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Fundamentals of Modern RF Wireless Receivers

A Short Tutorial

To be compliant with multistandard applications, the RF front end of a modern transceiver must satisfy several challenging tasks such as large operative bandwidth, low noise, and high linearity. Over the years, addressing such requirements has significantly changed the radio architecture toward an ultimate solution based on current mode signal processing and passive mixers. In this article, after a brief description of the typical structure of a wireless receiver, voltage and current mode signal processing will be compared showing why a fully current-mode approach is more suitable in deep-scaled CMOS technologies [1], [2].

The article is divided in four sections. In "Structure of a Wireless Receiver," the structure of a wireless receiver is presented starting form the analysis of the typical operative scenario present at its input. In "Characterization of a Wireless RX," some metrics are introduced to characterize the performance of the receiver, while in "Voltage Versus Current Mode Front Ends," voltage and current mode signal processing are compared. The article ends with some examples of current-mode building blocks.

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1. Structure of a Wireless Receiver

The typical input of a wireless receiver is reported in Figure 1. The portion of spectrum allocated for the standard is called RX band and contains several channels. One of the channels hosts the wanted signal while the others can be occupied by interferers called in-band blockers. All the other signals outside the RX band are called out-of-band-blockers and are interference coming from other RF sources not regulated by the standard.

In order to allow an easier analog-to-digital conversion, the analog front end of the radio processes the input signal performing three main tasks: amplification,

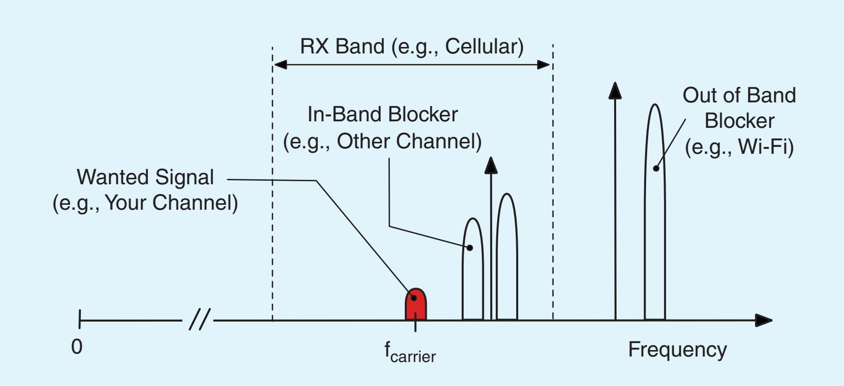

FIGURE 1: Input signal of a wireless receiver.

down-conversion, and filtering. The down-conversion is the operation of moving the wanted signal, which is modulated around a high-frequency carrier f_c, to a lower frequency. Among these three tasks, to filter in-band and out-of-band blockers is generally the most challenging operation and is crucial to relax the performance required to the analog-to-digital converter for the digitalization of the wanted signal. In the following, these three tasks will be discussed identifying the key building blocks for each operation.

Signal Amplification

Since the received signal is very small, the first blocks of the receiver chain is typically an amplifier called *low-noise amplifier* (LNA). The aim of the LNA is to amplify the input signal adding a minimum quantity of noise.

Since the LNA is the first block of the RX chain, it is also the interface with the external components of the radio. For this reason, the LNA input impedance must match with the one provided by the external element preceding it (e.g., the antenna). This impedance is indicated with RS and its value is typically 50 Ω.

Signal Down-Conversion

Once the wanted signal has been amplified by the LNA, it is shifted from RF to a lower frequency. This operation is performed by multiplying the RF signal with a sinusoid called local oscillator (LO). As shown in Figure 2, the multiplication of the wanted signal by the LO corresponds to a convolution in the frequency domain that generates four contributions: two centered at the sum of the carrier frequency (f_c) and the LO frequency (f_{LO}) and two at their difference. The frequency f_{LO} is chosen very close to f_c to have the contributions generated at the difference close to zero (i.e., at $f_c - f_{LO}$). Notice that, regardless the position of the wanted signal at RF, to tune the LO allows to down-convert the signal always at the same difference $f_c - f_{LO}$. This significantly simplifies the design of the blocks following the mixer.

Although the multiplication with a sinusoid is very effective to move the RF wanted signal around dc, there is another signal that, if convoluted with the LO, generates two contributions exactly at the same frequency of the down-converted wanted signal [Figure 3(a)]. This signal is called image and before the down-conversion is located symmetrically to the wanted signal with respect of the LO frequency. To avoid this overlap, which could degrade our reception, the modern receivers adopt the Hartley architecture drawn in Figure 3(b). The Hartley architecture exploits the asymmetry of the spectrum between the sine and the cosine functions performing a dual paths down-conversion scheme. The path with the cosine multiplication is called in-phase path (I) while the other one is called quadrature path (Q). As shown in Figure 3(b), after the sine multiplication, a 90° shift is required to put in phase the two paths before the recombination. When the outputs of the two paths are summed together, the wanted signal is preserved and the image rejected.

Filtering In-Band and Out-of-Band Blockers

Since the wanted signal is modulated around a high-frequency carrier, to filter in-band and out-of-band blockers before the down-conversion is very challenging because it would require a very narrow-band filter centered on f_c. The quality factor of such filter should be generally greater than 500. Unfortunately, in the GHz range, resonant filters achievable with a modern CMOS processes do not exceed a quality factor of 20. For this reason, out-of-band and in-band blockers are filtered through a two-step strategy, the formers at RF and the latter after the signal down-conversion (Figure 4).

To filter an out-of-band blocker at RF, an external surface acoustic wave

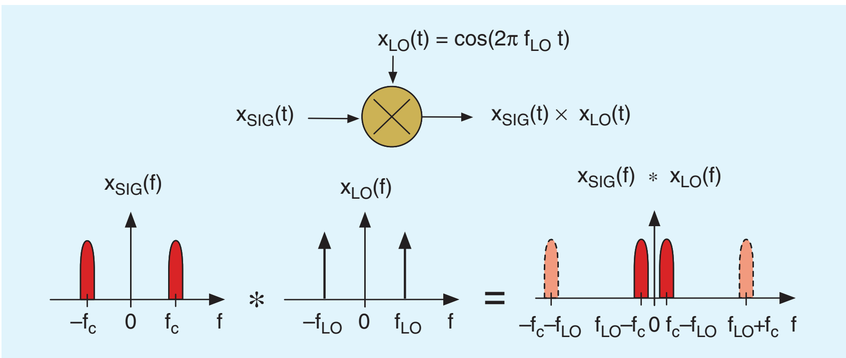

FIGURE 2: RF signal down-conversion.

