# Analysis of an 1.8 - 2.5 GHz Multi-Standard High Image-Reject Front-End

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## ABSTRACT

In this paper, the analysis of a multi-standard, high image-reject front-end is presented. The front-end is designed to be used for Digital European Cordless Telephone (DECT) systems and wireless communications systems, which operate in the 2.4 GHz Industrial Scientific Medical (ISM) band (like Bluetooth). A double-quadrature low-IF architecture is chosen, because it can provide a high image rejection ratio (*IRR*) and a flexibility in terms of different systems. In the selected architecture, an RC polyphase filter is applied as an RF I/Q generator. Building block specifications are optimized in order to achieve two goals: a high *IRR* and a high sensitivity of the receiver. Assuming a mismatch of 1% between the elements of the RC polyphase filter and 2° phase mismatch in the local oscillator path, a simulated *IRR* of 63 dB is obtained in the frequency range $1.8 - 2.5$ GHz. In spite of the fact that the RC polyphase filter deteriorates the front-end noise figure ($NF$), a $NF$ of 6 dB and a sensitivity of $-93$ dBm for the DECT receiver has been achieved.

## 1. INTRODUCTION

The wireless market is changing very rapidly. Pushed by customer requirements, new systems for wireless communications are emerging very fast. For setmakers it is very useful to have a cost effective front-end solution which can be applied for various communication standards. Considering the frequencies used by the DECT standard (1.9 GHz) and by the ISM band, it is possible to design a front-end which can be used for the DECT systems and for systems in the 2.4 GHz ISM band (like Bluetooth). A key question is then: what are the requirements for such a multi-standard front-end?

The ISM is a free band and numerous systems are located in this band (for example, Wireless Local Area Network [1], Bluetooth). As a result, a lot of interference will be present in the ISM band. In order to prevent the corruption of the wanted signal by the image signal, the front-end must provide a high image rejection.

The first step in the design of a multi-standard front-end is the selection of an architecture. A double-quadrature low-IF architecture is selected (Fig. 1) [2] in which a RC polyphase filter is chosen to implement RF I/Q generation. The arguments for this choice are given in section 2. The limitation in achievable *IRR* are analyzed in section 3 and the simulations results are presented in section 4. The choice of the RC polyphase filter for RF I/Q generator has the advantage that the implementation is straightforward, but it also degrades the front-end $NF$ which will deteriorate the receiver sensitivity. Special attention is given to this problem. The simulations results of the overall front-end $NF$ are presented in section 5.

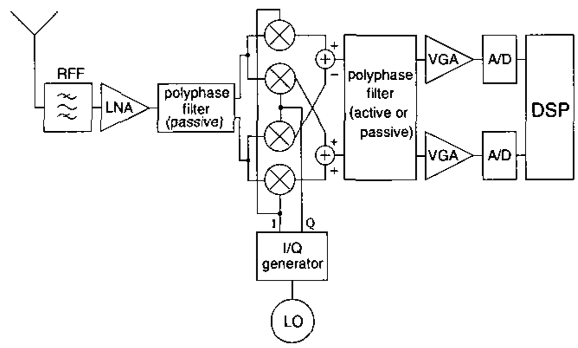


Figure 1. The double-quadrature low-IF architecture

## 2. FRONT-END ARCHITECTURE

The architecture selection has been done with the intention to select the architecture which can provide flexibility for different standards and a high *IRR*. Driven by market requirements to reduce cost, CMOS technologies migrate towards deep submicron processes. So sensitivity of an architecture to the technology scaling also influences the selection. The most frequently used architectures: superheterodyne, zero-IF and low-IF will be considered.

The major source of problems in the superheterodyne architecture is the high-frequency image-reject filter (HFIRF). The high requirements for this filter (Q factor is between 9 and 30 and the order is higher than 6) will make integration very difficult [2] , while discrete solutions are not cost-effective. The flexibility for different standards is low. Firstly, it is very difficult to make the HFIRF tunable. Secondly, integrated HFIRF can't achieve high *IRR* with reasonable *IF* and the *IRR* can't be improved after the down-conversion. Therefore, such front-end can't be applied in systems requiring an even higher *IRR*.

