common electrical circuit theory says that for any voltage source, maximum power is available when load resistance is equal to the source resistance. Under this condition, the available power from the thermal noise source is kBT , Figure 6. The concept of equivalent source temperature is used for convenience when the statistics of any signal resembles gaussian noise. The true source is replaced with an equivalent thermal noise source that produces a noise power at the same level, Figure 7. The equivalent noise temperature, T_s , may be a function of frequency, but should be flat over the filter bandwidth B .

If we connect a thermal noise source to a microwave amplifier input, and measure the amplifier output power at various source temperatures, we will find that the output power is the noise source power times the amplifier gain plus a constant term, Figure 8. This constant is noise added by the amplifier, and we model that noise by thermal noise at the amplifier input.

Amplifier noise is not entirely thermal. That is, it is not directly proportional to the amplifier temperature, and is generally a function of frequency. GaAs or InP HFET (HEMT) devices are now used almost exclusively to build low-noise microwave amplifiers. In addition to being inherently low-noise, high gain, and stable, they can be designed to cool well - that is, much of their noise is thermal and decreases when cooled, Figure 9.

In a cascade of amplifiers, the noise contribution of a particular stage is divided by the gain of preceding stages, Figure 10. Therefore, one wants the first stages of a receiver to have high gain and low noise. Most radio astronomy receivers have cryogenically cooled amplifiers for the first stages of gain. Figure 11 shows that when losses exist at the input of a receiver, two negative effects plague the system. An ohmic loss adds noise at a level proportional to its temperature, plus it attenuates the input signal. The net effect is to both add noise, and to multiply the effective noise temperature of following stages.

5 Frequency Conversion

Figure 12 shows a receiver using a mixer for frequency conversion. The mixer has at its output multiple instances of the input RF and LO signals shifted up and down in frequency by integer multiples. Use of filters allows selection of the m,n coefficients for the output signal. Use of mixing within a receiver provides several advantages:

Tunability: By using a narrow IF filter and a variable LO frequency, the detected RF frequency can be varied over a wide range.

Cost: Amplifiers, transmission lines, and other components are generally less expensive at IF frequencies.

Stability: Large amounts of gain at one frequency can cause instability due to leakage from the output circuits coupling back to the input stages. In a heterodyne receiver, the gain is distributed at two or three frequency ranges, reducing the likelihood of feedback coupling. Also, amplifiers tend to be less sensitive to temperature and other environmental effects.

RFI: Down-conversion of a frequency range increases the ratio of frequencies in the IF passband. Filtering out unwanted signals at the IF is simplified because filter responses scale by the ratio of frequencies. Hence fewer filter resonators are required for a given level of rejection.

6 Filters

Filters are necessary in most any receiver system, to reject interference, select desired mixer products, and to define the receiver passband. Digital filters provide many advantages, but currently most filters remain analog designs for reasons of cost and simplicity, particularly at frequencies above 100MHz.

The most common general purpose filter response is the *Tchebyscheff* or *Equal-Ripple* type. The Tchebyscheff equations describe filter insertion loss response using only two parameters: the number of resonators, and the magnitude of the passband loss ripple. Figure 13 shows the Tchebyscheff equations and a lowpass filter response with five resonators, for 0.1, 1, and 3dB passband ripple. Higher passband ripple yields steeper rejection slopes above the cutoff frequency, but keep in mind that the filters reflect power rather than absorb it. So, 3dB passband ripple means half the input power is reflected back toward the source at the peaks of the ripples.

Figure 14 shows the Tchebyscheff lowpass response for filters with 3, 5, 7, and 9 resonators and 0.1dB passband ripple. Such curves can be used to determine the minimum number of poles required to achieve specified rejection. For example, if 40dB rejection is required at a frequency 1.5 times the filter cutoff frequency, at least 7 poles are required. Note that one faces diminishing returns as poles are added, because the improvement in rejection at a given frequency decreases as the number of poles increase.

Once an appropriate lowpass response has been chosen, equations are available [8] to determine circuit elements that will realize the response. Figure 15 shows the classic lowpass filter topology using inductors and capacitors. At RF frequencies, lumped elements have significant parasitic elements which must be accounted for. Computer modeling and optimization is often necessary. At microwave frequencies, transmission line elements must substitute for lumped elements to achieve acceptable filter performance.