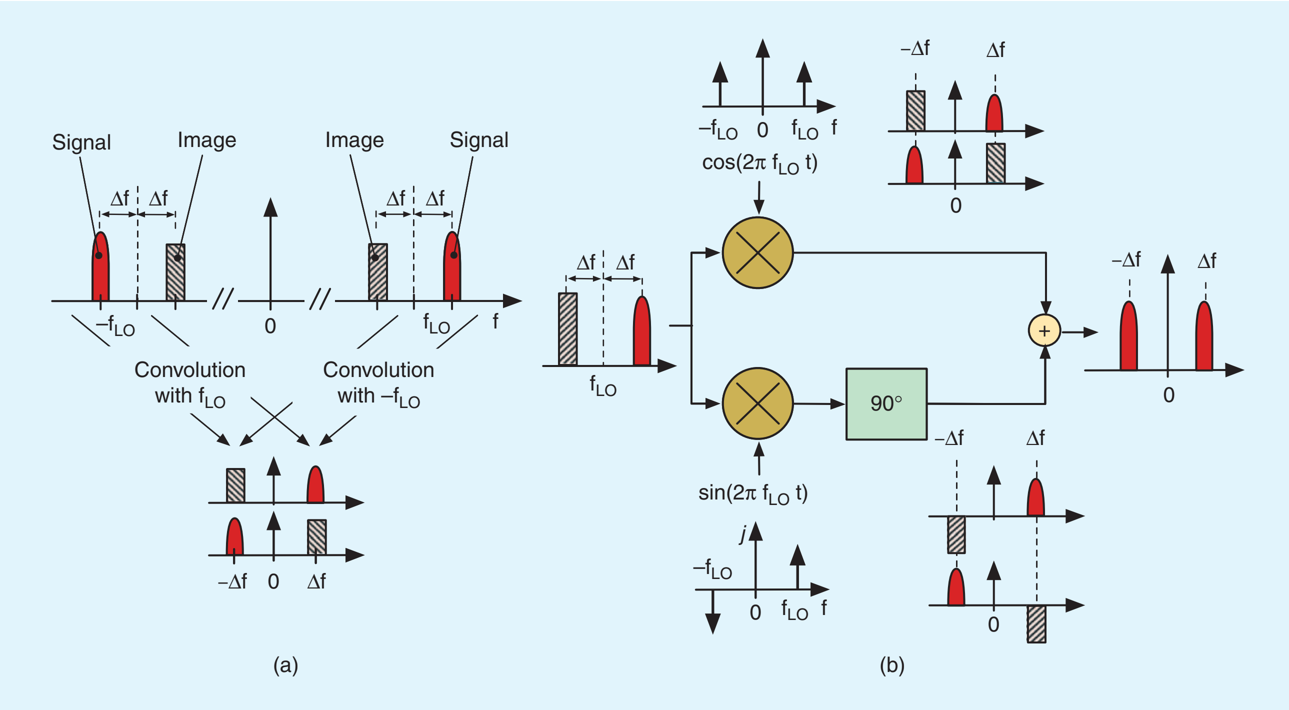

FIGURE 3: (a) Down-conversion of the image signal and (b) rejection of the image.

(SAW) filter is interposed between the antenna and the LNA. The SAW filter is based on an acoustic resonant cavity that leads to very high quality factors. However, this filter is not tunable, and so it can be used only to filter those interferers that are outside the band reserved to the standard (i.e., the out-of band blockers).

The in-band blockers are filtered after the down-conversion to avoid the need of a tunable high quality factor filter. This operation is realized by the channel selection filter following the mixer. Although in-band blockers cannot be filtered before the mixer, to filter them after the down-conversion has a great advantage. Since the down-converted wanted signal is always located at the same frequency ($f_c - f_{LO}$) independently on the channel used at RF, the channel selection filter doesn't need to be tunable.

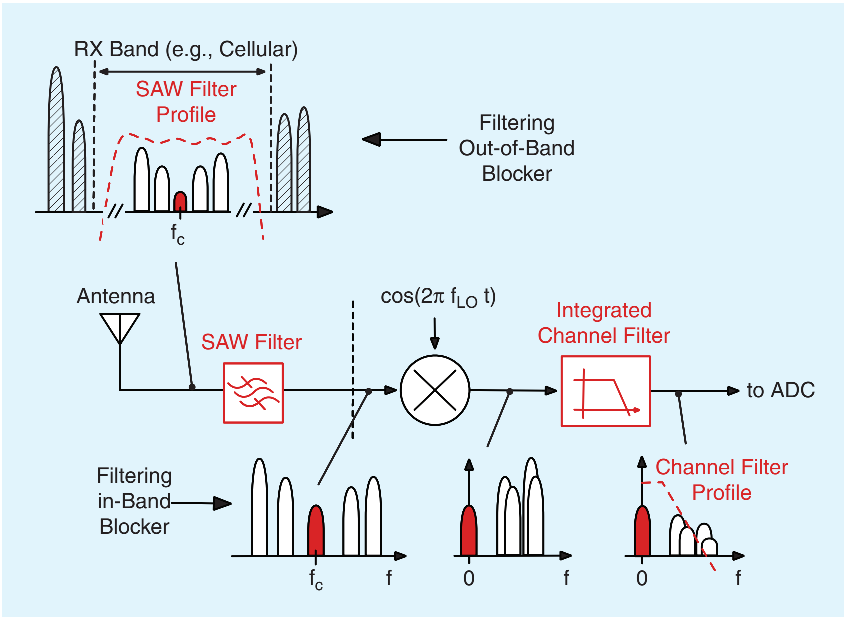

FIGURE 4: Two-step filtering strategy.

