SELF CALIBRATING LOW IF DIGITAL IMAGE REJECTION RECEIVER FOR MOBILE COMMUNICATIONS

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ABSTRACT

Here we present and develop a receiver capable of capturing two RF channels at the same time with a single RF front end and only one IF stage. The idea is to use a low IF digital image rejection receiver that can separate two adjacent RF channels with a negligible cochannel's image interference. We analyze two procedures of compensating, in the IF range, any gain and phase misadjustment generated in the RF mixing section that could produce some residual images in any of the channels. The first one needs the help of an internal reference or pilot signal whereas the second one implements a blind procedure that only needs the current working signals.

I. INTRODUCTION

The idea of using a low IF image rejection receiver for mobile communications can be an advantageous alternative to the classical 2-stage IF analog receiver and the new zero-IF digital receiver concept [1]. The proposed scheme needs only one downconversion to a low IF plus an image rejection stage to avoid any possible in-band image channel to be superimposed to the desired one. The IF image rejection stage needs the I+Q components to selfcompensate, at least in one of the branches, the image band with the help of a Hilbert Transformer (HT). This well known concept can be reused today as a clear alternative for implementing IF digital receiver to that of the zero-IF digital one, since it does not suffer of the dc-compensation problems as the zero-IF version does.

Moreover, image rejection receivers allow the use of RF front-end filters with relatively poor selectivity. This is at the expense of introducing an image cancellation device which is realized by means of a hybrid circuit in many analog systems. For the carrier frequencies and bandwidths used in modern mobile communications, the image-rejection ratio attained by those analog devices (25-35 dB) is sometimes insufficient. The reason for this can be traced down to the amplitude and phase inbalance of the quadrature demodulator. In principle the hybrid must be adjusted for compensation of an unknown (although close to 90°) phase angle and amplitude inbalance (close to 0) and this is done with some difficulty in practice. Figure 1 shows a realistic scheme of our proposal, where α and ε are the gain and phase misadjustments respectively. Observe how the two RF bands at the input can be totally separated in the IF range by using a compensation linear network that cancels their interferences in the upper and lower branches. Now, two complete receivers can demodulate both signals separately and extract the corresponding information (digital or analog).

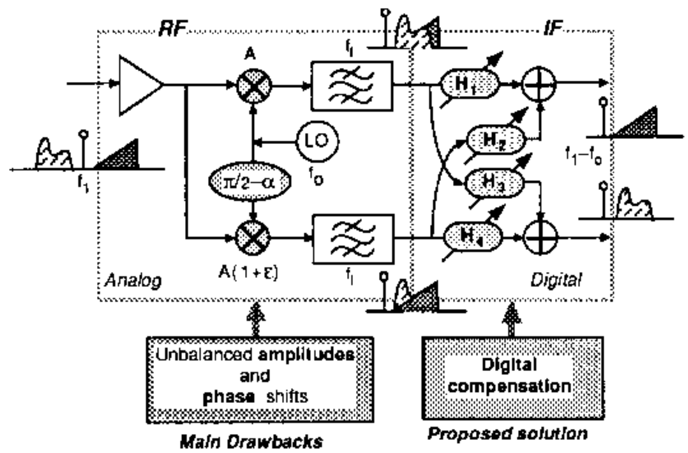

Figure 1: Low IF image rejection system

II. REJECTION PERFORMANCE WITH ε, α

In an ideal situation of identical IF filters and balanced branches ($\varepsilon=\alpha=0$), the compensation filters H_1 to H_4 have to implement the functions $H_1=H_4=1$ and $H_2=H_3=HT$, however, in the case of $\varepsilon\neq0$, $\alpha\neq0$ and identical IF filters they change theoretically to $H_1=H_3^*=\cos(\alpha)-\sin(\alpha)HT$ and $H_2=H_4^*=HT$, and finally in the realistic case of unbalanced branches (non ideal phase splitter, mixers and IF filters) H_1 to H_4 should be short memory and, in general, unknown filters. In fact, the signal to image ratio (assuming equal input powers) in any branch for a mixing operation with unbalanced branches is given by [3]

$$C(\varepsilon,\alpha)=\frac{1+(1+\varepsilon)^2+2(1+\varepsilon)\cos\alpha}{1+(1+\varepsilon)^2-2(1+\varepsilon)\cos\alpha} \qquad (1)$$

Observe in Figure 2 the maximun attainable image rejection levels when we try to compensate a generic unbalanced mixing stage with a network of ideal HTs. There is a clear even symmetry in the amplitude and phase inbalance directions. Observe also how slight impairments (below 0.5 dB and 4 degrees) set an upper rejection bound to 27 dB.

