A Single-Chip 900 MHz CMOS Receiver Front-End with a High Performance Low-IF Topology

Jan Crols, *Student Member, IEEE*, and Michel S. J. Steyaert, *Senior Member, IEEE*

***Abstract*— An analog receiver front-end chip realized in a 0.7 μm CMOS technology is presented. It uses a new, high performance, downconverter topology, called double quadrature downconverter, that achieves a phase accuracy of less than 0.3° in a large passband around 900 MHz, without requiring any external component or any tuning or trimming. A high performance low-IF receiver topology is developed with this double quadrature downconverter. The proposed low-IF receiver combines the advantages of both the classical IF receiver and the zero-IF receiver: an excellent performance and a very high degree of integration. In this way, it becomes possible to realize a true fully integrated receiver front-end that does not require a single external component and which is, different from the zero-IF receiver, nonetheless totally insensitive to parasitic baseband signals and self-mixing products.**

I. Introduction

In the last few years, the world of wireless communications has been changing very rapidly [1]. One aspect of this evolution is the ever increasing level of integration of the transceivers, driven by the need for both a better portability and a lower cost prize. The consequence of this evolution is that the zero-IF downconversion topology is used more and more in receivers for mobile wireless telecommunications applications [2], [3]. The zero-IF receiver can be implemented with a much higher degree of integration than the conventional IF (heterodyne) receiver. This has made the zero-IF downconverter the preferred topology in the research towards the fully integrated single-chip transceiver.

However, the zero-IF topology has two, well known, major drawbacks that makes achieving an acceptable performance with it very hard [4], [5]. Firstly, its baseband configuration makes the zero-IF topology highly sensitive to parasitic baseband signals like dc offset voltages and self-mixing products caused by crosstalk between the RF and the LO signals. The second source of performance reduction is inherent to any analog integrated multipath topology. Excellent matching between the different downconversion paths is required but limited in analog implementations. The effects of mismatch, i.e., phase and amplitude errors, degrade the signal quality because they result in a reduced mirror signal suppression.

The use of multipath topologies is nowadays in analog applications still very limited. In digital and software applications, matching is always perfect, and multipath topologies are used widely. In these applications, two parallel signals are regarded as one complex signal on which different complex signal processing steps are performed. In this paper, it is shown how the complex signal method can be extended to the world of analog systems with its inherent imperfect matching between parallel processing paths. It is demonstrated in this paper that many more analog multipath signal operations, other than the classical quadrature downconversion and low-pass filtering, are possible. The complex signal method is used to get a clear insight in both the wanted complex signal operation and the unwanted operations caused by mismatch. With these analog complex signal operators and the complex signal analysis method for analog systems, many new receiver topologies can be synthesized. This paper discusses the different possibilities and selects one topology, named the low-IF receiver topology, as the most preferable.

The low-IF receiver is, like the zero-IF receiver, also a multipath topology that can be implemented in a highly integrated way. It uses an IF of a few hundred kHz and is therefore not sensitive to parasitic baseband signals like dc offset voltages and self-mixing products. In this way, the low-IF receiver combines the advantages of both the IF and the zero-IF receiver. It can have a high performance and a high degree of integration at the same time. The main drawback of the low-IF receivers is that they are more sensitive to a good mirror signal suppression, because, in opposition to the zero-IF receiver, here the mirror signal can be larger than the wanted signal. With the complex signal method, it is demonstrated that the phase error of the quadrature downconverter is the main limitation for the mirror signal suppression. A typical quadrature downconverter with a phase error of 3° [6], results in a maximal mirror signal suppression of 26 dB. Therefore, this paper presents a newly developed fully integrated CMOS quadrature downconverter with a phase accuracy of 0.3°, a key building block for high quality low-IF receivers. Its topology is synthesized with the complex signal method and analog complex signal operators, and it is shown how this quadrature downconverter in combination with an LNA and a synthesizer can form the complete and fully integrated analog low-IF receiver front-end.