Design of a bandpass filter often begins with selection of a lowpass prototype response, which then can be mapped to a bandpass response by simple substitution of the frequency variable. Figure 16 illustrates the variable substitution and a Tchebyscheff 0.1dB ripple, 5-pole response mapped to frequency of 19.8 to 20.2. Figure 17 shows how a lowpass prototype circuit can be transformed to a bandpass topology. Inductors are replaced by series resonators and capacitors by shunt resonators. Good lumped elements are generally not available at microwave frequencies, so microwave filters are realized using transmission line resonators. Various transmission line forms (microstrip, stripline, coaxial, waveguide) may be used, based on tradeoffs involving size, cost, and performance.

7 Receiver Stability

Figure 18 shows a greatly simplified diagram of a total power receiver. The receiver gain is represented by G , and typically consists of many amplifiers,

mixers, and other microwave and IF components with net gain well over 100 dB. The IF signal presented to the detector is band-limited gaussian noise with effective noise bandwidth of B . The output voltage of the square-law detector is proportional to the power of the input signal. The detected output is then integrated with time constant τ , represented by a RC circuit. The output voltage V_o , proportional to the detected power, has both DC and AC components. The ratio of these two components is inversely proportional to the square root of the product of B and τ [7], assuming the receiver gain and noise temperature is constant. A radiometer's sensitivity is often defined as the change in antenna equivalent temperature which results in a change of V_o equal to the rms value of V_{ac} . However, since a change in G also results in a change in V_o , the receiver gain must be much more stable than V_{ac}/V_{dc} for periods greater than τ .

Gain stability is achieved by temperature control, both active and passive, design of components that are not sensitive to vibration and shock, and avoidance of stress on cables and connectors by careful alignment of components. Even so, sophisticated switching techniques are often required to achieve adequate sensitivity.

8 Receiver Linearity

Any practical amplifier, mixer, or other active device has limited power-handling capability. At some level of output power, the gain begins to become non-linear, and eventually the output power saturates, figure 19. For sinusoidal inputs, the output voltage becomes clipped and approaches a square wave, producing harmonics in the output frequency spectrum. Sum and difference terms arise in the output spectrum if multiple frequencies are present at the input [9].

The most troublesome intermodulation products are usually odd-order difference terms. The strongest of these are the third-order products, Figure 20. For two closely spaced input tones F1 and F2, the 2F1-F2 and 2F2-F1 products fall near F1 and F2 and are often impossible to filter out. A log-log plot of the third-order product level has a slope of 3:1. For example, if the level of F1 and F2 is dropped by 10 dB, the third-order products drop by 30 dB, a relative improvement of 20 dB.

The third-order performance of a component is usually specified by the *third-order intercept*, a fictitious power level where the fundamental and third-order curves intersect. The device is usually not capable of actually producing that level of power, but knowing this point and the slopes, the relative third-order levels can be calculated for any reasonable fundamental output power. For example, if the fundamental tones are 20 dB below the intercept point for a particular amplifier, the third-order products will be 60 dB below the intercept, or 40 dB below the fundamental tones.

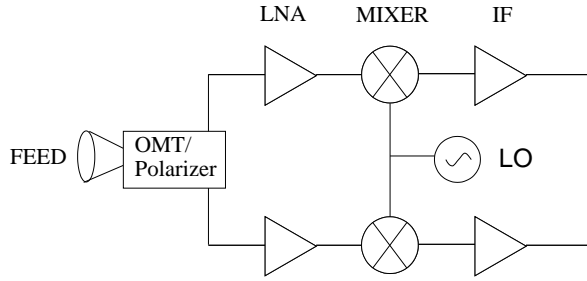


Figure 1: A simplified block diagram of a typical radio astronomy receiver.

9 Summary

The most common topology for centimeter-wave radioastronomy receivers is dual-polarization heterodyne with cooled HEMT low-noise amplifiers. The feed, OMT, polarizer, and first amplifiers are often the most critical components. Gain stability has a large effect on data quality and observing efficiency and must be considered and measured during receiver design and construction. The linearity and power handling of the receiver determines its performance in the presence of strong unwanted signals such as interference from man-made terrestrial or space-borne sources.

