Integrated LNA and Mixer Basics

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ABSTRACT

Basic theory and operation of low noise amplifiers and mixers are presented. Important figures of merits of these two devices such as gain, noise figure, compression point, and third order intercept point are introduced and derived. Measurement methods of these figures of merit are also described.

LNA

Low noise amplifiers (LNAs) are widely used in wireless communications. They can be found in almost all RF and microwave receivers in commercial applications such as cordless telephones, cellular phones, wireless local area networks, and satellite uplinks and downlinks and in military applications such as doppler radars and signal interceptors. Depending upon the system in which they are used, low noise amplifiers can adopt many design topologies and structures—those used in military applications tend to be discrete, large in size, and consume high power, whereas those in commercial applications aim toward high integration and low voltage and bias currents. The LMX2215 and LMX2216B, for example, can be classified into the latter catergory. LNAs are usually placed at the front-end of a receiver system, immediately following the antenna. A band pass filter may be required in front of it if there are many adjacent interfering bands leaking through the antenna, but this filter generally degrades the noise performance of the system. The purpose of an LNA is to boost the desired signal power while adding as little noise and distortion as possible so that retrieval of this signal is possible in the later stages in the system. With this in mind, low noise amplifier designers have developed many design concepts and theories applied to low noise amplifiers and important figures of merit used to characterize and compare their performance. These concepts and figures of merit are discussed in the following sections.

MIXER

Mixers are found in virtually all wireless communication systems. They are frequency translating devices that convert input signals from one frequency to another by mixing these signals with another signal of known frequency. One reason frequency translation is a necessary process in wireless transmission is that information signals such as human speech or digital data are usually low frequency signals and are not suitable for a wireless channel. Another is that wireless channels are common channels that are shared by many signals and these signals must be separated into different frequency bins so that electronic circuits (which contain frequency selective components) can keep them from destructively interfering with each other. Among many other properties, frequency is one that is most easily exploited in signal identification.

Mixers can be classified into two broad categories: passive or active. The most commonly available and used are passive diode mixers since they are easier to design and more thoroughly understood. Active mixers, on the other hand, involve transistors and the most popular ones are built from the basic Gilbert cell structure. Some higher frequency active mixers exploit the nonlinear characteristics of high gain transistors and can perform the mixing action using only one transistor. Among these types, the Gilbert cell structure has the most desirable characteristics in terms of isolation and harmonic suppression due to its balanced structure. The LMX2215, LMX2216B, and LMX2213B use the Gilbert cell (the LMX2216B is the 3V equivalent of the LMX2215, and the LMX2213B is the LMX2216B without the LNA).

Most down converting mixers are three-port devices, as shown in *Figure 1.* They take two input signals: the RF and the LO (local oscillator) signals. The output is a mixing product of these two inputs and is an intermediate frequency (IF) signal. There are self-oscillating mixers which provide their own LO signal by having an internal resonating element coupled with the RF input. The LMX2215 and LMX2216B require external LO drives.

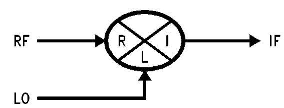

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FIGURE 1. Three-Port Mixer

Mixers perform the mixing operation by multiplying the two input signals. The output, IF, is the product of the two signals RF and LO, and it contains the sum and difference of the two input frequencies. In receivers, the lower frequency component is usually the desired one and can be obtained by lowpass filtering the mixer output signal. Derivations of the mixer effect are shown below in the section on nonlinearities (OIP_3).