The quadrature downconverter that is presented here is in many ways different from the conventional realizations [6]. The conventional realizations use an RC-CR network for the quadrature generation of the LO signal and two Gilbert multipliers for the downconversion. The generation of the

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The authors are with the Katholieke Universiteit Leuven, ESAT-MICAS, B-3001 Heverlee, Belgium.
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quadrature signal is based on the 90° phase-difference between a pole and zero with the same cutoff frequency. Due to its asymmetric structure is the phase accuracy of an RC-CR quadrature generator typically about 3°. Only with tuning and trimming can its phase error be made lower than 1° [7]. Here, sequence assymmetric polyphase filters are used for a high quality broad-band quadrature generation. Both the RF and the LO input signals are put in quadrature and mixed down with four CMOS downconversion multipliers of which the output signals are combined two by two. This new topology is called the double quadrature downconverter. In order to obtain an overall phase error of 0.3°, each quadrature generator of the double quadrature downconverter requires only a phase-error of 4°, resulting in no need for matching, tuning, or trimming.

The presented quadrature downconverter is realized in a 0.7 μm CMOS process, and it does not require any external component. The realization in CMOS has been chosen to prove the abilities of submicron CMOS for the integration of high performance RF circuitry. The trend to a full and low cost integration of the complete wireless system on a single chip, including both the analog and the digital part, will make CMOS a key technology for advanced RF circuitry development.

In the second part of this paper, the complex signal method for the analysis of analog multipath signal processing architectures is introduced. With this technique, the low-IF receiver is synthesized and analyzed, and the qualities of this receiver architecture towards a high degree of integration and a high performance are demonstrated in Section III. Section IV deals with the design and implementation of the double quadrature downconverter. The results of the realized quadrature downconverter are given in Section V. The downconverter has single-ended RF and LO inputs, it delivers fully differential I- and Q-signals at its output with an amplitude error smaller than 0.5 dB and a phase error smaller than 0.3° in a passband from 500–900 MHz. Its noise figure is 24 dB for a 12 dBm LO signal, and its third-order interception point (IP3) is more than 27 dBm. Its input bandwidth is as high as 900 MHz, and its output bandwidth is 3 MHz.

II. Synthesis of Analog Multipath Systems

The analysis of multipath systems is greatly simplified when at each point the set of parallel signals is treated as one signal vector. This is perfectly possible for any set of independent signals. By transforming these signal vectors to a new vectorspace, a new set of independent signals can be defined. These new signal sets give, when properly chosen, a clear insight in their frequency information, and special frequency operations can be defined on them.

A two-dimensional signal vector (i.e., a set of two independent signals, e.g., an I, Q-signal) can be represented as a complex signal

$$\begin{aligned} y(t) &= [i(t) \quad q(t)] = i(t) + jq(t) = y_r(t) + jy_i(t) \\ Y(j\omega) &= [I(j\omega) \quad Q(j\omega)] = I(j\omega) + jQ(j\omega) \\ &= Y_r(j\omega) + jY_i(j\omega). \end{aligned} \tag{1}$$

Any frequency component of a signal can be written as a sum of a positive ($e^{+j\omega}$) and a negative ($e^{-j\omega}$) signal

$$\begin{aligned} A(\omega) \cdot \cos(\omega t + \varphi(\omega)) &= \frac{A(\omega)}{2} \cdot [\cos(\omega t + \varphi(\omega)) + j \cdot \sin(\omega t + \varphi(\omega))] \\ &\quad + \frac{A(\omega)}{2} \cdot [\cos(\omega t + \varphi(\omega)) - j \cdot \sin(\omega t + \varphi(\omega))] \end{aligned} \tag{2}$$

$$\begin{aligned} A(\omega) \cdot [\cos(\omega t + \varphi(\omega)) + j \cdot \sin(\omega t + \varphi(\omega))] &= A(\omega) \cdot e^{j\varphi(\omega)} \cdot e^{j\omega} \\ A(\omega) \cdot [\cos(\omega t + \varphi(\omega)) - j \cdot \sin(\omega t + \varphi(\omega))] &= A(\omega) \cdot e^{-j\varphi(\omega)} \cdot e^{-j\omega}. \end{aligned} \tag{3}$$

Equations (1) and (3) show that a signal vector can be made that has only a positive or a negative frequency component. In a more general way, it can be said that the positive and negative frequency information content of a complex signal vector can be completely independent. It becomes thus important to always represent a complex signal vector with its positive and negative frequency information.