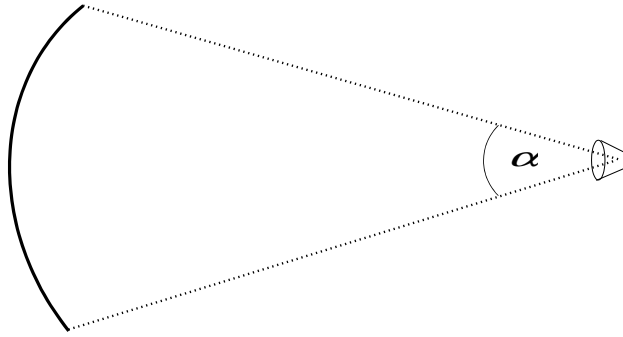


Figure 2: The angle subtended by the reflector as seen by the feedhorn is an important parameter. As the angle gets small, the feedhorn size increases in terms of wavelengths in order to maintain high efficiency and avoid excessive spillover.

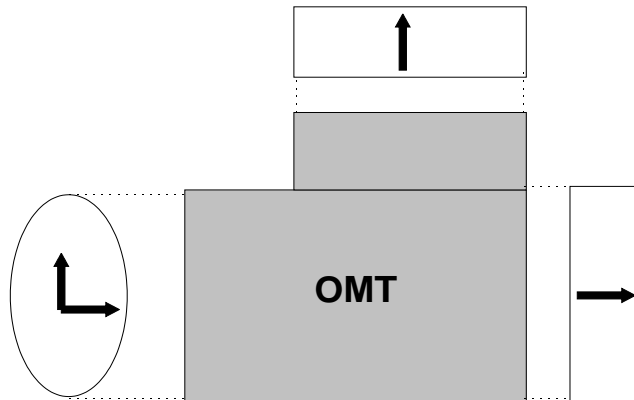


Figure 3: The Orthomode Transducer (OMT) separates dual-linear polarizations at its square or round input port into two output ports, usually rectangular waveguide or coax.

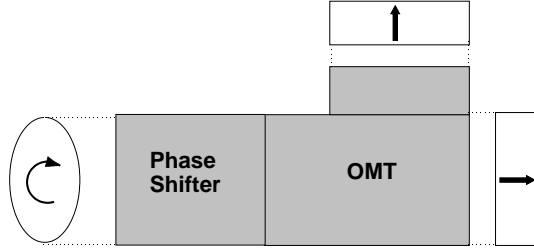
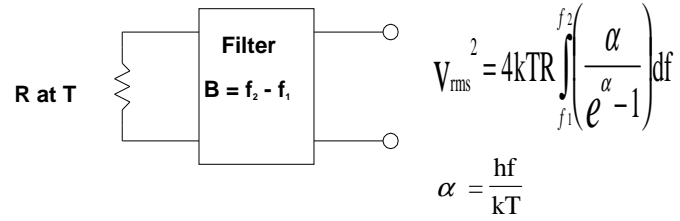


Figure 4: The addition of a phase shifter, which retards one polarization by 90° , at the input of an OMT causes the device to respond to circular polarizations.



$$\text{If } \alpha \ll 1, \quad v_{\text{rms}}^2 = 4RkBT$$

Figure 5: Random voltage fluctuations generated by ohmic conductors were discovered experimentally by Johnson [5], and theoretical equations were developed by Nyquist [6].

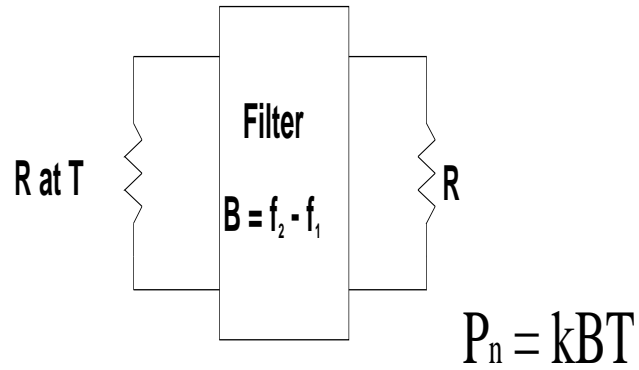


Figure 6: The available thermal noise power under matched impedance conditions is given by $P_n = kBT$.

