AN INTEGRATED GSM/DECT RECEIVER: DESIGN SPECIFICATIONS

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Memorandum No. UCB/ERL M97/82

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1.0 Introduction

The following document is a preliminary set of specifications for the second generation multi-standard receiver. These specifications should be treated as a living document subject to change; careful attention should be paid to the revision date.

A brief summary and description will be given for all of the standards under consideration on this project. Then a detailed explanation is given of the method used to calculate specifications for the overall receiver and each of the individual receiver components. These calculation methods were then applied to two receiver architectures under consideration, Wideband IF w/ Double Conversion (WIF) and Low-IF single conversion. Clear easy to follow specifications are given for each receiver component with respect to equivalent input noise, input voltage IP3, maximum required swing, 1dB compression point, input IP2 of the receiver baseband and dynamic range.

It was decided early in this project that the wideband IF with double conversion architecture would be used for the second generation receiver. Therefore, the specifications for the different blocks have been considerably more refined (and practical) for Wideband IF verse the other architecture, Low-IF single conversion.

1.1 Updates

05/09/97 - Specs changed for all of the mixers. The dynamic range of the ADC was lowered from 16 to 14 bits. GPS has been dropped from the possible standards under consideration. PCS 1900 has been dropped from consideration.

10/12/97 - Specs. on the baseband have changed. Requirements for the anti-alias filter and the ADC have been revised. The standards now being implemented are DCS 1800 and DECT. The ADC sampling frequency is set to 44.8MHz.

10/14/97 - Assuming a conversion directly to baseband for the second mixer stage, a 16 bit ADC can now potentially be used. Therefore, less attenuation is needed from the anti-alias filter on the blockers for both GSM and DECT.

12/4/97 - The ADC is now set to 15 bits while the baseband anti-alias filter requirements have been changed. Also, a new IF-to-Baseband mixer has been designed. The second frequency translation is now realized with a passive MOS ring mixer.

1/19/98 - Currently we are examining a 13bit, 13.5bit, and 14bit ADC. Some modifications to the baseband filter are now being considered.

1/26/98 - The specifications for the ADC were frozen at 14bits of resolution for GSM. Also, it was decided to place a variable gain stage immediately after the mixer to accommodate gain variation in the RF receiver blocks and thus reduce the required dynamic range of the ADC. The variable gain stage at baseband is used only to guarantee that the largest signal present is set to the maximum range of the ADC.

2/15/98 - The frequency plan for the synthesizers have been completely revised. It was found that the crystal frequency was far too low, requiring a large divider and hence an unacceptable amount of noise enhance from the divider to the local oscillator output. To accommodate the new required crystal reference frequency for both DCS 1800 and DECT a high IF was required at the output of the first mixer. This then eliminated the possibility of using a passive ring mixer for the second stage frequency translation. Therefore, the second mixer stage has now been converted from a sampling ring like mixer to a Gilbert cell based mixer.

2.0 Summary of E-GSM/DCS1800/PCS1900/ DECT Standards

A few highlights of the GSM, DECT and GPS specification as they apply to the receiver are given. The specifications are outlined with respect to noise figure, intermodulation, blocking performance, image rejection, and maximum inband desired signal.

GSM is a standard that was developed by the european standards committee[1]. The original version of GSM was used in the 900MHz band throughout europe. Then on the request of the english, an upbanded version of GSM was added in the 1800MHz band, this was original called PCN for personal communications network[3]. PCN was later renamed to DCS1800. An extra 10MHz was added to the lower end of both the receive and transmit bands to create Enhanced-GSM or E-GSM. The european GSM is currently being deployed in the United States in the 1900MHz band and is named PCS1900. The digitally enhanced cordless telecommunications (DECT) standard [2](originally known as digital european cordless telephone standard) was developed with the idea of creating mini-cells which would cover a couple square blocks in a residential neighborhood.

The following section gives a brief description of the radio physical layer specifications for all of the standards under consideration.

2.1 Frequency Bands

Shown in figure 1 are the frequency bands of interest for this project. The channel spacings for GSM, Enhanced GSM, DCS 1800, and PCS 1900 are 200kHz wide, while DECT consist of 10 channels with a bandwidth of 1.728 MHz. The single GPS channel for the Coarse Acquisition (C/A) code is 2 MHz wide. Originally, all of these standards were under considerations. However, to demonstrate both multi-carrier and variable channel bandwidth both DCS 1800 and DECT were selected as the target standards to implement for a demonstration prototype.