Structure of the Receiver

The complete structure of the receiver, obtained combining the building blocks used to perform the tasks of amplification down-conversion and filtering, is drawn in Figure 5.

The antenna is followed by the SAW filter used to attenuate the out-of-band blocker. Hence, the LNA amplifies the RF signal and provides an input impedance that matches with the SAW filter. The LNA precedes the I and Q mixers driven by the local oscillator, which supplies cosine and a sine signals by a quadrature generation scheme. Finally, the channel selection filter and the ADC follow the mixer.

The scheme in Figure 5 differs from the Hartley architecture shown in Figure 3(b) since two elements are missing: the 90° shift, and the recombination between I and Q paths. The reason is that, generally in integrated

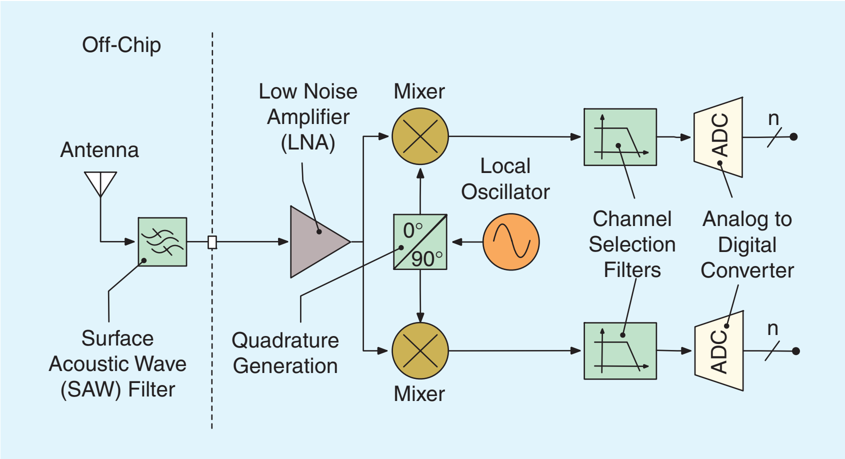

FIGURE 5: Typical structure of a wireless receiver.

solutions when the channel is shifted around DC, phase shift and recombination are done in digital domain after the ADC, where to realize these operations is simpler.

The structure in Figure 5 can be used for two different kinds of receivers, the direct conversion architecture, where the signal is downconverted exactly around DC (i.e., $f_{LO} = f_c$), and the low-IF architecture, where $f_c - f_{LO}$ is only slightly above DC and in-band blockers can still be rejected using a low-pass filter instead of a band-pass one.

2. Characterization of a Wireless RX

Now that the radio architecture has been defined, it is important to introduce some metrics in order to characterize the performance of the wireless receiver. In this section, four metrics will be introduced to quantify the quality of the input matching, the amount of noise and distortion introduced by the receiver, and the spectral purity of the local oscillator.

Input Matching (S_{11})

The quality of the impedance matching realized by the LNA can be evaluated using the magnitude of the reflection coefficient S_{11}. The S_{11} is defined as

$$S_{11} = \frac{Z_{in,LNA} - R_s}{Z_{in,LNA} + R_s}, \qquad (1)$$

where $Z_{in,LNA}$ is the input impedance of the LNA and RS the driving impedance. The magnitude $|S_{11}|$ represents the ratio between the amplitude of the reflected wave compared to the amplitude of the incident wave. When no reflected way is present the matching is perfect and $|S_{11}| = -\infty$ dB. A reasonable good matching is obtained when $|S11| < -10$ dB, which means that more than 90% of the power is transferred from the driving stage to the LNA.

Noise

The noise of the receiver is characterized in terms of noise factor (nf) and noise figure (NF) defined as

$$nf = \frac{SNR_{in}}{SNR_{out}}, \qquad NF = 10\mathrm{Log}_{10}(nf) \qquad (2)$$

where SNR_{in} is the signal-to-noise ratio at the input of the receiver and SNR_{out} is the signal-to-noise ratio at the output of the receiver. NF and nf represent the excess of noise introduced by the receiver and can be also expressed as function of the noise coming from the antenna (N_{Rs}) and the noise added by the receiver itself reported at its input ($N_{RX,in}$):

$$nf = 1 + \frac{N_{RX,in}}{N_{Rs}}. \qquad (3)$$

The minimum nf achievable is equal to 1 when no noise is added (i.e., $N_{RX,in} = 0$), having $SNR_{out} = SNR_{in}$. While $N_{RX,in}$ depends on the receiver implementation, N_{Rs} is fixed and has a power spectral density equal to kT (i.e., –174 dBm/Hz), where k is the Boltzmann's constant and T the operative temperature expressed in Kelvin. The noise figure of the receiver sets the minimum power that the wanted signal must have to be detectable. This power is called sensitivity and is given by

$$P_{sens}|_{dBm} = -174 + 10Log_{10}B + NF + SNR_{min}|_{dB}.$$

The first two terms represent the noise N_{Rs} integrated in the bandwidth B of the wanted signal , while SNR_{min} is the minimum signal-to-noise ratio required at the output of the receiver to demodulate the wanted signal.

Distortion

In addition to the noise, the analog section of the receiver can distort the input signal due to the presence of nonlinearities in the signal transfer function. The distortions can be divided in two categories: hard distortions and weak distortions. The former involves signals that explore widely the input-output characteristic, and the latter are generated when signal swing around the operative point is limited. This article will focus on the weak distortions that, in presence of blockers, can produce undesired components overlapped to the wanted signals.

The weak distortions can be studied starting from a Taylor expansion of the input–output characteristic around the operative point:

$$y(t) = a_o + a_1 x + a_2 x^2 + a_3 x^3 + \cdots \qquad (4)$$

where y(t) and x(t) are, respectively, the output and the input signals. When the characteristic described by (4) is explored by two tones, with same amplitude (A) but different frequencies (ω_1 and ω_2), several components are generated at the output. Among them, the following four contributions will be considered:

$$a_1 A(\cos(\omega_1 t) + \cos(\omega_2 t)) \qquad (5)$$

$$\frac{3}{4}a_3A^3(\cos(2\omega_1-\omega_2 t)$$
$$+\cos(2\omega_2-\omega_1)t). \qquad (6)$$

The first two contributions are at the same frequency of the input signals and are called fundamental tones. They differ from the input by the factor a_1, which is the linear gain in our receiver. The other two contributions, proportional to a_3 and located at $2\omega_1$-ω_2 and $2\omega_2$-ω_1, are called third-order intermodulation products (IM3).