The possibility of creating a signal, albeit a complex signal pair, which has only a positive frequency content gives many design opportunities because the main problem in receiver design is the fact that the multiplication in a downconverter with a sinus is a multiplication with both a positive and a negative frequency. This is the well known problem of the mirror frequency. This problem is nowadays solved by using either an external high IF and an external high frequency high-Q mirror signal suppression filter (IF receiver) or by using no IF (zero-IF receiver). If there would exist a multiplying operation defined on complex signal pairs that would preserve the convolution laws, there would be a third method: a downconversion to an IF frequency by multiplication with only a positive frequency, resulting in no downconversion of the unwanted mirror signal frequency, no need for a high IF, and no need for an external high Q HF filter. Such a multiplication for complex signal pairs is given by the complex multiplication

$$\begin{aligned} z(t) &= y(t) \cdot x(t) = (y_r(t) + jy_i(t)) \cdot (x_r(t) + jx_i(t)) \\ &= (y_r(t) \cdot x_r(t) - y_i(t) \cdot x_i(t)) \\ &\quad + j(y_r(t) \cdot x_i(t) + y_i(t) \cdot x_r(t)) \end{aligned} \tag{4}$$

$$\begin{aligned} Z(j\omega) &= Y(j\omega) \otimes X(j\omega) \\ &= (Y_r(j\omega) + jY_i(j\omega)) \otimes (X_r(j\omega) + jX_i(j\omega)) \\ &= (Y_r(j\omega) \otimes X_r(j\omega) - Y_i(j\omega) \otimes X_i(j\omega)) \\ &\quad + j(Y_r(j\omega) \otimes X_i(j\omega) + Y_i(j\omega) \otimes X_r(j\omega)). \end{aligned} \tag{5}$$

Fig. 1 gives the block diagram realization of (4) and (5). It reduces to the classical quadrature multiplier when only a single signal has to multiplied with a complex signal pair. The problem with an actual analog integrated implementation of this block diagram is that (4) and (5) only hold when the conversion gains of the four mixers are perfectly matched. Perfect matching does not exist in analog circuits, and the higher the operating frequencies, the more difficult the matching becomes. A mismatch between the four conversion gains can be split up as a mismatch between the gains in the different

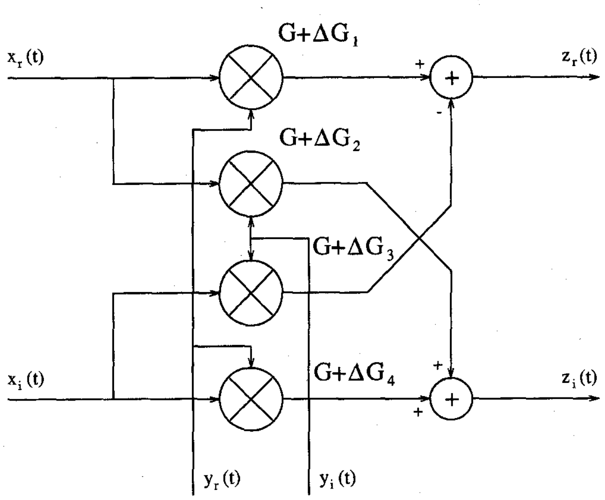

Fig. 1. Mixer topology for the multiplication of two complex signal pairs.

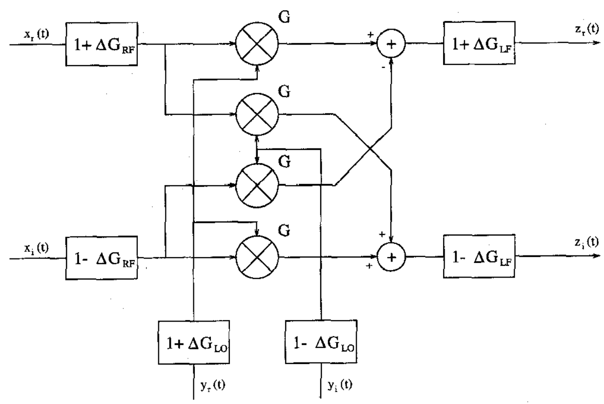

Fig. 2. Conversion gain mismatch in the complex signal multiplier and its equivalent scheme.

signal paths. These mismatches have been added in Fig. 1 as ΔG_i. The equivalent scheme is shown in Fig. 2.