The main problem in zero-IF architecture is DC offsets. They are a result of self mixing and oscillator pulling, which depend on capacitive and substrate coupling. With CMOS scaling, the dimensions of the circuit will be further reduced, which will cause stronger coupling and increase DC offsets. The method of the DC offset cancellation depends, very much, on the system where front-end is applied. AC coupling can not be applied in the systems with the modulation schemes having a high energy content at the center frequency of their spectrum. DC offset cancellation in the digital domain can be applied in the systems which use Time Division Multiple Access (TDMA) technique where the blind slots are implemented (e.g. DECT). Summarizing, the flexibility for different standards is low.

The main advantage of the low-IF architecture is that the image rejection is done after the down-conversion, by a polyphase filter. So any changes in the system operating frequency can be solved by using another local oscillator (LO) frequency. In the case that an higher *IRR* is required than the *IRR* provided by the polyphase filter, there is the possibility to suppress the image signal, further, in the digital domain. Problems like DC offsets and integration of HFIRF are avoided. All this makes low-IF architectures very flexible for different standards. In low-IF architectures achievable *IRR* depends mainly on the I and Q matching in the LO path. The *IRR* can be increased by making a more accurate I/Q generator in the LO path or by adding a polyphase filter in the RF path which will work as RF I/Q generator [2]. The second solution is better because higher *IRR* can be achieved. In this way, double-quadrature low-IF architecture is obtained (see Fig.1). Overall *IRR* can be calculated as:

$$IRR\,[\mathrm{dB}] = \mathrm{IRR_{RF}}\,[\mathrm{dB}] + \mathrm{IRR_{LO}}\,[\mathrm{dB}] \tag{1}$$

$IRR_{RF}$ is determined by the matching in RF polyphase filter and $IRR_{LO}$ by the matching in the I/Q generator in LO path.

## 3. COMPLEX TRANSFER FUNCTION AND IMAGE REJECTION OF THE POLYPHASE FILTER

In the Fig. 2, the topology of the RC polyphase filter is presented.

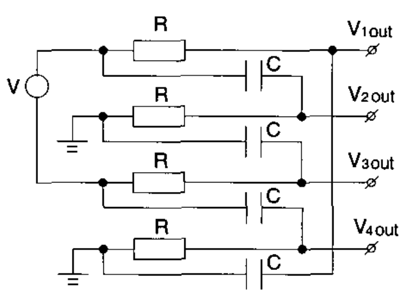


Figure 2. The topology of the RC polyphase filter

$$\begin{aligned} V_I &= V_{1out} - V_{3out} && (2)\\ V_Q &= V_{2out} - V_{4out} && (3) \end{aligned}$$

In [3] it was shown that the ratio of the amplitudes $\left|V_I/V_Q\right|$ depends on the frequency and the phase difference $\varphi_Q - \varphi_I$ is always 90°.

The complex transfer function of the RC polyphase filter is defined in the following way:

$$H_c = \frac{V_I + jV_Q}{V} = \frac{V_{outc}}{V} \tag{4}$$

In the case of one, two and three stage RC polyphase filter, the complex transfer functions are given in [3].

Using the complex transfer function it is very easy to define the *IRR* and to see the limitations for the *IRR*. The process of down-conversion in the frequency domain is represented in the Fig.3. $A_w$ is the amplitude spectrum of the wanted signal and $A_i$ of the image signal. There are two