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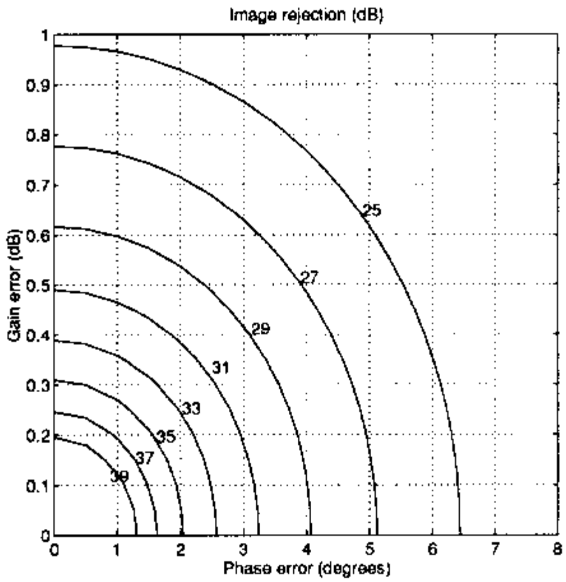

Figure 2: Image rejection performance

This maximum rejection might be in some practical systems (for instance GSM or DECT) an important drawback for the utilization of this kind of receivers. This justifies the implementation of the full compensation system afterwards. As reported in [2, 3] this can be efficiently done by digital means with finite precision arithmetic, thus avoiding the cumbersome adjusting of mechanical devices, inaccuracy, etc, typical of analog devices. Of course the values of amplitude and phase inbalances are assumed to be known. In practice this means that the compensation system must be calibrated for the inbalance values found in a specific device.

III. DESIGN OF THE COMPENSATION UNIT

Assuming a perfect balanced RF mixing operation, the image can be completely removed by compensating the 90º degrees phases between both branches using the scheme proposed in Figure 3. This is a theoretic scheme where all the 90º phase shiftings are assumed to be perfect. Actually, the real 90º shifters of a real the system are implemented in two very different manner. whereas the ones in the mixing unit are narrow band (RF band), the ones in the Image separation unit are usually wideband (low IF band). This fact impacts directly in their implementation

However, whenever phase error exists, the image at IF can be completely removed by inserting the corresponding phase compensation in the image separation unit. The corresponding equivalent scheme is shown in Figure 4 only for the upper branch.

Observe how the exact knowledge of α is enough to remove totally the image signal in the upper branch. In this scheme the $-\alpha$ shifter has the following transfer function

$$H_{-\alpha}(f)=\begin{cases} e^{j\alpha}, & f>0 \\ e^{-j\alpha}, & f<0 \end{cases} = \cos\alpha - \sin\alpha H_{\pi/2}(f) \qquad (2)$$

here we observe that any phase shift can be directly implemented by using a Hilbert transformer.

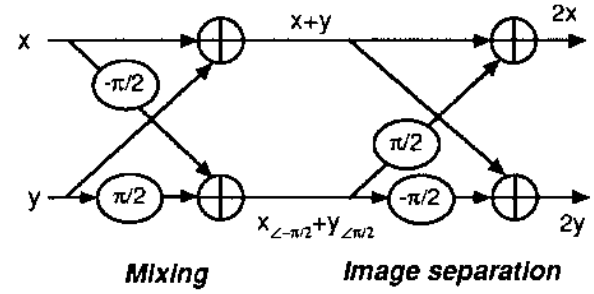

Figure 3: Balanced mixing and compensation

Denoting now x_u and x_l as the outputs at the mixing section in the upper and lower branch respectively and introducing the gain compensation in the lower branch we obtain (see Figure 4)

$$\begin{aligned} x_u &= x + y \\ x_l &= (1+\varepsilon)\left(x_{\angle\alpha-\pi/2} + y_{\angle\pi/2-\alpha}\right) \end{aligned} \qquad (3)$$