A gain mismatch between two parallel signal paths is equivalent with a crosstalk between positive and negative frequencies. Fig. 3 shows this: the gain mismatch can be rewritten as the sum of two parallel signal operations, the wanted perfectly matched gain operation, and a gain operation as large as the mismatch combined with a frequency mirror operation (positive frequency components become negative frequencies and vice-versa, as proven in (6)). The crosstalk is thus equal to $\Delta A/A$. An amplitude error of 1% gives a -40 dB crosstalk between the positive and negative frequencies. For the complex mixer of Fig. 2, this results in an imperfect mirror signal suppression. The equivalent gain mismatch in the LO signal path makes the multiplication with a pure single positive frequency impossible. There is always also the multiplication with a small negative frequency component. Gain mismatch in the LF path results also in a crosstalk of the unwanted mirror signal, situated at the negative IF, to the wanted signal,

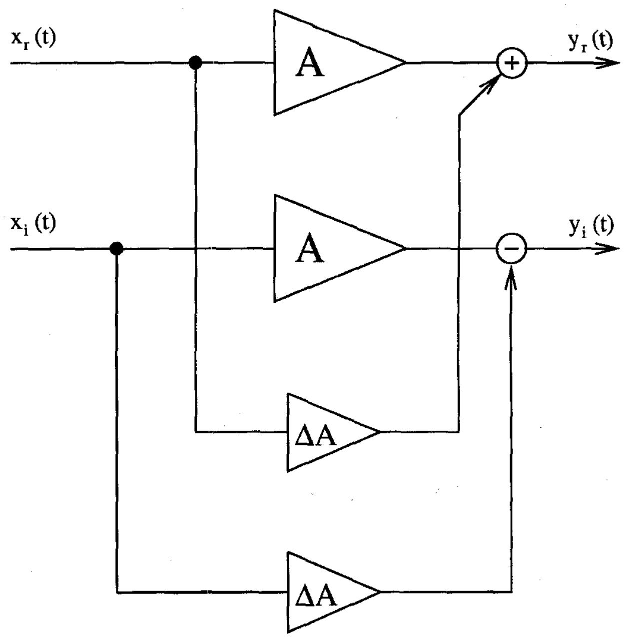

Fig. 3. Equivalent scheme for gain mismatch between two parallel signal processing paths.

situated at the positive IF

$$\begin{aligned} e^{j\omega_c t} &= \cos(\omega_c t) + j \cdot \sin(\omega_c t) \rightarrow \sin(\omega_c t) + j \cdot \cos(\omega_c t) \\ &= j \cdot e^{-j\omega_c t} \\ e^{-j\omega_c t} &= \cos(\omega_c t) - j \cdot \sin(\omega_c t) \rightarrow \sin(\omega_c t) - j \cdot \cos(\omega_c t) \\ &= -j \cdot e^{j\omega_c t}. \end{aligned} \tag{6}$$

The result of a phase error between two parallel signal paths is also an unwanted mirroring of signals to their opposite frequencies. Equation (7) gives the effect of a phase error on a positive frequency component

$$\begin{aligned} e^{j\omega_c t} &= \cos(\omega_c t) + j \cdot \sin(\omega_c t) \rightarrow \cos(\omega_c t + \Delta\varphi) \\ &\quad + j \cdot \sin(\omega_c t - \Delta\varphi) \\ &= \cos(\Delta\varphi) \cdot (e^{j\omega_c t} - j \cdot \tan(\Delta\varphi) \cdot e^{-j\omega_c t}). \end{aligned} \tag{7}$$

Apart from the wanted positive frequency, the total transfer function generates again also a negative and unwanted frequency component. The ratio between the unwanted and the wanted signal is $\tan(\Delta\varphi)$, which can be taken equal to $\Delta\varphi$ for small values of $\Delta\varphi$. This means that a phase error of 1° results in a -35 dB crosstalk between positive and negative frequencies.