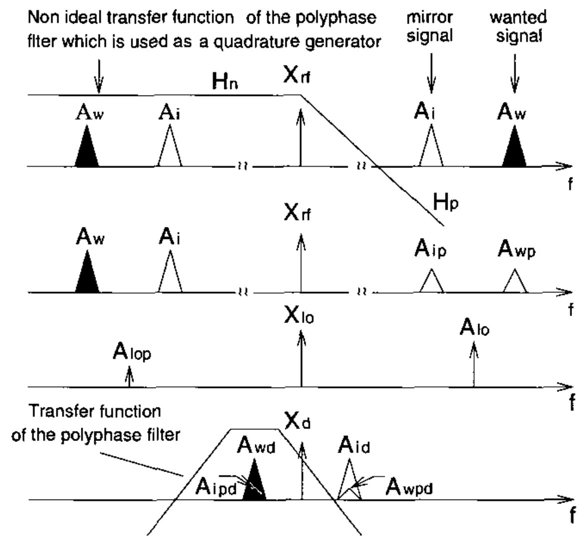


Figure 3. The down-conversion in the frequency domain

sources of limitations. The first is the topology of the RC polyphase filter. From the formulas for complex transfer function [3], it can be seen that infinite *IRR* can be achieved only in a few frequency points. The second source is the spread in the absolute values and mismatches between the R and C elements. This means that I/Q generation, done by RC polyphase filter, is not ideal, which explains the presence of the signals $A_{ip}$ and $A_{wp}$. Signal $A_{lop}$ is result of mismatches in the LO path. Due to the convolution between the signals $A_{ip}$ and $A_{lop}$, signal $A_{ipd}$ will fall in the band of the wanted signal and will limit the *IRR*. The *IRR* can be calculated using (5).

$$IRR = \frac{A_{wd}}{A_{ipd}} \frac{A_i}{A_w} \tag{5}$$

Further, the following equations can be written:

$$\begin{aligned} A_{wd} &= |H_n| A_{lo} A_w && (6)\\ A_{ipd} &= |H_p| A_{lop} A_i && (7)\\ H_I &= H_{Ir} + jH_{Ii} && (8)\\ H_Q &= H_{Qr} + jH_{Qi} && (9)\\ H_p &= H_I + jH_Q && (10)\\ H_n &= H_I^* + jH_Q^* && (11)\\ IRR_{LO}[dB] &= 20log(A_{lo}/A_{lop}) && (12) \end{aligned}$$

Finally, the $IRR_{RF}[dB]$ can be expressed, using (1) and (6) ... (12), as:

$$IRR_{RF} = 10log\frac{(H_{Ir}+H_{Qi})^2+(H_{Qr}-H_{Ii})^2}{(H_{Ir}-H_{Qi})^2+(H_{Ii}+H_{Qr})^2} \qquad (13)$$

## 4. $IRR_{RF}$ SIMULATIONS RESULTS

In order to see the influence of the spread in absolute values and mismatches between resistors (R) and capacitors (C) in the RC polyphase filter to the $IRR_{RF}$, statistical simulations were performed for the RC polyphase filters with one, two and three stages. The simulator is set for 50 trials and takes randomly values of the R and C elements from the interval $[-4\sigma, 4\sigma]$ around the nominal value. For the spread in ab-

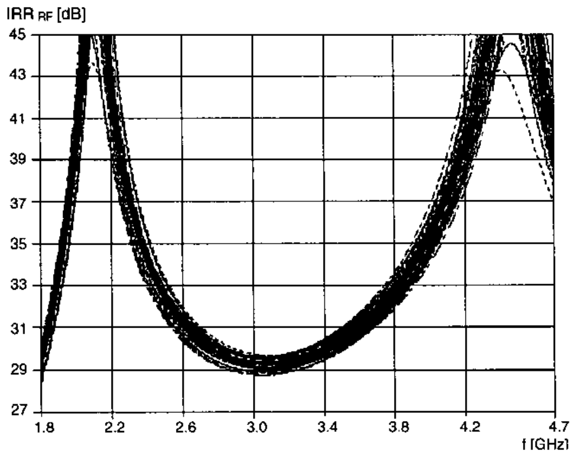


Figure 4. $IRR_{RF}$ for RC polyphase filter with two stages in the case that $\sigma_m = 1\%$

solute value of the R and C elements, the following values were used: $\sigma_R = 4\%$ and $\sigma_C = 5\%$. For the mismatch the simulations are performed with the following values of the $\sigma_m$: 1%, 2% and 5%.

Simulations for the spread in the absolute value and for mismatch are done separately. In the case of simulations for the spread in absolute value, the simulator is set to take the values for all resistors (R) in the RC polyphase filter completely correlated. The same is done for all capacitors. In the case of simulations for mismatch, those values are taken completely uncorrelated. In order to find the worst case $IRR_{RF}$, the spread in the absolute value and mismatch must be taken together. This can be done by simulating the mismatch with the worst case for the spread in absolute value.