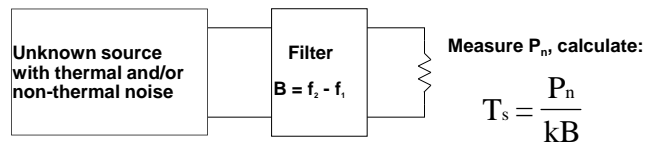
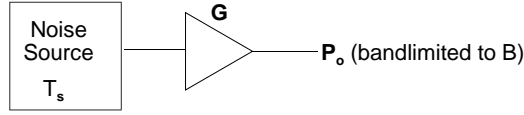


Figure 7: An unknown source of gaussian noise can be characterized by the temperature of a thermal source which produces the same noise power over the frequency range of interest.

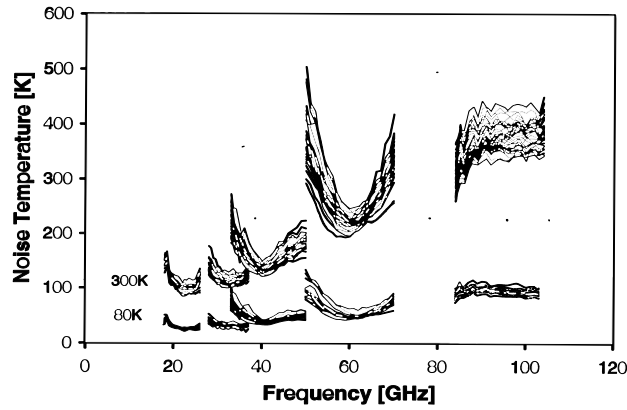


$$P_o = GkBT_s + K$$

$$\text{Define } K = GkBT_e$$

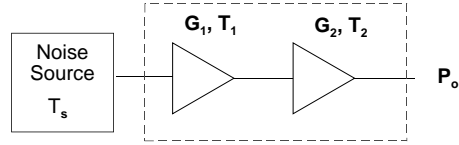
$$\text{Then, } P_o = GkB(T_s + T_e)$$

Figure 8: Amplifiers generate noise internally by thermal and non-thermal mechanisms. The noisy amplifier may be represented by an equivalent noiseless amplifier with a thermal noise source at its input.



Data courtesy M. Pospieszalski of NRAO Central Development Laboratory

Figure 9: Measured data for over 100 low-noise HFET amplifiers produced by the NRAO Central Development Lab for a recent project. Data from five amplifier types at 300 Kelvin and at 80 Kelvin is shown. Note the noise generally increases with frequency and decreases markedly with cooling.



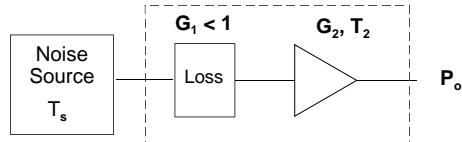
$$P_o = G_1 G_2 k B T_s + G_1 G_2 k B T_1 + G_2 k B T_2$$

or,

$$P_o = G_1 G_2 k B (T_s + (T_1 + T_2/G_1))$$

So, Amplifier Cascade has equivalent noise $T_1 + T_2/G_1$

Figure 10: The noise and gain of the first amplifier in a cascade dominates the total noise temperature. The first amplifier in a receiver is chosen to have low-noise and high gain for this reason.



Let $L = 1/G_1$, then for ohmic loss at physical temperature T_o ,

the effective noise temperature of the loss is $(L-1)T_o$.

Effective noise temperature of the loss - amplifier cascade

is: $(L-1)T_o + LT_2$.

Figure 11: Loss before the first receiver amplifier adds noise due to its ohmic loss, and effectively multiplies the noise temperature of following stages.

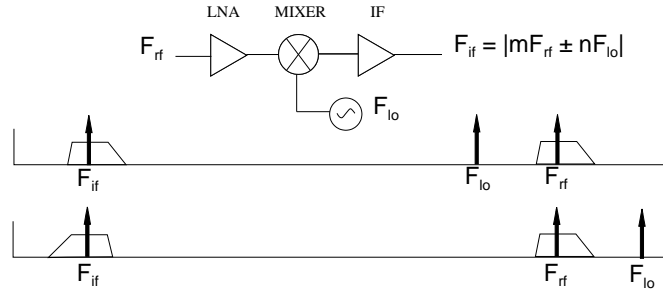


Figure 12: A simplified block diagram of a frequency down-converter is shown. The middle diagram illustrates a upper-sideband conversion in which the LO is below the RF band. The bottom diagram illustrates a upper-sideband conversion in which the LO is above the RF band. Note that in the latter case, the RF spectrum is inverted at the IF frequency. The selection of upper or lower sidebands and the IF range is defined by filters which are not shown.