From the output spectrum plotted in Figure 6(a), it is possible to notice that fundamental and IM3 tones are equally spaced in frequency. This means that if at the input of the receiver the wanted signal and two large blockers are equally spaced, the IM3 generated by the two blockers will fall above the wanted signal, causing a deterioration of the SNR.

When the output amplitude is plotted versus A in a log-log plot [Figure 6(b)], the fundamental amplitude grows with a slope of 1, while IM3 grows with a slope of 3 [due to the A^3 factor present in (6)]. Although a_1 in typically much greater than a_3, due to the different slope, fundamental and IM3 curves could intercept each other in a point called third-order intercept point (IP3). Figure 6(b) shows that generally the two curves do not intercept each other since they bend for large amplitude due to the presence of hard distortions. Hence the IP3 is typically an extrapolated point. The input amplitude for which it has the intercept point is given by

$$A_{\mathrm{IIP3}}=\sqrt{\frac{4a_1}{3a_3}}. \qquad (7)$$

The amplitude A_{IIP3} depends only on a_1 and a_3 and so it is a good parameter to characterize the third-order nonlinearity of the receiver. The IM3 generated by two blockers that can corrupt the wanted signal can be evaluated directly by the IIP3 using following equation:

$$\mathrm{IM3}\,|_{\mathrm{dBm}}=3P_{\mathrm{blocker}}\,|_{\mathrm{dBm}}-2\mathrm{IIP3}\,|_{\mathrm{dBm}} \qquad (8)$$

where the power of the two blockers (P_{blocker}) has been assumed equal for simplicity.

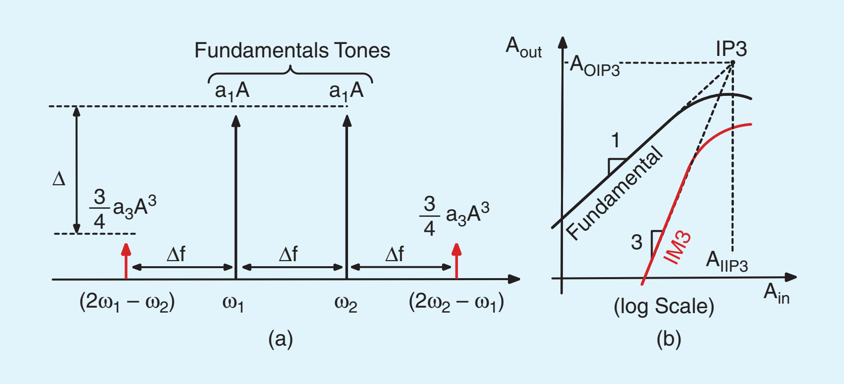

FIGURE 6: (a) Output spectrum due to a third-order nonlinearity and (b) IP3 definition.

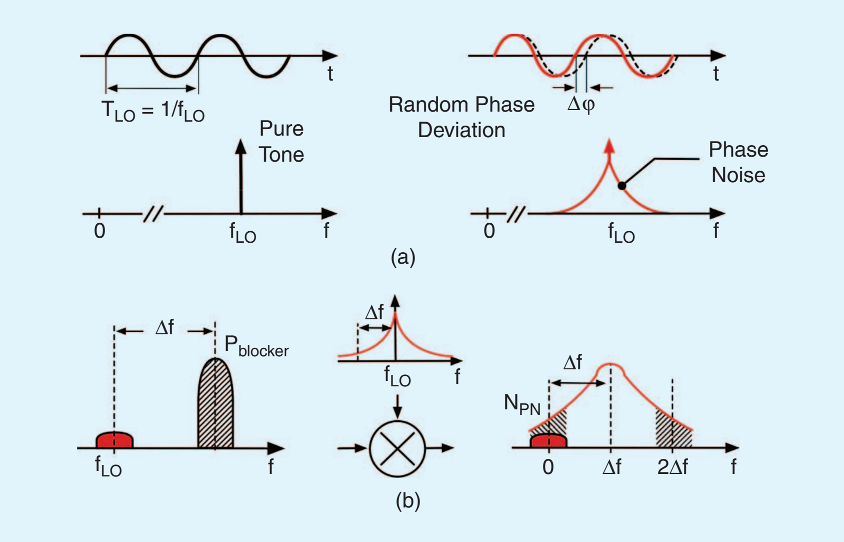

FIGURE 7: (a) Phase noise and (b) reciprocal mixing.

Spectral Purity of the Local Oscillator

The local oscillator used to downconvert the wanted signal to a lower frequency in theory should be a pure tone. However, due to the presence of noise, the sinusoid has a random deviation of its phase that in the frequency domain appears as a skirt of noise around a carrier [Figure 7(a)]. This noise is characterized in terms of phase noise obtained normalizing the power of the noise to the power of the carrier. The phase noise is measure in dBc (i.e., dB carrier) and its spectral density in dBc/Hz.

A large phase noise around the LO can affect the detection of the wanted signal when a large blocker is present at the input of the mixer. As shown in Figure 7(b), the large blocker is convolved with the phase noise producing an additional noise contribution above the wanted signal. This phenomena is called reciprocal mixing, and the amount of noise added is given by

$$N_{\mathrm{PN}}\,|_{\mathrm{dBm/Hz}}=P_{\mathrm{blocker}}\,(f_{\mathrm{LO}}+\Delta f\,)|_{\mathrm{dBm}}$$
$$+PN(\Delta f\,)\,|_{\mathrm{dBc/Hz}}, \qquad (9)$$

where P_{blocker} is the power of the blocker, f_{LO} the LO frequency, Δf the frequency offset from the carrier.