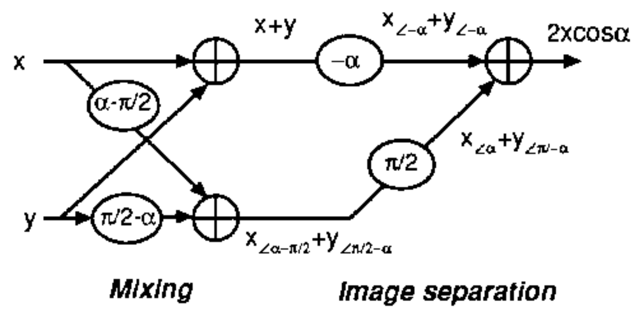

Figure 4: Unbalanced mixing and compensation

Therefore, considering this last scheme and equations (2) and (3) we can provide the final implementation for the compensation network. It is shown in Figure 5

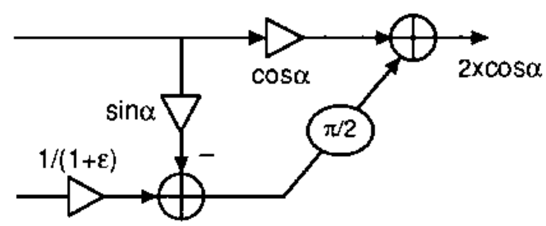

Figure 5: Compensation network (upper branch)

In summary, to remove the image signal, in any of the branches we only need to add 3 multipliers to the balanced compensation network.

IV. CALIBRATING THE COMPENSATION NETWORK

We propose in this section two different methods of calibrating the digital IF compensation network in order to remove as much as possible the image band from the wanted one in the desired branch (upper or lower). The first approach is oriented to calibrate the network in the manufacturing phase of the transceiver (we assume stable impairments with time or alternatively very slow drifts). Here we assume to have access to an internal RF test signal at the RF stage and the corresponding reference in the IF section that allows to learn, for instance adaptively, the compensation coefficients. This method is schemed in Figure 6

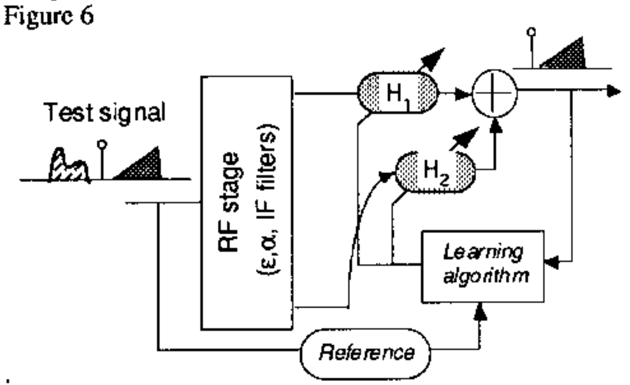

Figure 6: Aided or trained calibrating system

Here the learning algorithm can use the proper internal reference therefore allowing the setting of a good residual image that can be cancelled adaptively. Actually, the easiest procedure is to use a test signal with just only one band (x or y) in order to eliminate the need of the internal IF reference.

In the following we will denote δ and β as the two unknowns to compensate the misadjustments ε and α respectively and w_u and w_l the two outputs at the compensation unit, that is

$$\begin{aligned} w_u &= x_u \cos\beta - x_{\hat{u}} \sin\beta + x_{\hat{l}} / (1+\delta) \\ w_l &= x_u \cos\beta + x_{\hat{u}} \sin\beta - x_{\hat{l}} / (1+\delta) \end{aligned} \qquad (4)$$

where, for symplicity in the notation, the circumflex (^) in the subindeces denotes a Hilbert transformation.

In order to calibrate the system of Figure 6, let us assume the following situation (x=0 and $y \neq 0$). Here, according to Figure 5, we expect an image signal in the upper branch, theferore denoting this output as the system error $e = w_u$, we can easily set up the corresponding adaptive gradient algorithm. Denoting γ as the $1/(1+\delta)$ factor and making linear the system error around β=0 and γ=1 we obtain

$$\begin{aligned} e &\approx x_u + \gamma x_{\hat{l}} - x_{\hat{u}}\beta = x_u + \mathbf{w}^t\mathbf{z} \quad \text{where} \\ \mathbf{w} &= [\gamma \;\; \beta]^t \;\; \text{and} \;\; \mathbf{z} = [x_{\hat{l}} \;\; -x_{\hat{u}}]^t \end{aligned} \qquad (5)$$