CONCEPTS

Noise

Noise in electrical systems is defined as random fluctuations in voltage and current. It can be generated internally by components employed in the system or externally by electrical radiation from other systems or induced mechanical vibrations. RF and microwave oscillators, for example, are very susceptible to external radiation if they are not properly shielded. They are also susceptible to mechanical vibrations, a phenomenon called microphonics, if they are not sufficiently isolated from physical contact with nearby objects. Integrated low noise amplifiers are, on the other hand, most vulnerable to noise that is generated by their own transistors and resistors. Transistors exhibit flicker noise, which is caused by a change in conductance caused by a relatively slow process (e.g. the exchange of charge with surface traps or metallic impurities through tunneling), and shot noise, which is due to random one-way crossings of some barrier by discrete quantities of charge. For amplifiers at radio and microwave frequencies, flicker noise is negligible since it's power spectrum has a 1/f property. The

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power spectral density of flicker noise is described by equation (1) below:

$$G_i(f) = C_1 \frac{I^a}{f^b} \quad (1)$$

where a ~ 1 to 2, b ~ 1, and C_1 is a device dependent constant. Shot noise power, however, depends on the net total current crossing the pn junctions, and its power spectral density is given by

$$G_i(f) = qI \quad (2)$$

where q is the electronic charge and I is the total current. Resistors exhibit thermal noise, which is generated by the random movement of electrons inside the resistive material at a non zero absolute temperature. The thermal noise power (per unit of frequency) of resistors, thus, depends on temperature and the resistance value of the resistors. However, the available thermal noise power depends solely on temperature. Equation (3) gives the power spectral density of thermal noise

$$G_v(f) = KTR \quad (3)$$

where K is the Boltzmann's constant, T is the absolute temperature in Kelvins, and R is the resistance.

The combined effect from the noise sources mentioned above and all other possible noise sources is often treated as though it were caused by only thermal noise. Moreover, LNAs are sometimes specified, not by their noise figure, but by their noise temperature, the temperature at which a resistor would generate the equivalent noise power.

Noise Figure (NF)

Noise figure is noise factor in decibel units (dB) and is an important figure of merit used to characterize the performance of not only a single component but also the entire system. It is one of the factors which determine the system sensitivity. Noise factor is defined as the input signal to noise ratio divided by the output signal to noise ratio. For an amplifier, it can also be interpreted as the amount of noise introduced by the amplifier seen at the output besides that which is caused by the noise of the input signal. Mathematically,

$$F = \frac{S_i/N_i}{S_o/N_o} = \frac{S_i/N_i}{G_a S_i/(N_a + G_a N_i)} = \frac{N_a + G_a N_i}{G_a N_i} \quad (4)$$

$$NF = 10 \log (F) \quad (5)$$

where S_i, and N_i, represent the signal and noise power levels available at the input to the amplifier, S_o and N_0 the signal and noise power levels available at the output, G_a the available gain, and N_a the noise added by the amplifier. For a mixer which is used in applications where the desired signal power is contained in only one sideband, N_i is interpreted as the input noise contained in only one sideband. Therefore, in specifying noise figure for mixers, the term single-sideband or double-sideband must be noted to indicate how N_i was measured. In most communication receivers, single-sideband noise figure is the "true" noise figure and is 3 dB higher than double-sideband noise figure.

Design for Optimum Noise Performance

Based on the above equations, noise models for transistors can be developed. Furthermore, analysis of these models shows that for an amplifier using bipolar transistors, the only noise determining factor is the input match (which can also be translated into a bias current dependence). If resistors are also employed in the matching networks, then these will affect the noise performance as well. For each transistor operating at a particular frequency and bias current, there exists an optimum input match $\Gamma_{opt} = R_{opt} + jX_{opt}$ (6), which will yield an optimum noise figure F_{opt} This input match can be obtained by measurements using a noise figure meter and a vector network analyzer. A matching network designed to present this optimum impedance at the input of the transistor yields the optimum noise performance. The noise figure of the resulting amplifier can be calculated using the following formula,

$$F = F_{opt} + \frac{4R_n}{Z_0} \frac{|\Gamma_s - \Gamma_{opt}|^2}{|1 + \Gamma_{opt}|^2 (1 - |\Gamma_s|^2)} \quad (7)$$

where R_n = noise resistance

Z_0 = system impedance

Γ_s = input reflection coefficient seen by the device (see *Figure 1*).