3. Voltage Versus Current-Mode Front Ends

In this section, voltage and current-mode signal processing will be compared, highlighting the reasons why the modern designs are moving toward fully current-mode solutions. The effectiveness of these two approaches will be evaluated on the three main tasks performed by the receiver: RF amplification, down-conversion, and filtering.

Voltage Versus Current Mode in RF Amplification

Since the amplification in the RF domain is realized by the LNA, voltage or current-mode operation is set by the ratio between the output impedance of the LNA and the input impedance of the mixer. In voltage mode, the LNA drives the mixer as a voltage source developing a voltage gain at RF. On the contrary, in current mode approach, the LNA inject an RF current into the low input impedance of the mixer and no RF voltage gain is developed. Since conventionally the LNA is assumed driven by a voltage source, in current mode operation is also called low noise transconductor amplifier (LNTA).

The main limit of the voltage mode solutions relies on the generation of the RF voltage amplification, which is challenging due to the presence of parasitic capacitances at the output of the transistors. In order to increase the gain achievable, the capacitances can be resonated with integrated inductors. However, this strategy has two main drawbacks. The first one is that integrated inductors need large areas, increasing the cost of the design. The second problem is that a resonant load cannot be used in a wideband receiver or must be tuned if several standard must be covered (e.g., in multistandard applications). The only advantage in the use of a resonant load is that out-of-band blockers are mildly filtered relaxing the linearity requirements of the radio.

Voltage Versus Current Mode in Down-Conversion

Voltage and current mixers are very similar. In both cases, the input signal is multiplied with the LO by using transistors acting as switches (Figure 8). The input and the output of the mixer are alternatively connected and disconnected, resulting in a multiplication by a square-wave, with a period equal to $1/f_{LO}$. Only the first harmonic of the square wave is used to down-convert the signal obtaining a conversion gain equal to $2/\pi$.

During the conduction phase, when the receiver is implemented in a CMOS technology, the switches can operate in triode region (passive mixers) or in saturation region (active mixers). Since the CMOS technologies evolve optimizing the transistor in the triode region, passive mixers have become the natural choice for the modern design.

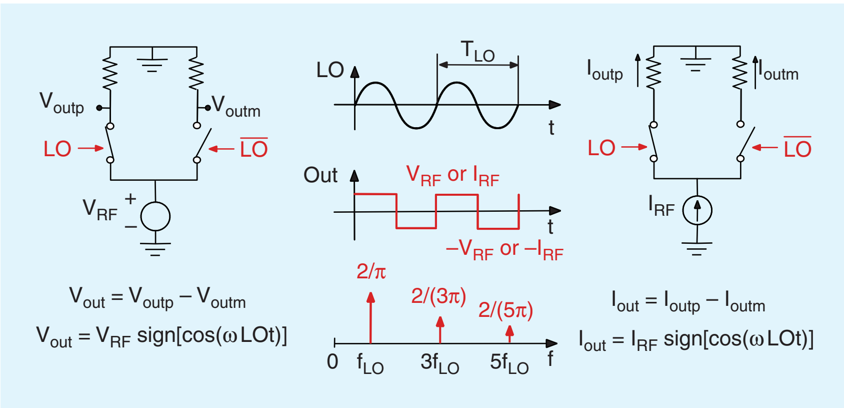

FIGURE 8: Voltage and current-mode mixers.

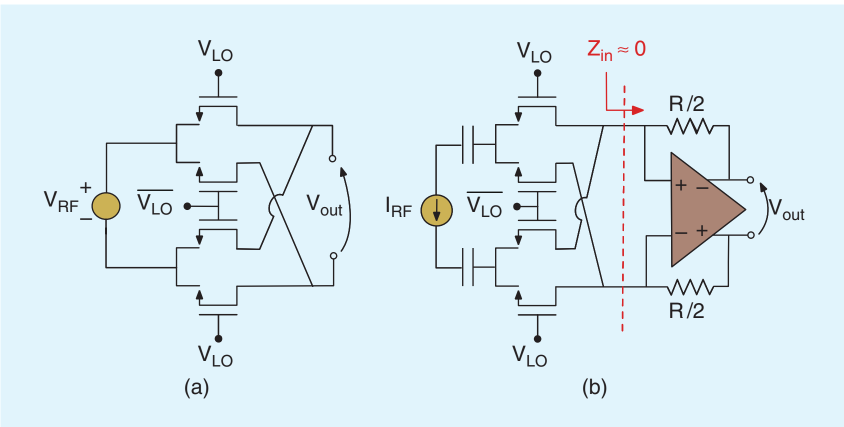

FIGURE 9: Fully differential (a) voltage and (b) current-mode passive mixers.

The CMOS implementations of fully differential voltage and current mode passive mixers are seen in Figure 9. In both cases, the input signal is connected to the output through four switches. Using a two-phases LO, the polarity of the input signal is reversed half of the time, having a multiplication by a square wave. In the conducting phase, a low switch resistance is assured, driving the gate with a large LO amplitude and limiting the voltage swing at the source and at drain of the transistors.

The limitation of the voltage swing at the input of the mixer is a severe constraint in the voltage mode implementation where a voltage gain is developed at RF [Figure 9(a)]. On the contrary, the current mode-passive mixer in Figure 9(b) does not suffer from this issue since the signal is sensed by a trans-impedance amplifier (TIA), which provides a low input impendence assuring a small voltage swing also in presence of large RF current signals [3].

Voltage Versus Current Mode in Signal Filtering

Since, after the down-conversion, the wanted signal is still relatively small and surrounded by large in-band blockers, the channel selection filter should be a low-pass filter with low noise in the signal passband and high linearity in the filter stop-band.

At low frequency, filters are realized by using capacitors as reactive elements. Since the capacitor is an integrator for the charge, the best way to realize a low-pass filter in voltage mode is to sense the output signal across the capacitor [Figure 10(a)]. Unfortunately, the capacitor also integrates the noise charge generated by the resistor and when the signal is sensed across its terminal, this noise is added to it. This is because the noise and input signal in Figure 10(a) have the same transfer function to the output.