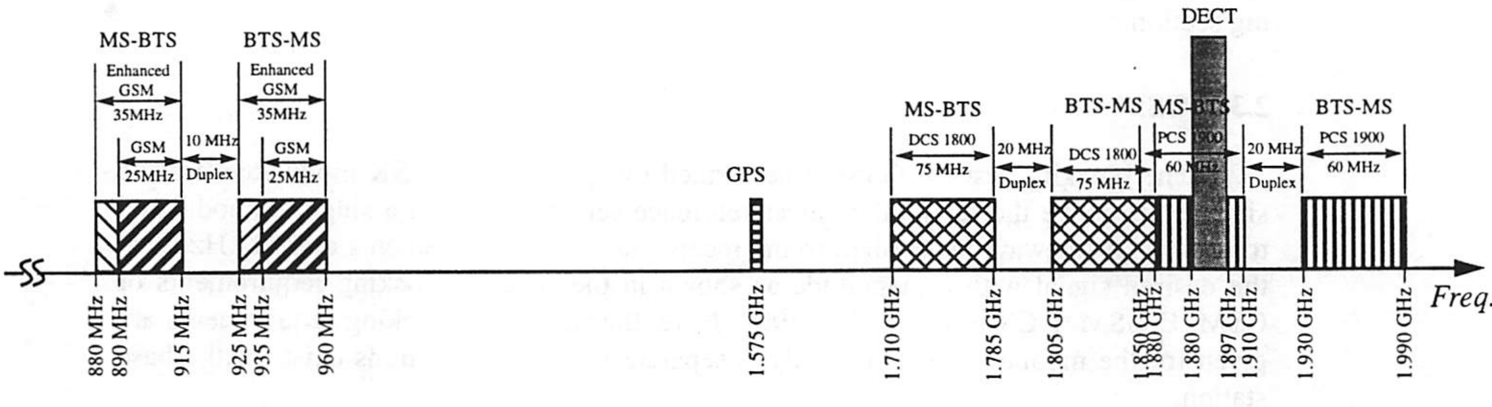


FIGURE 1. Frequency Bands of interest for the (GSM / DECT / GPS) Receiver.

2.2 Sensitivity and Noise Figure Requirements

The sensitivity requirements for the three standards under consideration are shown in table 1 with the corresponding input SNR (See definitions in section 4.1.1), required Carrier-to-Noise Ratio (CNR) to maintain the minimum BER outlined in the standards, along with required NF. Definitions for sensitivity, input noise floor, and the required carrier-to-noise ratio are given in section 4.0.

TABLE 1. Sensitivity and Signal Levels

Standard	Sensitivity (dBm)	Input Noise (dBm)	Input SNR (dB)	Required C/N (dB)	Required NF (dB)
GSM	-102	-120.8	18.8	9	9.8
E-GSM	-102	-120.8	18.8	9	9.8
DCS1800	-100	-120.8	20.8	9	11.8
PCS 1900	-102	-120.8	18.8	9	9.8
DECT	-83	-112.3	29.3	10.3	19.0
GPS	-130	-110.9	-19.0	N/A	14

2.3 Blocking Requirements

A summary of the blocking requirements and the corresponding blocking test that must be performed to comply with DECT, and all version of GSM are given in the following section.

2.3.1 GSM

The blocking test for GSM is performed by applying a GMSK modulated desired signal 3dB above the required receiver reference sensitivity. Then a single unmodulated tone (simple sinewave) is applied to the receiver at discrete increments of 200 kHz from the desired signal with a magnitude as shown in the specific blocking requirements of GSM, E-GSM, DCS1800, and PCS1900. Note, the following blocking requirements are given for the mobile station(MS) only, a separate set of specifications exist for the base station.