Phase errors are not so important as amplitude errors in the complex mixer structure of Fig. 2 as long as its operating frequency is lower than its bandwidth. Phase errors are more dominant in filter operations performed on complex signal pairs. There are basically three different filter operations that can be performed on complex signal pairs. The first one is given in Fig. 4. Putting the same filter in the two parallel paths gives the filter transfer function for both the positive and negative frequencies. Amplitude and phase differences between the two transfer functions will cause frequency crosstalk.

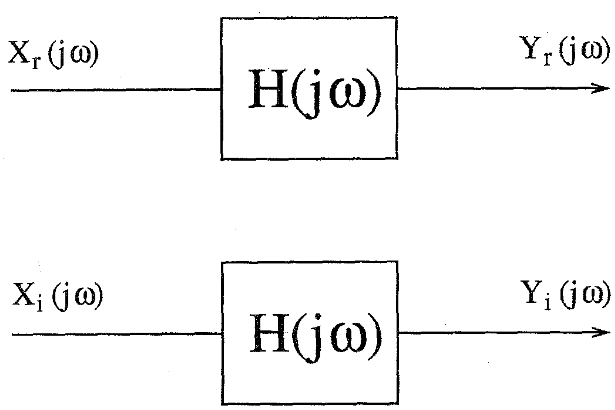

Fig. 4. The classic two-port filtering for complex signals.

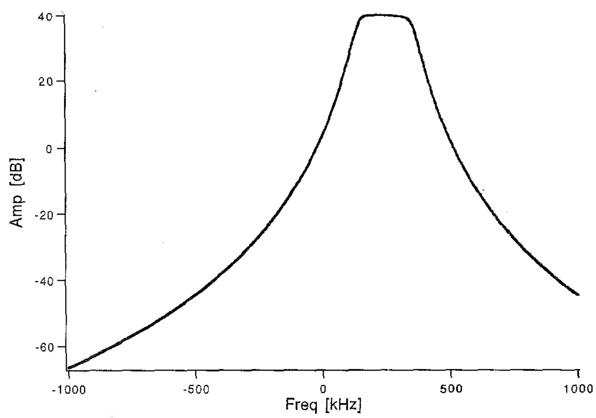

Fig. 5. The transfer function for positive and negative frequencies of an active integrated fifth-order sequence asymmetric polyphase filter.

A second class of filters for complex signal pairs are called active sequence asymmetric polyphase filters [8]. These filters are band-pass filters that have a passband for either only positive or negative frequencies. This makes them suitable for the suppression of the mirror signal after a quadrature downconversion (after quadrature downconversion the mirror signal is situated at frequencies opposite of the wanted signal). The transfer function of an active asymmetric polyphase is a linear frequency translated version of a low-pass filter. For a first-order low-pass filter is this

$$H_{lp}(j\omega) = \frac{1}{1 + j\omega/\omega_o}$$
$$H_{bp}(j\omega) = \frac{1}{1 - j\omega_c/\omega_o + j\omega/\omega_o} = \frac{1}{1 - 2jQ + j\omega/\omega_o}. \quad (8)$$

Fig. 5 shows the transfer function of a frequency translated fifth-order low-pass Butterworth filter. The linear transformation $H_{bp}(j\omega) = H_{lp}(j\omega - j\omega_c)$ ensures that there is only a passband for positive frequencies. The aspects of synthesis and implementation of active asymmetric polyphase filters are described in [9].

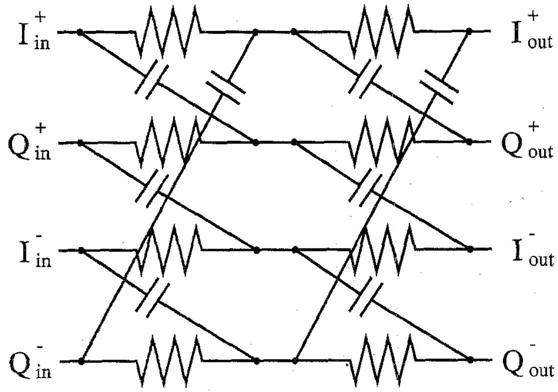

Fig. 6. A passive sequence asymmetric polyphase filter.