In the current mode filter drawn in Figure 10(b), the RC network is fed by a current source and the signal is not sensed across the capacitor but as a current coming out from the resistor. The filter transfer function is still low pass because the capacitance provides a high impedance at the low frequency while at high frequency drains the majority of the input signal. However, in this case the noise injected by the resistor has a high-pass transfer function to the output since at the low frequency the capacitor is an open circuit and there is no path for the noise current to flow out from the resistor. On the contrary, at high frequency, the low impedance used to filter the input signal lets the noise coming out form the resistor. This intrinsic high-pass shaping of noise, and also of the distortions [4], makes the current more approach more suitable for the implementation of the channel selection filter.

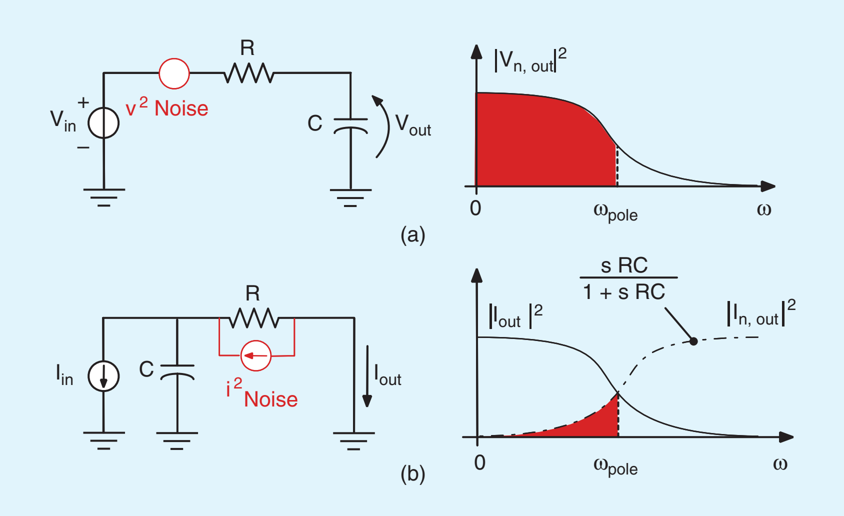

FIGURE 10: First-order low-pass filter in (a) voltage and (b) current mode.

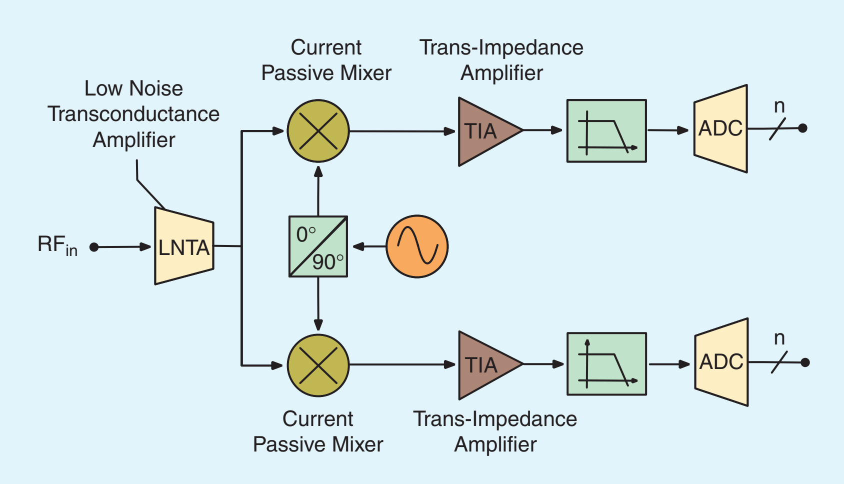

FIGURE 11: Structure of a current-mode wireless receiver.

4. Current-Mode Architecture and Building Blocks

From the considerations of the previous section, the architecture of the modern RF wireless receivers evolved in the one drawn in Figure 11. An LNTA drives I and Q current-mode passive mixers followed by a TIA used to sense the down-converted current. After that, the signal is processed by the cascade of the channel selection filter and the ADC. In the following some considerations regarding LNTA, mixer and TIA implementations will be discussed in details.

Current Mode LNTA

An efficient implementation of the LNTA relies on how the input matching is performed. An easy way to provide the impedance matching is to add in parallel to input of the transconductor a resistor R_s [Figure 12(a)]. The noise factor of such LNTA can be evaluated starting from the definition given by (3) assuming three noise contributors: the noise coming from the antenna (i.e., $N_{Rs} = 4kTR_s$), the noise of the resistor used to synthesize the matching, and the noise of the transconductor (assumed equal to $4kT/g_m$, where g_m is the gain of the transconductor). The expression for nf is given by

$$nf_{\text{LNTA}} = 1 + 1 + \frac{4}{g_m R_S} > 3\text{dB}. \quad (10)$$

In (10), the first term comes directly from the definition used in (3), the second one is given by the resistor R_s used to synthesize the matching, and the third one by the transconductor. Since to synthesize a matching with the source we used a resistor equal to R_s, the noise added is equal to the noise N_{Rs} and so the nf cannot be smaller than 2 (i.e., NF > 3dB).

Among the feed-forward solutions, one that is very popular is the boosted common gate LNTA.

The only way to reduce the impact of the noise of the input impedance below 1 is to insert a feedback or a feed-forward path around the transconductor. As will be shown in the examples reported in the following, feedback and feed-forward can reduce only the noise of the matching network, while the only way to reduce noise of the transconductor is to amplify the signal in front of the transconductor itself.

The use of feedback is more complicated in current-mode approach since no voltage gain is developed at the output of the LNTA. This explains why the use of feed-forward approach has become so popular in current-mode receivers. In the following the three example of LNTA reported in Figure 12 will be discussed.

Inductive Degenerated LNTA

One of the feedbacks techniques that does not involve the output of the transistor is source degeneration. In particular, in the LNTA drawn in Figure 12(b), the input matching is realized by an inductive degeneration where the inductance L_s is transformed into a resistance equal to $\omega_t L_s$ (where $\omega_t = g_m/C_{gs}$) [5]. This is possible since the current injected by the transistor is in quadrature with the input current flowing through C_{gs}. The network is completed adding an inductance L_g at the gate of the transistor in order to resonate the reactive part of the input impedance given by C_{gs}. The obtained series resonance provides also an amplification of the input signal at the input of the transconductor proportional to the quality factor (Q) of the resonance.