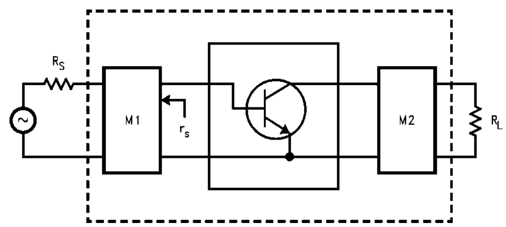

FIGURE 2. Typical LNA Topology

Note that if the input match is perfect, the noise figure is F_{opt}. This value is usually not achievable in practice and tradeoffs between noise performance, match to available filters, gain, and stability is often required.

Noise Figure Measurements

Noise figure can be measured using a noise figure meter, which consists of a noise source and an RF receiver. The noise source is placed at the input of the device under test (DUT), and the output of the DUT is connected to the receiver (see *Figure 3*). There are several methods which noise figure receivers use to calculate noise figure, one of which involves computing the Y factor. With this method, the noise source (an avalanche diode) is cycled between two effective noise temperatures: T_h and T_c, shown in *Figure 4.* T_h corresponds to the hot temperature, when the diode is bias with a DC current, and T_c corresponds to the cold temperature, when the diode is off. The receiver detects the noise power at the output of the DUT under these two temperatures and computes the straight-line noise characteristics, from which the noise added, N_a, can be determined. Along with N_a, the noise figure meter also measures the available gain of the DUT to compute the noise figure using equations (4) and (5). *Figures 3* and *4* below illustrate the measurement setup and the straight-line noise characteristic.

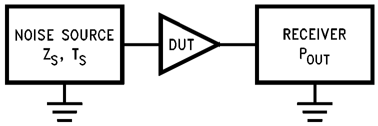

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FIGURE 3. Noise Figure Measurement Setup

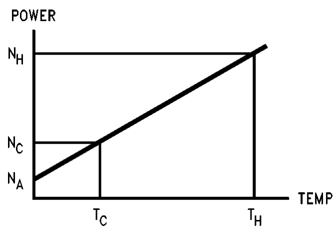

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FIGURE 4. Noise Power vs Noise Source Temperature

GAIN (G)

At radio and microwave frequencies, efficiency in transmission of signal power is of great importance. For this reason, RF and microwave circuits are optimized for power gain instead of voltage or current gain as commonly found in most low frequency circuits. The unit of power used to specify absolute power level is the dBm, or decibels referenced to 1 mW. Power levels in dBm can be computed from the equation

$$P(\text{dBm}) = 10 \log \left(\frac{P(\text{mW})}{1 \text{ mW}}\right) \tag{8}$$

In cases where the load impedance is known or assumed, equivalent voltage levels can be used to specify power levels indirectly. In these cases, the unit dBμV is often used. A similar equation converts μV units to dBμV units.

$$V(\text{dB}\mu\text{V}) = 20 \log \left(\frac{V(\mu\text{V})}{1\ \mu\text{V}}\right) \tag{9}$$

The importance of power transfer is one of the reasons for which power gain, and not voltage or current gain, is often used to specify RF and microwave devices. Many different types of power gain are used in RF engineering. The type used here is called transducer gain, which is defined as the ratio of the power delivered to the load to the available power from the source,

$$G = \frac{P_{out}}{P_{in}} = \frac{V_{out}^2/R_L}{V_{in}^2/4R_s} = 4\,\frac{R_s}{R_L}\,\frac{V_{out}^2}{V_{in}^2} \tag{10}$$

where V_{out} is the voltage across the load R_L, and V_{in} is the generator voltage with internal resistance R_s. In terms of scattering parameters, transducer gain is defined as

$$G = 20 \log (|S_{21}|) \tag{11}$$

where S_{21} is the forward transmission parameter, which can be measured using a network analyzer.