Tchebyscheff filter response in dB:

$$\text{Tcheby}(n, \omega, \epsilon) := \begin{cases} \left(10 \log \left(1 + \epsilon \cdot \cos(n \cdot \arccos(\omega))\right)^2\right) & \text{if } \omega \leq 1 \\ \left(10 \log \left(1 + \epsilon \cdot \cosh(n \cdot \operatorname{acosh}(\omega))\right)^2\right) & \text{if } \omega > 1 \end{cases}$$

$n := 5$ Number of Resonators

$\epsilon_1 := 10^{\left(\frac{0.1}{10}\right)} - 1$ 0.1dB Ripple $\epsilon_2 := 10^{\left(\frac{1}{10}\right)} - 1$ 1dB Ripple $\epsilon_3 := 10^{\left(\frac{3}{10}\right)} - 1$ 3 dB Ripple

$\omega := 0.1, 0.15, .3$

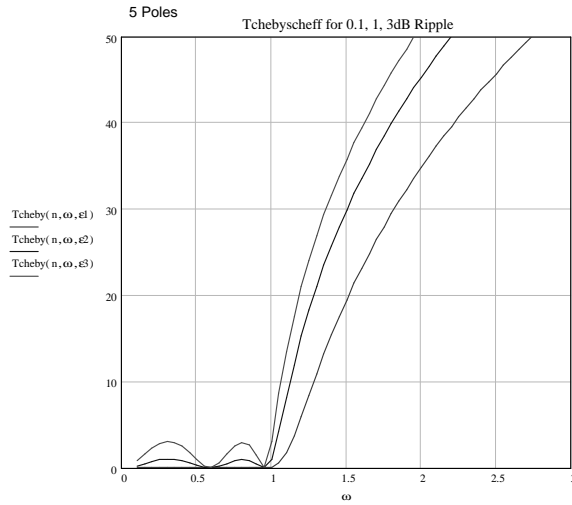


Figure 13: Tchebyscheff low-pass filter response is shown, along with the governing equations, for three levels of passband ripple. The x-axis is frequency, normalized to the filter cutoff frequency. Note how the filter rejection curve gets steeper as passband ripple increases.

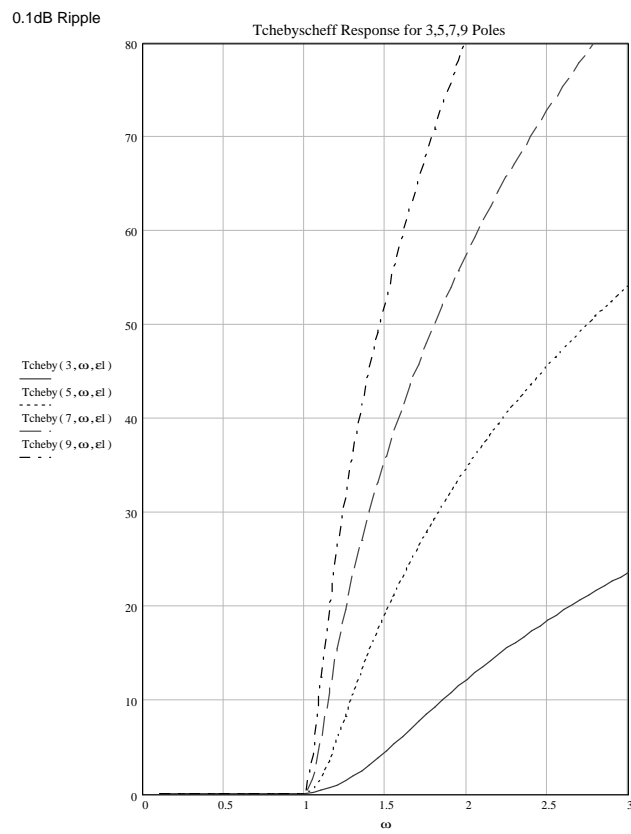


Figure 14: Tchebyscheff response for 3, 5, 7, or 9 filter resonators (poles).