The third class of filters for complex signals are the passive sequence asymmetric polyphase filters [10]. Fig. 6 shows one stage of such a filter. It is a symmetric and repetitive version of the classical all-pass RC-CR filter structure that is classically used for the generation of the quadrature LO signals. Where the amplitude and phase errors of classical RC-CR filters are highly sensitive to absolute variations of the R and C values, this is not the case for the symmetric structure. Due to its repetitive four-input four-output structure, it is possible to put several of these filter stages in cascade, allowing the synthesis of more complex and especially more broad-band transfer functions. The latter makes the filter less sensitive to absolute R and C variations.

III. The Low-IF Receiver

A. The High Frequency Part

The concept of the low-IF receiver starts from the observation that, as illustrated with (4) and (5), a high frequency signal can be downconverted to a lower frequency (IF) without the problem of the mirror signal suppression when the multiplication is done with a single positive frequency (which can only be represented by a signal pair). Fig. 7 is an illustration of these operations in the frequency domain for both the positive and negative frequencies. Fig. 7(a) shows the downconversion by multiplication with a sine (i.e., a positive and a negative frequency). This operation superimposes the wanted and the mirror signal by bringing them down to the same frequencies. Fig. 7(b) and (c) gives the downconversion by multiplication with a single positive frequency for, respectively, the zero-IF and the low-IF receiver. In the low-IF receiver are the mirror and the wanted signal not superimposed but downconverted to opposite IF frequencies.

The downconversion to an IF by multiplication with a single positive frequency has large consequences. They are illustrated with Fig. 8. It is now not necessary anymore to do any mirror signal suppression at high frequencies before the downconversion and the IF may be situated at low frequencies (about one to two times the bandwidth of the wanted signal). Both aspects result in an integratability that is as good as the integratablity of the zero-IF receiver. The zero-IF receiver is

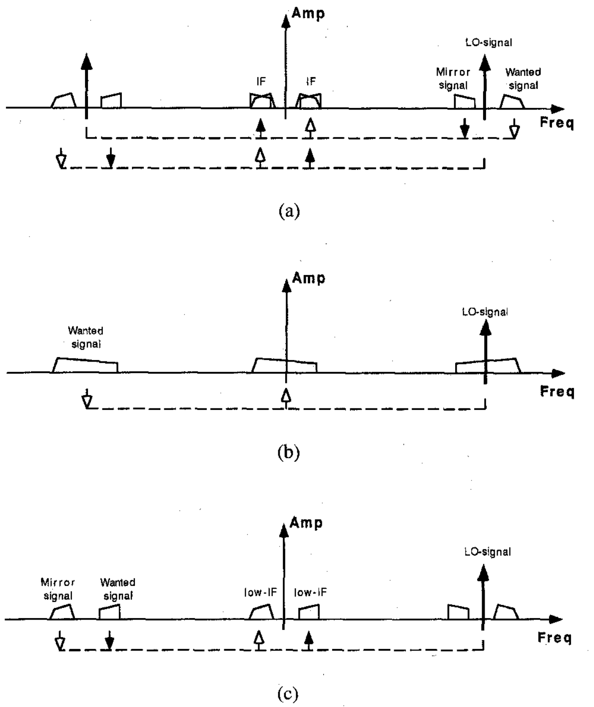

Fig. 7. The downconversion process in (a) an IF receiver. (b) a zero-IF receiver. (c) a low-IF receiver.