1 dB COMPRESSION POINT (P_{1dB})

A measure of amplitude linearity, 1 dB compression point is the point at which the actual gain is 1 dB below the ideal linear gain. For a memoryless two-port network with weak nonlinearity, the output can be represented by a power series of the input as

$$v_o = k_1 v_i + k_2 v_i^2 + k_3 v_i^3 + \ \ldots \tag{12}$$

For a sinusoidal input,

$$v_i = A\cos \omega_1 t \tag{13}$$

the output is

$$v_0 = \frac{1}{2} k_2 A^2 + \left(k_1 A + \frac{3}{4} k_3 A^3\right) \cos \omega_1 t + \frac{1}{2} k_2 A^2 \cos 2\omega_1 t + \frac{1}{4} k_3 A^3 \cos 3\omega_1 t \tag{14}$$

assuming that all of the fourth and higher order terms are negligible. For an amplifier, the fundamental component is the desired output, and it can be rewritten as

$$k_1 A \left[1 + \frac{3}{4} (k_3/k_1) A^2\right] \tag{15}$$

This fundamental component is larger than k_1 A (the ideally linear gain) if $k_3 > 0$ and smaller if $k_3 < 0$. For most practical devices, $k_3 < 0$, and the gain compresses as the amplitude A of the input signal gets larger. The 1 dB compression point can be expressed in terms of either the input power or the output power. Measurement of P_{1dB} can be made by increasing the input power while observing the output power until the gain is compressed by 1 dB.

P_{1dB} is an important characteristic of a device since it indicates the upper limit of the power level of the input signal without saturating the device and generating nonlinear effects.

THIRD ORDER INTERCEPT (OIP_3)

Third order intercept is another figure of merit used to characterize the linearity of a two-port. It is defined as the point at which the third order intermodulation product equals the ideal linear, uncompressed, output. Unlike the P_{1dB}, OIP_3 involves two input signals. However, it can be shown mathematically (similar derivation as above) that the two are closely related and $OIP_3 \approx P_{1dB} + 10$ dB. Theses two figures of merit are illustrated in *Figure 5* below.

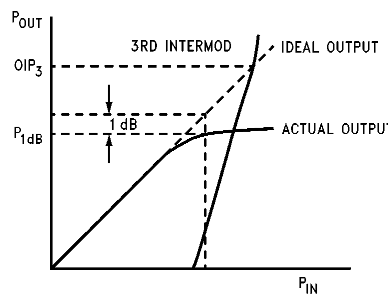

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FIGURE 5. Typical P_{out}–P_{in} Characteristics

Third order intermodulation products are important since their frequencies are located close to the wanted signal frequency, making them more difficult to be rejected by practical filters. If the two-port network is an LNA used in a receiver, intermodulation products at the output of the LNA can mask out signals from adjacent channels. For example, the third order intermodulation products resulted from 2 channels at 1 MHz apart, f1 = 408 MHz and f2 = 409 MHz, will be at 407 MHz and 410 MHz, a 1 MHz offset from f1 and f2. Similarly, two channels f4 = 411 MHz and f5 = 412 MHz will produce intermodulation products at 410 MHz and 413 MHz. If f3 = 410 MHz is the desired signal, it will be interfered with by the intermodulation products created by its adjacent channels f1, f2, f4, and f5.