In first approximation the noise factor of this amplifier is given by

$$nf_{\mathrm{LNTA}} = 1 + \frac{\gamma}{Q^2 g_m R_S}, \qquad (11)$$

where for the transistor has been assumed a noise source reported at its gate equal to $4kT\gamma/g_m$. Compared to (10), the term associated to the input termination disappeared, since R_s is synthesized starting from an inductor that is in theory noiseless. In addition to that, the noise of the transconductor is divided by Q^2, taking advantage of the signal amplification produced by the series resonance.

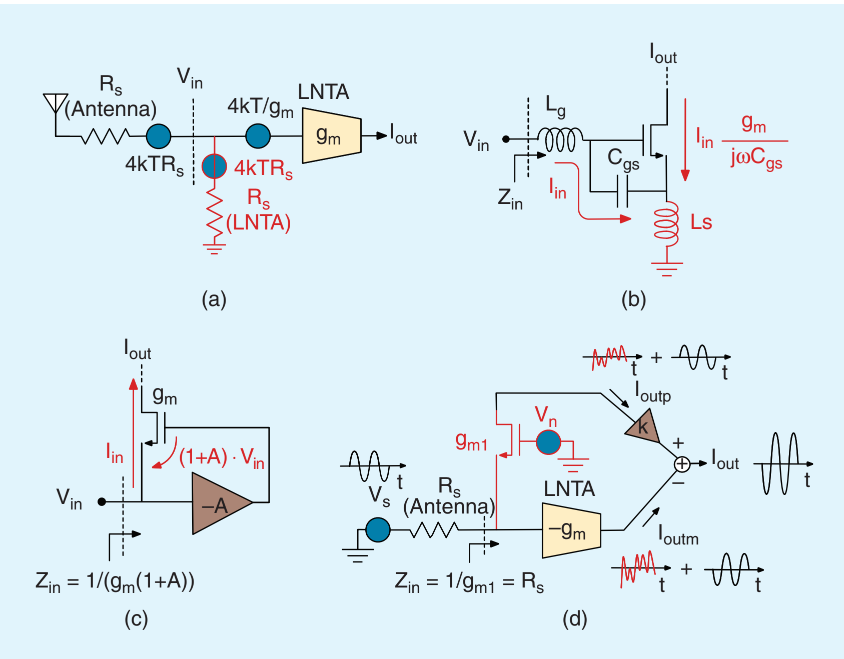

FIGURE 12: LTNA: (a) simple resistive termination, (b) inductive degeneration, (c) boosted gm, and (d) noise cancelling.

This LNTA is one of the less noisy available in literature; however, there are two drawbacks. The inductive degenerated amplifier is intrinsically narrowband since relies on a series resonance whose implementation requires two inductors (in the GHz range generally L_g is external).

Boosted Common Gate LNTA

Among the feed-forward solutions, one that is very popular is the boosted common gate LNTA reported in Figure 12(c) (many variations of the theme are present in literature). In this case the input signal is injected into the source of a transistor using the transconductor itself to synthesize the matching. In addition, an amplification (A) is applied between the source and the gate to minimize the noise of the transistor. This voltage amplification can be noiseless if realized by a reactive element as transformer [2], [6].

The presence of the feed-forward path A transforms the input impedance of the transistor in $1/(g_m(1+A))$ that must be set equal to R_s. The noise factor for this LNTA is given by

$$\begin{aligned} nf_{\mathrm{LNTA}} &= 1 + \frac{\gamma}{(A+1)^2 g_m R_S} \\ &= 1 + \frac{\gamma}{(A+1)}. \end{aligned} \qquad (12)$$

This expression looks very similar to (11). Only two contributions are present since the input impedance is realized with the transconductor itself without the addition of any noisy components. The noise produced by the transistor is divided by $(A+1)^2$ thanks to the amplification by $A+1$ between gate and source. However in this case, the matching condition forces $g_m = 1/(R_s(A+1))$ leading to

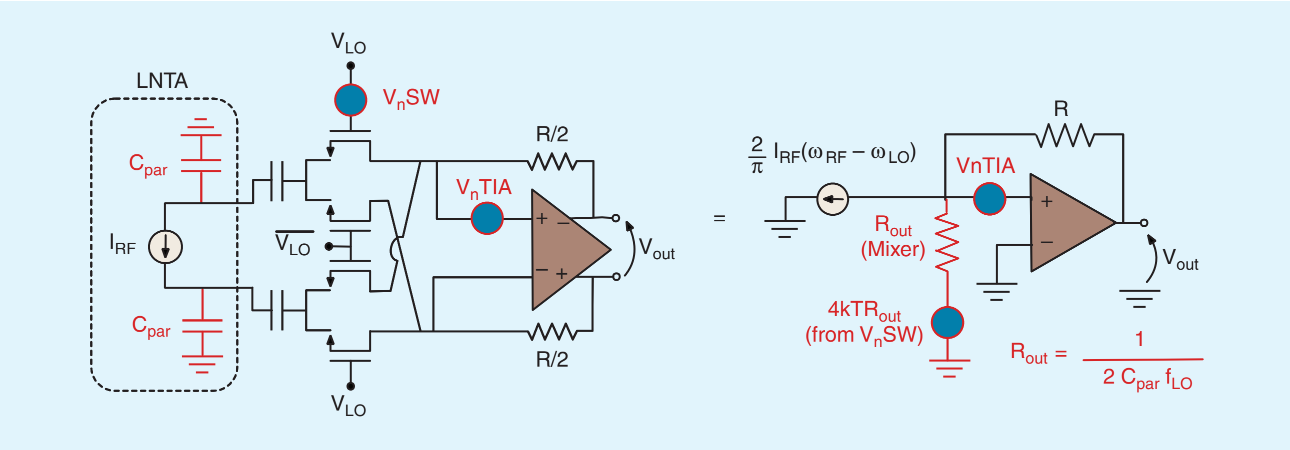

FIGURE 13: Equivalent circuit for current-mode mixer.

an overall noise factor that is only inversely proportional to $A+1$.