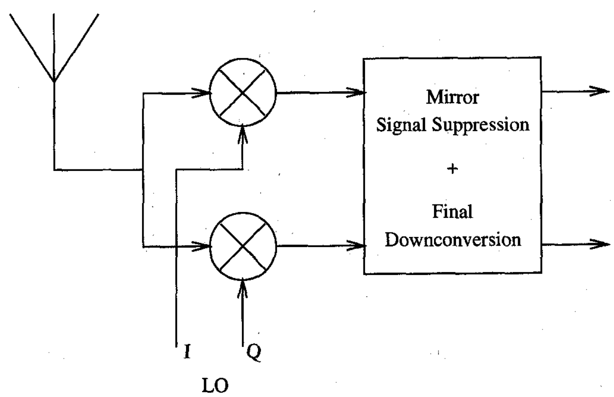

Fig. 8. A downconverter for a low-IF receiver topology.

actually only a special case (IF = 0) of a receiver with a downconversion by means of multiplications with a positive frequency. The use of a zero-IF makes the topology however highly sensitive the parasitic baseband signals like dc offset voltages, second-order distortion products and self-mixing products.

The finite matching between the mixers and, more important, the sensitivity of the RC-CR quadrature generator for the LO, make it impossible to generate a perfect single positive frequency. This is not so important for a zero-IF receiver. Here is the mirror signal equal to the wanted signal and a typical 3° phase accuracy (25 dB mirror signal suppression) will suffice for most applications. This is not the case when

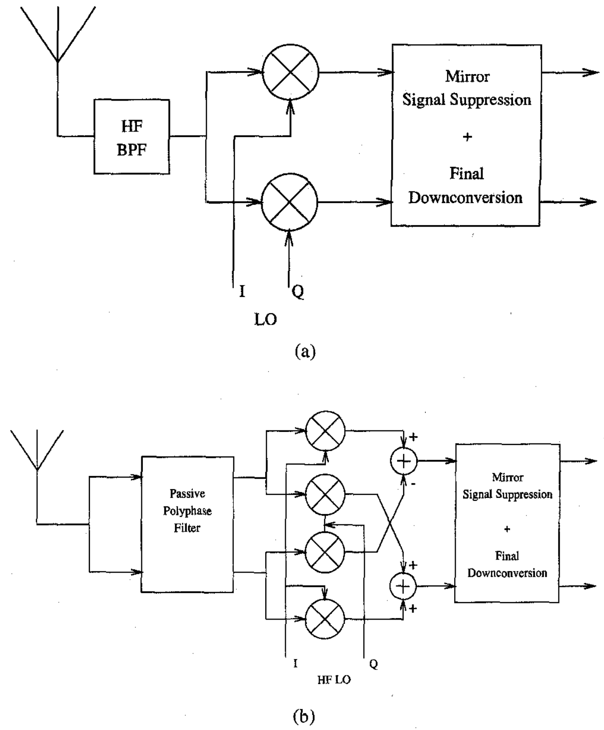

Fig. 9. Low-IF downconversion with (a) an extra external HF filter. (b) a sequence asymmetric polyphase filter.

the downconversion with an imperfect positive frequency is done to an IF frequency. The mirror signal is now completely different from the wanted signal and it can, depending on the application and the exact position of the IF, be up to 20 dB higher than the wanted signal, resulting in a required phase accuracy of 0.3°. This is an accuracy which, with today's typical quadrature downconverters, can only be achieved with the use of extensive tuning and trimming.

It is thus still necessary to do some mirror signal suppression at high frequencies before the downconversion. Fig. 9 gives two possibilities to do this. A quadrature downconverter combined with the classical HF filter, shown in Fig. 9(a), gives only a limited improvement. A HF filter, which can not be integrated, is still necessary, albeit with reduced specifications, and these reduced specifications are only available when still a high IF (between 10 and several hundred MHz) is used. This requires for a further downconversion with a second mixer stage before the wanted signal can be sampled.

A better alternative, which truly combines the advantages of both the IF and the zero-IF receiver, is given in Fig. 9(b). A more close observation of the downconversion process by multiplication with a single frequency component shows that it is only the mirror signal situated at negative frequencies that is superimposed on the wanted signal situated at the IF, while the wanted signal is downconverted from positive frequencies. This means that it is not necessary to suppress the mirror signal at both positive and negative frequencies, as is done with the classical high-Q HF filter. The suppression of only

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