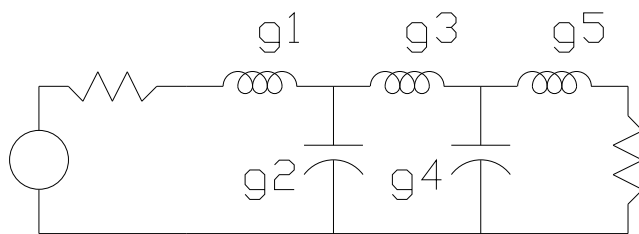


Figure 15: The classic lowpass filter circuit topology using ideal (lumped) inductors and capacitors.

Lowpass to Bandpass Mapping

$$\omega_{lpf}(\omega, \omega_0, B) := \frac{1}{B} \cdot \left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \right)$$

$$\omega_1 := 19.8 \quad \omega_2 := 20.2 \quad \omega_0 := \sqrt{\omega_1 \omega_2} \quad B := \frac{\omega_2 - \omega_1}{\omega_0}$$

$$\omega := 18, 18.05, 22$$

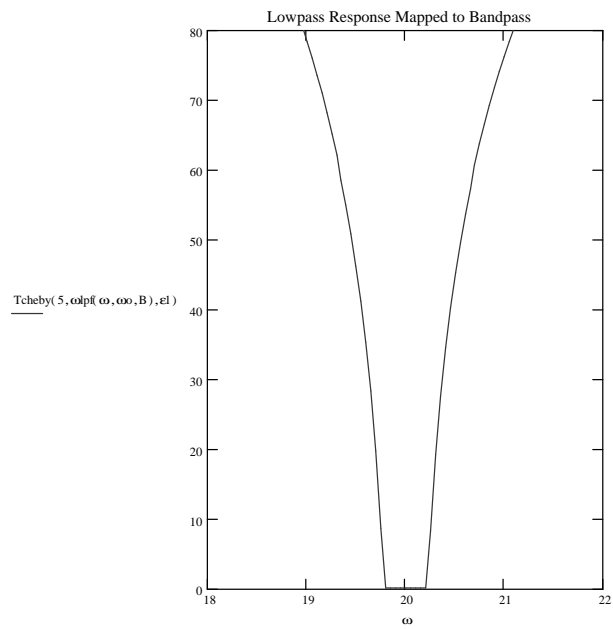


Figure 16: A lowpass filter response can be mapped to a bandpass response using a substitution of the frequency variable.

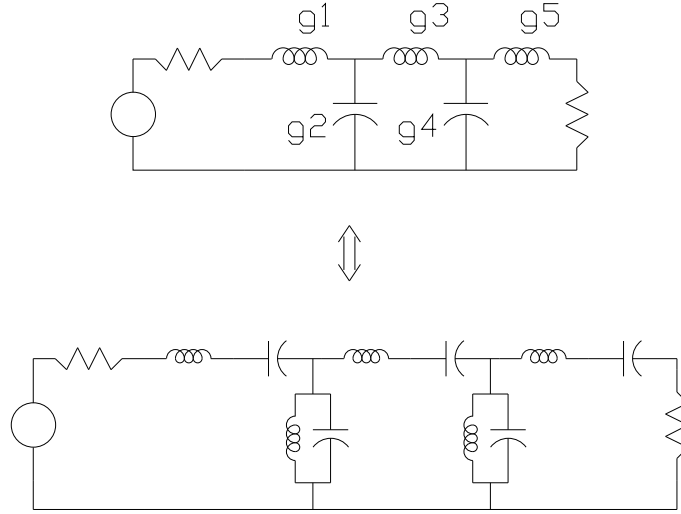


Figure 17: The lowpass topology can be transformed to a bandpass topology, replacing inductors with series resonators and capacitors with shunt resonators. Equations for the transformation are given in [8].