The blocking requirements are similar among the different GSM standards with some exceptions which are outlined below. By far, one of the most difficult specifications to meet in GSM is the 3 MHz blocker which is typically 76 dB above the carrier. Other unique features of the blocking specifications are what is referred to as "spurious response frequencies" which are a set of exceptions which relax the requirements in a selected range of frequencies. The frequency of the relaxed blocking requirements are selected by the user and each channel is allowed a different set of spurious response frequencies. For example, if we set a spurious response frequencies for channel 800 in DCS 1800 then move to channel 805 we can again assign a new set of spurious response frequencies.

The blocking test are performed over the entire spectrum for each possible desired channel.

2.3.1.1 GSM 900 (Mobile Station Receiver)

Figure 2 is the blocking profile for GSM 900. Unique features of GSM 900 over other versions of GSM are that all out-of-band blockers are at 0 dBm with exception of the spurious response frequencies. Also, the difference between GSM and Enhanced GSM (E-GSM) is that an extra 10MHz is added to the lower end of both the transmit and receive bands of GSM to form E-GSM. Assuming E-GSM is used, the 3MHz inband blocker extends out 10MHz beyond that last channel on the low end of the mobile station receive spectrum before the first out-of-band blocker begins. Likewise, on the high end of the E-GSM mobile station receive band, a 20MHz guardband exists between the last channel and the beginning of the first out-of-band blocker. The guardbands are important, particularly for integrated receivers where the RF frontend filter will be the only attenuation of the first out-of-band blockers.

Spurious Response Frequencies:

When a spurious response frequency is selected the blocking requirement is relaxed to -49 dBm at the frequency which the blocker is applied. A separate set of spurious response exceptions can be applied to each channel.

6 inband frequencies may be selected with a maximum of three adjacent frequencies assigned as spurious response exceptions.

24 out-of-band spurious response frequencies are allowed with a maximum of three adjacent frequencies assigned as spurious response exceptions.

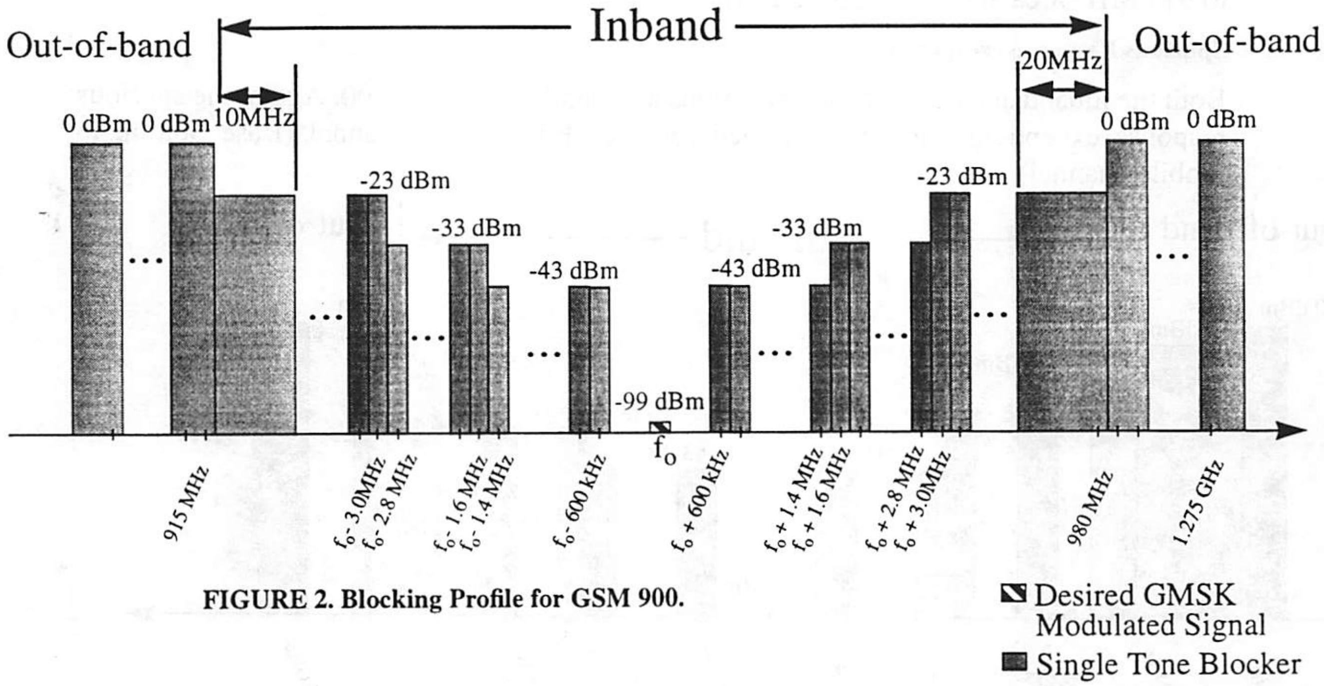


FIGURE 2. Blocking Profile for GSM 900.