To see how third order intermodulation products come about, assume that the input to a two-port with the same output-input relationship as stated in the above section consists of a sum of two sinusoids:

$$v_i = A\,(\cos \omega_1 f + \cos \omega_2 t) \tag{16}$$

Then, the output voltage is

$$\begin{aligned} v_o = {} & k_1 A\,(\cos \omega_1 t + \cos \omega_2 t) \\ & + k_2 A^2 (\cos \omega_1 t + \cos \omega_2 t)^2 \\ & + k_3 A^3 (\cos \omega_1 t + \cos \omega_2 t)^3 + \ldots \end{aligned} \tag{17}$$

Expanding these square and cube terms and ignoring the higher order terms, the output voltage is seen to contain not only harmonics of each of the two individual input frequencies but also the intermodulation terms:

$$\frac{3}{4} k_3 A^3 \cos(2\omega_2 \pm \omega_1)t$$

and

$$\frac{3}{4} k_3 A^3 \cos(2\omega_1 \pm \omega_2)t$$

The amplitude of these terms are proportional to the cube of the amplitude of the input signals; therefore, these terms increase three times faster than the fundamental term as the input signals increase, as can be seen in *Figure 5.*

MATCHING

Matching and microwave circuit design are to some designers synonymous. It is the act of making the source and load impedances matched to achieve the desired amount of power reflected and power transferred. Matching is required if the circuit is to yield optimum gain and return loss. Poorly matched devices can cause large amount of reflected power, poor noise performance, and low gain. For an LNA, power reflected caused by improper input match can travel back to the antenna and be re-radiated. Poor input match can also reduce the gain of the LNA and causes the system to have non-optimum noise performance.

Amplifiers can achieve maximum gain and return loss when they are presented with conjugate impedances at the input and output ports. There are two types of matching networks: resistive and reactive. Resistive matching networks rely on resistive elements for matching, usually have wider bandwidths, and consume more power than their reactive counterparts, which use lossless elements (capacitors and inductors). Simple matching networks can be designed with the help of the Smith chart, but more complicated ones often require the use of a computer and some type of network synthesis software.

Standard input and output impedances of most microwave instruments are 50Ω. Therefore, microwave and RF devices are designed to have 50Ω input and output impedances so that they can be easily characterized. In a communication system, however, not every component can be designed or optimized for 50Ω impedances due to other constraints. While most RF ceramic or helical filters have 50Ω impedances, most available SAW filters used to filter intermediate frequencies, for example, exhibit 200Ω impedances, IF ceramic filters usually have impedances of 330Ω, and crystal filters have 1 kΩ impedances. So, devices that are designed to be used with these components may have input or output impedances that are different from 50Ω and need matching networks to perform the necessary impedance transformation for proper characterization. In this case, simple narrow band LC matching networks can be designed to operate at the frequency of interest. Narrow band matches are also useful to reduce NF in some devices and to trade current for voltage in low headroom power amplifiers (such as 3V devices). Shown below is an example of a 50Ω to 200Ω matching network

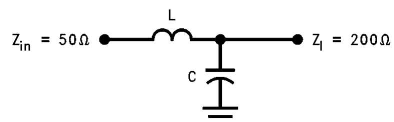

FIGURE 6. 50Ω–200Ω Matching Network

The actual values of inductance and capacitance for the above network depend on the frequency of operation, f, and can be obtained using the following equations

$$\frac{L}{C} = 10000, \qquad LC = \frac{0.75}{(2\pi f)^2} \tag{18}$$

In general, for a step-up transformer where $Z_{in} < Z_l$ and both are real impedances, the following equations apply:

$$\frac{L}{C} = Z_{in} Z_l\,, \qquad LC = \frac{1 - Z_{in}/Z_l}{(2\pi f)^2} \tag{19}$$

MIXER CONVERSION GAIN

Conversion gain of mixers is defined as the delivered IF power divided by the available input RF power. The term conversion is used to refer to the frequency converting action of the mixer. Conversion gain can be measured using similar method and equipment setup as those used to measure amplifier gain. More details on conversion gain measurement are deferred to a later application note.

MIXER ISOLATION

Isolation is a measure of how much power is coupled from one port to the next. The two most useful isolation measurements are LO-to-IF isolation and LO-to-RF isolation. The former indicates how much LO power leaks through the output IF port, and the latter indicates how much LO power leaks through the input RF port. LO appearing at the output IF port can be attenuated easily by a lowpass filter since the two frequencies are far apart, but it is more difficult to suppress at the RF port. LO leakage through the RF port usually results in a re-radiation through the antenna if the mixer is used as the first downconverter in a wireless receiver.