with these definitions the corresponding stochastic gradient algorithm is

$$\begin{aligned} \mathbf{w}(0) &= [1 \;\; 0]^t \\ \mathbf{w}(n+1) &= \mathbf{w}(n) - \mu e(n)\mathbf{z}(n) \end{aligned} \qquad (6)$$

where μ is the corresponding adaption step. In fact if we use in 6 the following modified version of the error

$$e_m = x_u \cos\beta + \mathbf{w}^t\mathbf{z} \qquad (7)$$

instead of the complete linearized version given in equation (5) we obtain a better unbiased result. For instance, assuming a mixing section with misadjustment α=5°, ε=0.8 dB (an original image rejection below 27 dB), using a pure tone for signal y to drive the adaptive system with an adaption step μ=.05, we obtain the results shown in Figure 7, that is, after 100 iterations the adaptive algorithm has converged to the right solution . For these new gain and phase, the image rejection is now above 65 dB.

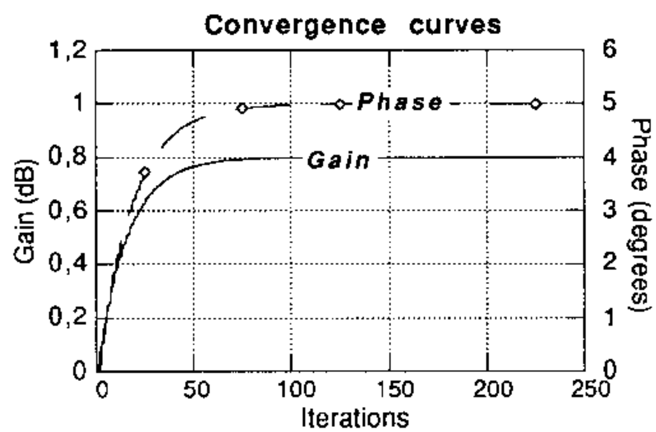

Figure 7: Convergence results for β and δ

Although these good results validate the procedure, it requires an internal generator to drive the adaptive system properly. Since this pilot signal has to be really implemented in the RF section of the receiver, it could probably be a penalty in the implementation of cheap receivers. To avoid this possible drawback, we present a second blind approach which is oriented to obtain a selfcalibration in both branches simultaneously in order to provide two uncorrelated outputs. Figure 8 ilustrates this concept. Here we implement a blind scheme where we try to separate both channels by imposing a zero cross-correlation between both branches as objective function. This situation is interesting from the point of view of having just a single receiver (a radio base station) receiving two independent bands in any FDM system (GSM, DECT, etc). The main consideration here is the important saving in electronic RF components. Concerning the possible local minima of an objective function based on the nulling of the cross-correlation between the output of both branches we can affirm that a good guess of the initial settings would lead the network coefficients to the global minimum.

Since the RF stage presents, in general, only slight deviations from the ideal balanced situation we can initalize again the algorithm to the nominal values, thus expecting a good convergence to the global minima.

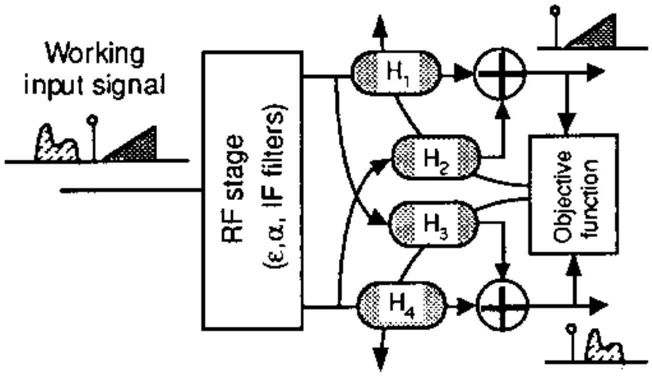

Figure 8: Blind selfcalibrating system

Using equations (4) we can compute the cross correlation between both branches as

$$R_w(m)=E\{w_u(n)w_l(n+m)\}=R_u(m)\cos 2\beta-\frac{R_l(m)}{(1+\delta)^2}+ \\ +R_{u\hat{u}}\sin 2\beta+\frac{2R_{\hat{l}u}(m)\cos\beta+\left(R_{\hat{u}\hat{l}}(m)+R_{\hat{l}\hat{u}}(m)\right)\sin\beta}{1+\delta} \quad (8)$$

where R_u, R_l, R_{ul}, etc, are the correlation functions at the input of the compensation network.