This LNTA is wideband, and it can have a noise factor smaller than 2. However, it suffers from a very low transconductance gain that is equal to $1/R_s$.

Noise-Cancelling LNTA

The last LNTA presented is the noise-cancelling amplifier shown in Figure 12(d) [7], [8]. In this case, the input impedance of the LNTA is realized with a transistor in common gate configuration. For a proper matching its transconductance must be set equal to $1/R_s$.

The common gate stage creates an additional feed-forward path in parallel to the main path given by the transconductor g_m. The noise produced by the common gate stage has an inverting transfer function in both paths. On the contrary, the input signal has an inverting transfer function only in the main path. This makes it possible to cancel the noise of the common gate stage by subtracting the signal between the two paths, after a proper renormalization of the gain. The noise factor for this LNTA is given by

$$nf_{\mathrm{LNTA}} = 1 + \frac{\gamma}{g_m R_S}. \qquad (13)$$

As in the previous LNTA, the noise factor can be smaller than 2 since the noise of the input termination has been cancelled. However, compared to the inductive degenerated LNTA the noise of the transconductor is not reduced since there is no signal amplification in front of it.

Current-Mode Passive Mixer

The key element in the design of a current-mode passive mixer [Figure 9(b)] is the parasitic capacitance (C_{par}) present at its input. The value of this capacitance is the main limitation in the noise performance and in the output impedance provided by the mixer.

To study the noise and the interaction of the mixer with the TIA, the current-mode passive mixer can be modeled with a Thevenin equivalent circuit having as current source the down-converted signal normalized by the conversion gain $2/\pi$ and as output impedance a resistor $R_{out} = 1/2C_{par}f_{LO}$ (Figure 13) [9].

In the noise analysis, the resistance R_{out} plays two different roles. The noise injected by the resistance R_{out} models the noise produced by the switches [10]. In addition to that, the

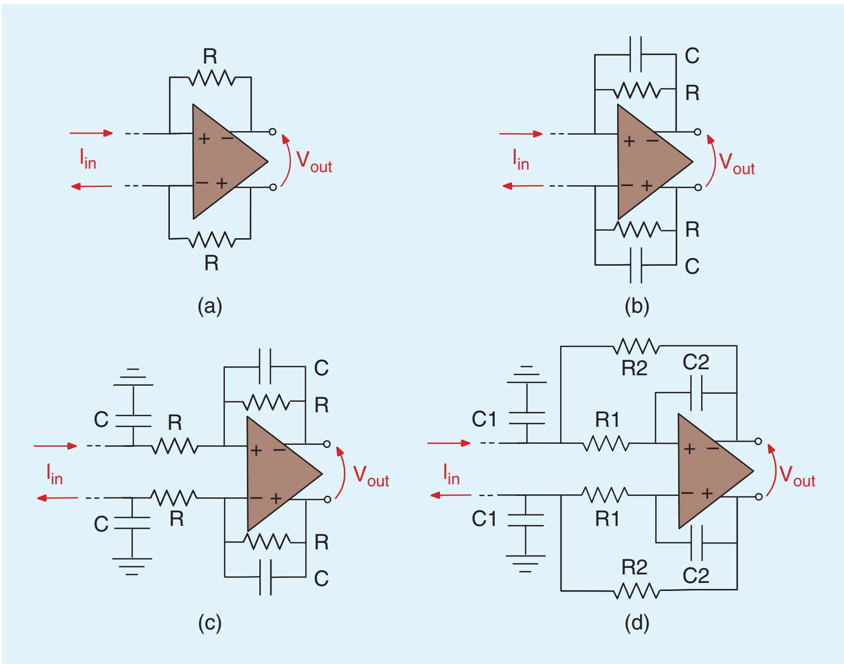

FIGURE 14: (a) Simple TIA, (b) first-order TIA, (c) second-order TIA with real poles, and (d) second-order TIA with complex poles.

presence of the fine impedance R_{out} amplifies the noise injected by of TIA, as it is possible to verify from a quick analysis of the circuit in Figure 13.

Notice that, although the finite output impedance of the mixer affects the noise performance of the TIA, it doesn't produce any loss on the signal transfer function because, from the point of view of the signal, the TIA offers an input impedance close to zero.

Filtering Trans-Impedance Amplifier

In order to provide a very low input impedance, the TIA is generally implemented by using an operational amplifier connected in feedback by a resistor [Figure 14(a)]. However in some wide-band implementation, also a simple common gate stage is used.

If the former solution is adopted, part of the channel selection filter can be embedded in the TIA by inserting a capacitor C in parallel with R [Figure 14(b)]. In this case, the voltage signal is filtered by a first-order low-pass profile. However the RC filter adopted is ineffective to reduce the power consumption of the TIA since the amplifier absorbs the current coming from the mixer unfiltered.

In order to reduce the power consumption of the TIA, a passive RC filter can be inserted in front of it [Figure 14(c)]. In this way, the current coming from the mixer is passively filtered by the capacitance before entering in the TIA. This approach proposed in [11] wand [12] has two main drawbacks. The first one is to increase the input impedance of the TIA, and the second one is that the resulting transfer function is the cascade of two real poles, much less selective than the one obtained with a couple of complex conjugate poles.

A second-order low-pass TIA with complex conjugate poles can be obtained starting from the solution in Figure 14(c) connecting the additional RC filter into the feedback loop of the TIA as proposed in Figure 14(d) [2]. In this structure R_1, R_2 the operational amplifier, and C_2 behave as an inductor. At low frequencies C_2 is an open circuit so the feed-forward gain in very high leading to an input impedance close to zero. As soon as the magnitude of the capacitance C_2 drops, the gain diminishes and the input impedance increases causing the circuit to behave like an inductor. The combination of this inductor with C_1 produces a second-order transfer function with complex conjugate poles.

Conclusions

In this short tutorial, the structure of the modem wireless receiver has been developed highlighting the most important tasks that the analog section of the radio must performed: amplification, down-conversion, and filtering. After that, an introduction to different metrics to characterize the receiver has been proposed. Voltage and current-mode approaches were compared, showing the reasons why modern solutions are converging toward a fully current mode design. Finally, some examples of current mode building blocks have been provided.

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