MIXER NOISE FIGURE (DSB vs SSB)

Mixer noise figures can be measured and specified in two ways: double side band or single side band. Double side band noise figure measurements involve measuring the noise power contained in both the IF and image components, whereas single side band measurements demand that the image component be filtered out, and only the noise power in the IF component is measured. The DSB method assumes that the gain of the DUT is the same at both image and intermediate frequencies, so it is not recommended for a narrowband DUT or a high intermediate frequency. The SSB method does not require this assumption but does require an external image frequency filter at the input of the DUT. SSB noise figure is used in most applications where the desired information is contained only in the intermediate frequency and the image frequency is rejected. If the DSB measurement method is employed, 3 dB must be added to the measured noise figure to arrive at the SSB noise figure number.

MEASUREMENT TECHNIQUES

Gain, Return Loss, and P_{1dB}

Gain, return loss, and P_{1dB} of the LNA can be measured using a standard scalar S parameter test set which includes a signal generator (e.g., HP8350), a scalar network analyzer (e.g., HP8757) with detector and directional bridge, a two-way splitter, and a variable attenuator. *Figure 7* shows the setup.

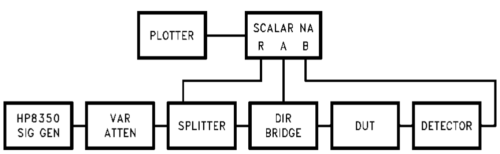

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FIGURE 7. Gain, RL, P_{1dB} Measurement Setup

After the signal generator is set to sweep over the desired frequency range and the variable attenuator at the desired value, the system can be calibrated using standard short and open terminations. Once calibrated, gain measurement can be obtained by setting the scalar analyzer to display the corrected (memory subtracted) channel B power divided by channel R power. This method allows the calculation to remain valid as the signal generator output power is changed. The variable attenuator value must be set such that the input power into the DUT is far (at least 10 dB) below the expected 1 dB compression point so that the DUT is operating in its linear region. To measure input return loss, the analyzer should display the corrected channel A divided by channel R power. Most analyzers allow dual channel displays, in which case, gain and return loss can be obtained in one plot. P_{1dB} can be obtained by gradually decreasing the attenuator value until the observed gain is 1 dB below the linear gain.

Intercept Point (OIP_3)

Output third order intercept point measurement requires two signal generators, a combiner, and a spectrum analyzer. The input of the DUT is a sum of two continuous wave RF signals at Δf apart, combined by the combiner, and the output is displayed on the spectrum analyzer. The power level of the two input signals are such that the DUT is operating in the linear range, and Δf is about a few hundred kHz or a few MHz. The intercept point is obtained by dividing the measured power level difference between the fundamental and the third order mixing product components (denoted by D in *Figure 8*) by 2 and adding the result to the power level of the fundamental component (P_0). The frequency spectrum observed on the spectrum analyzer may look similar to that illustrated by the *Figure* below.

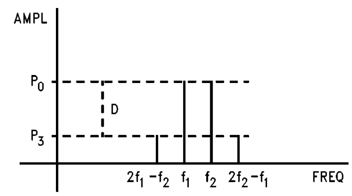

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FIGURE 8. Third Order Intermodulation

As shown above, the output third order intercept point is given by the following equation:

$$OIP_3 = P_0 + \frac{D}{2} \qquad (20)$$

This equation is a direct result of the fact the third order products grow three times faster than the fundamental term, as mentioned earlier.

CONCLUSION

Basic theory and operation of low noise amplifiers and mixers have been presented, together with the most important figures of merit and measurement methods. Also discussed were fundamental concepts on noise in electrical systems, particularly how it is generated and measured as applied to low noise amplifiers.

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