To impose a zero cross correlation, $\forall m$, between both branches we need to set up an infinite system of nonlinear equations in β and δ. However assuming that the input signals are usually wideband, in the IF range, and that both misadjustments allows a linearization of the system, we will use an equation set of just two linear equations. Furthermore, considering the properties of the impulse response of a Hilbert transformer and stationary correlation functions, i.e. $h(m)=-h(-m)$ and $R_{xy}(m)=R_{yx}(-m)$, we can rewrite after some analysis and approximations the expresion (8) for m=0 and m=1 as

$$R_w(0)\approx R_u(0)\cos 2\beta+\frac{2R_{ul}(0)\sin\beta}{1+\delta}-\frac{R_l(0)}{(1+\delta)^2} \\ R_w(1)\approx R_{u\hat{u}}(1)\sin 2\beta+\frac{2R_{\hat{l}u}(1)\cos\beta}{1+\delta} \quad (9)$$

If we now force a zero cross correlation for $R_w(m)$ and we make the linearization of (9) around [β δ]=[0 0], we finally obtain the system of equations

$$\begin{bmatrix}2R_{ul}(0) & R_u(0)+R_l(0)\\ R_{u\hat{u}}(1) & -R_{\hat{l}u}(1)\end{bmatrix}\begin{bmatrix}\beta\\ \delta\end{bmatrix}=\begin{bmatrix}R_l(0)-R_u(0)\\ -R_{\hat{l}u}(1)\end{bmatrix} \quad (10)$$

this system can be solved recursively using the proper correlation estimations, however we are not interested now in the recursive algorithm but in the final result for the gain and phase misadjustments. In this case we have used two GMSK (BT=0.5) uncorrelated input signals of equal power with a sampling rate of three samples per symbol. Furthermore we have assumed 4 different misadjustments situations for the mixing section : α=3 and 5 degrees and $(1+\varepsilon)_{dB}$=0.6 and 0.8 dB. To compute the correlation estimations we have used 10000 samples of x and y inputs. The results are shown in Figure 9. It shows the residual misadjustments with and without the insertion of the blind compensation network. Observe how in this case there still exist a significant residue. It is equivalent to an image rejection of about 46 dB (an improvement of around 20 dB for the 4 cases).

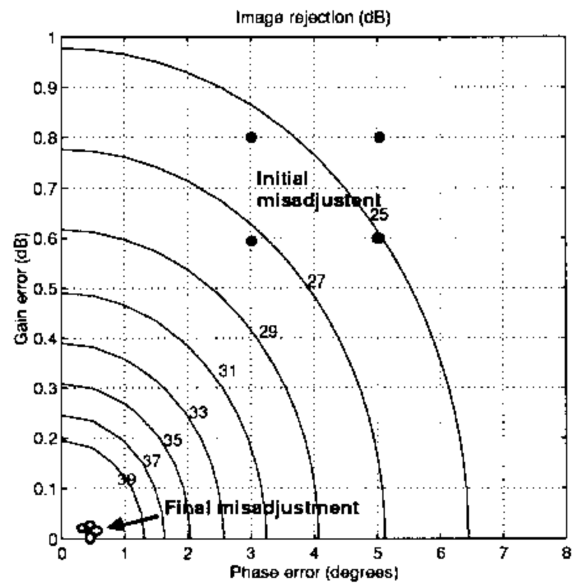

Figure 9: Performance of the blind scheme

V. CONCLUSIONS

We present in this work a receiver capable of receiving two independent RF channels with a single RF front end and just one low IF stage. The presence of image bands in the IF section for any of the two channels can be digitally separated by using a linear network consisting of two wideband Hilbert transformers. However, the gain and phase misadjustments present in the RF mixing unit may cause some residual uncompensated images. To deal with this relative important drawback, we also present two compensation methods (trained and blind ones) that estimate the gain and phase RF inbalances and therefore, it can extend significatively the rejection (or separation) capabilities of the complete low IF image rejection receiver.

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