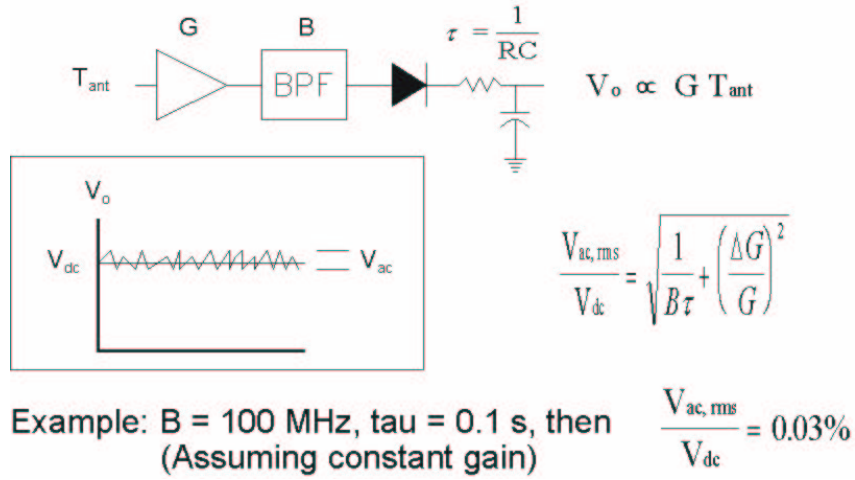


Figure 18: A simplified receiver detection diagram. The output voltage fluctuates about a DC value, and the amplitude of the fluctuations is inversely proportional to the square-root of the pre-detection bandwidth and the RC time constant, as long as gain fluctuations are kept small.

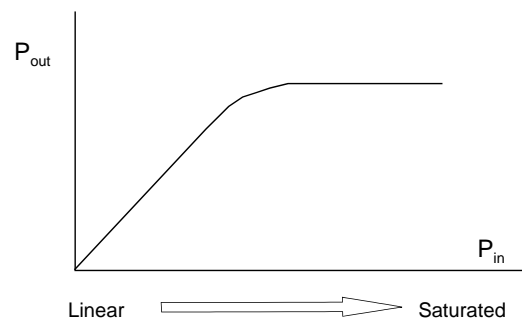


Figure 19: Amplifiers or other active components have limited output power capability.

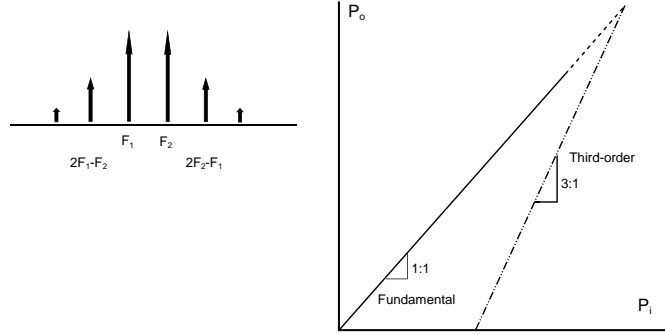


Figure 20: Non-linearity in a component gives rise to intermodulation products as strong input signals mix and produce products that can be difficult or impossible to remove by filters.

References

- [1] A.D. Olver, P.J.B. Clarricoats, A.A. Kishk, L. Shafai, *Microwave Horns and Feeds*, IEEE Press, 1994.
- [2] A. W. Rudge, K. Milne, A. D. Olver, P. Knight, *Handbook of Antenna Design, Volume I*, Peter Peregrinus Ltd., 1982.
- [3] J. Uher, J. Bornemann, U. Rosenberg, *Waveguide Components for Antenna Feed Systems: Theory and CAD*, Artech House, 1993.
- [4] H.C. Pazkowski, J. Whelehan, "Understanding Noise, Part I", IEEE MTT Newsletter, No. 118, Spring 1987.
- [5] J. B. Johnson, "Thermal Agitation of Electricity in Conductors", Physical Review, Vol. 32, pp. 97-109, July 1928.
- [6] H. Nyquist, "Thermal Agitation of Electric Charge in Conductors", Physical Review, Vol. 32, pp 110-113, July 1928.
- [7] J. D. Kraus, *Radio Astronomy*, McGraw Hill, 1966, Ch. 7.
- [8] G. L. Matthaei, L. Young, E. M. T. Jones, *Microwave Filters, Impedance-Matching Networks, and Coupling Structures*, McGraw-Hill, 1964.
- [9] S. A. Maas, *Non-Linear Microwave Circuits*, Artech House, 1988, pp1-16.