The values for R and C are optimized for maximum $IRR_{RF}$ in the band $1.8-2.5$ GHz. The values for resistors in the two stage RC polyphase filter were set identical in all stages ($R$), while the values for capacitors in the first ($C_1$) and in the second stage ($C_2$) are different. Neglecting the spread in the absolute values and assuming that the value for $R$ is known, the values for $C_1$ and $C_2$ can be calculated [3] using (14), where $f_1 = 1.8$ GHz and $f_2 = 2.5$ GHz.

$$f_1 = 1/(2\pi RC_1),\ f_2 = 1/(2\pi RC_2) \qquad (14)$$

But, due to the spread in the absolute values, the $IRR_{RF}$ curve will be shifted in the frequency, which will reduce $IRR_{RF}$ in the band $1.8-2.5$ GHz. Simulations showed that maximum $IRR_{RF}$ is achieved in the band $1.8-2.5$ GHz, if $f_1 = 1.4$ GHz and $f_2 = 3$ GHz. In the Fig. 4, the simulation results for $IRR_{RF}$ for RC polyphase filter with two stages are represented in the case that $\sigma_m = 1\%$. The simulation results for $IRR_{RF}$, in the band $1.8-2.5$ GHz, are summarized in the table 1 for one, two and three stages.

Table 1. Simulation results for $IRR_{RF}$ [dB]

| Stages | $\sigma_m = 1\%$ | $\sigma_m = 2\%$ | $\sigma_m = 5\%$ |
|---|---|---|---|
| 1 | 11 | 10 | 9 |
| 2 | 28 | 27 | 25 |
| 3 | 38 | 35 | 30 |

Assuming the phase mismatch of 2° in the LO path, the $IRR_{LO}$ of 35 dB can be achieved [2]. Hence, from the results in the table 1 and using (1), it can be seen that with two stage RC polyphase filter, the $IRR$ of 63 dB can be achieved in the frequency range $1.8-2.5$ GHz.

## 5. INFLUENCE OF THE POLYPHASE FILTER TO THE NOISE FIGURE OF THE FRONT-END

The noise figure ($NF$) of the front-end is approximated by considering only one path in the front-end chain. The noise factor $F$ is calculated using (15), which represent the extension of the Friis formula. Stage $i$ in the Fig. 1 is described with noise factor ($F_i$), which is calculated with respect to the output impedance of the previous stage, gain ($A_i$), input impedance ($R_{ini}$) and output impedance ($R_{ini}$). The RF out-of-band rejection filter (RFF) is described by the loss ($L$). The contribution of all the stages after the mixers in the front-end chain to the $F$ of the front-end is neglected. This is a reasonable assumption, because the gain introduced by the Low Noise Amplifier (LNA) and mixer reduces the contribution of the noise from the stages after the mixer.

$$\begin{aligned} F &= L\cdot F_2 + L\frac{F_3-1}{(\frac{R_{in2}}{R_{in2}+R_{out1}})^2 A_2^2 \frac{R_{out1}}{R_{out2}}} + \\ &+ L\frac{F_4-1}{(\frac{R_{in2}}{R_{in2}+R_{out1}})^2 A_2^2 A_3^2 \frac{R_{out1}}{R_{out3}}} \end{aligned} \qquad (15)$$

From (15) it is clear that the RC polyphase filter will increase $F$ of the front-end because $F_3 > 0$ and $A_3 < 1$. Therefore it is very important to see which $NF$ can be achieved and what are the design parameter values for the building blocks (LNA and mixer) in that case. These parameter values can be used as starting values in the building block design. As the calculations get complicated in the case of a double-quadrature low-IF architecture because of the non-unilateral character of the RC polyphase filter, simulations must be done.

The following information which relate to the $NF$ simulations is important:

- Using (15) the noise figure of the front-end can be expressed as: $NF$ [dB] = L [dB] + $\mathrm{NF_a}$ [dB]. $NF_a$ is the noise figure excluding the RF out-of-band rejection filter (RFF) (see Fig.1). So $NF_a$ can be simulated and in order to get overall $NF$, $L$ should be added.

- The noise figure of the LNA ($NF_{LNA}$) and mixer $NF_{MIX}$ are calculated assuming 50 $\Omega$ source resistance.

- The value of the resistors in the two stage RC polyphase filter is: $10 \leq R \leq 50\ \Omega$. The values for $C_1$ and $C_2$ can be calculated with (14), taking $f_1 =$ 1.4 GHz and $f_2 = 3$ GHz.

- The following data is used for the simulations: $L =$ 1.5 dB, LNA voltage gain ($A_{LNA} = 15\ dB$), LNA noise figure ($NF_{LNA} = 2\ dB$), LNA input impedance ($R_{ILNA} = 50\ \Omega$), LNA output impedance is taken as a parameter for $NF$ simulations: ($10 \leq R_{OLNA} \leq 50\ \Omega$), mixer voltage gain ($A_{MIX} = 20\ dB$), mixer noise figure ($NF_{MIX} = 10\ dB$), mixer input impedance ($R_{IMIX} =$ 500 $\Omega$).

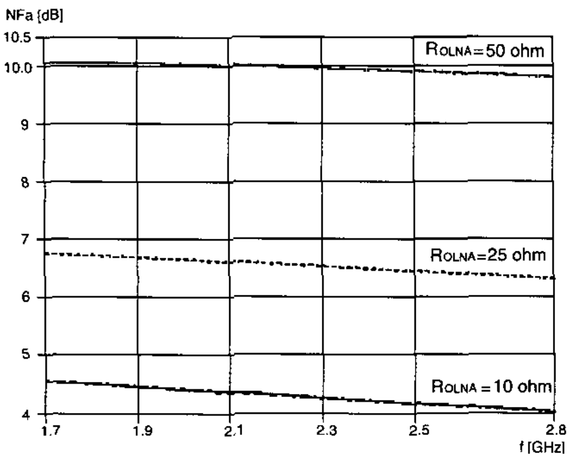


Figure 5. $NF_A$ versus frequency for the RC polyphase filter with two stages, with $R = 10\ \Omega$ and for three $R_{OLNA}$ values

The simulation results of $NF_a$ are given in the Fig.5 and Fig.6. From the simulations the following conclusions can be drawn:

- The front-end noise figure is strongly dependent on the output impedance of the LNA (Fig. 5).

- The front-end noise figure is strongly dependent on the resistance ($R$) which is used in the polyphase filter (Fig. 6).

- The front-end noise figure increases with the number of stages in the RC polyphase filter. For $R_{OLNA} = 10\ \Omega$ and $R = 10\ \Omega$ in case of a two stage RC polyphase filter ($C_1 = 11.2$ pF, $C_2 = 5.3$ pF) a $NF = 6\ dB$ is obtained. For $R_{OLNA} = 10\ \Omega$ and $R = 10\ \Omega$ in case of a three stage polyphase filter, $NF = 7.8\ dB$.

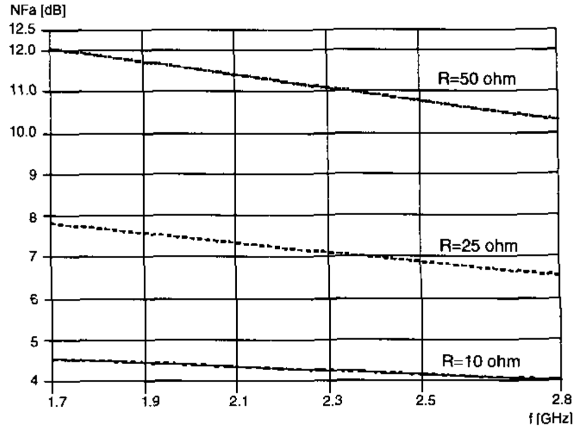


Figure 6. $NF_a$ versus frequency for the RC polyphase filter with two stages with $R_{OLNA} = 10\ \Omega$ and for three $R$ values

## 6. CONCLUSION

The feasibility of a 1.8 − 2.5 GHz multi-standard, high image-reject front-end is examined. It was explained that the double-quadrature low-IF architecture, in which RC polyphase filter is used as RF I/Q generator, provides flexibility and a high $IRR$, which are necessary for a front-end which can be applied in both DECT systems and wireless communications systems which operate in ISM band (like Bluetooth). The limitations in achievable $IRR$ are discussed and explained. The influence of the mismatch between $R$ and $C$ elements in the RC polyphase to the $IRR$ is evaluated by the statistical simulations. An $IRR$ of 63 dB is obtained in the case of 1% mismatch between the elements in the two stage RC polyphase filter and 2° phase mismatch in the local oscillator path. Simulations with behavioral models for the LNA and mixer showed that with the realistic building block specifications, it is possible to achieve a low $NF$ (6 dB) and thus provide a high sensitivity of the DECT receiver, namely −93 dBm.

## References

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