2.3.1.2 E-GSM (Mobile Station Receiver)

The blocking profile for E-GSM is shown in figure 4. The blocking mask is identical to GSM 900 with the single exception of a relieved spec. on some of the out-of-band blockers on the lower side of the MS band. Specifically the blocker requirement in the 905 MHz to 915 MHz area are reduced to -5dBm.

Spurious Response exceptions:

Both the inband and out-of-band exceptions are identical to GSM 900. Again, the spurious response exceptions can be reassigned for each BTS-to-MS channel (Base Station to mobile channel).

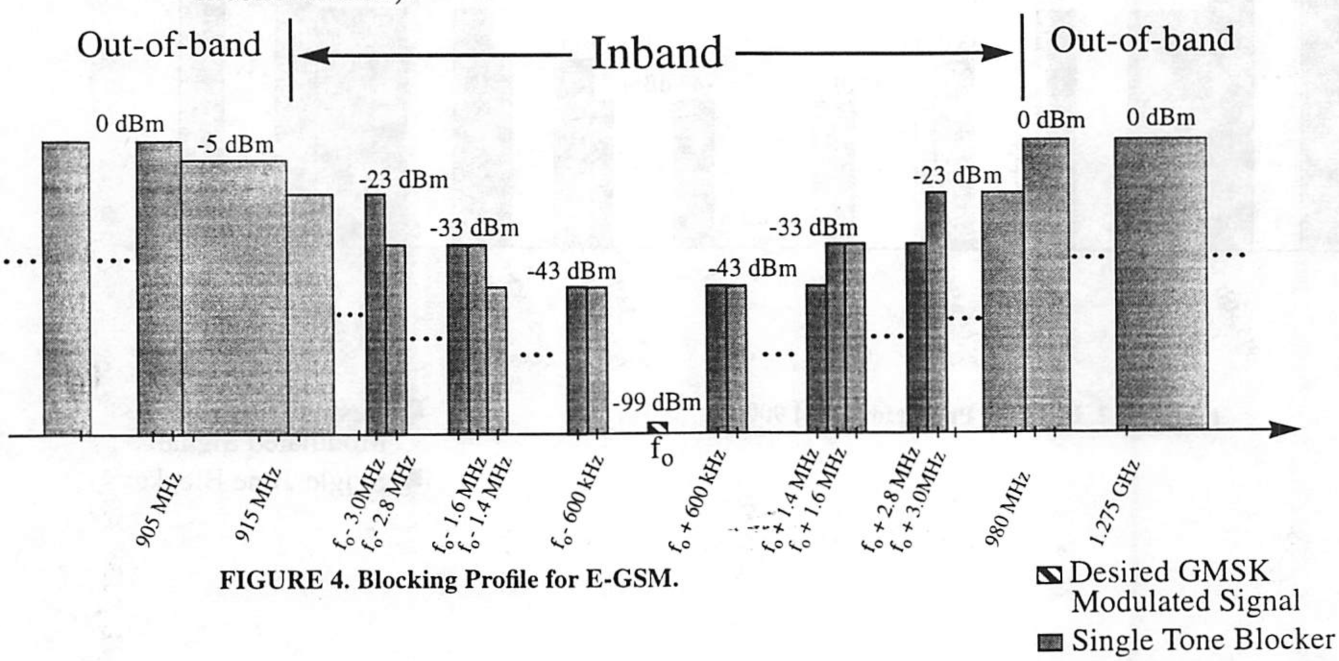


FIGURE 4. Blocking Profile for E-GSM.

2.3.1.3 DCS 1800 (Mobile Station Receiver)

Figure 5 is the blocking mask for DCS 1800. Unique characteristics of this band are that the desired inband signal is set to -97dBm instead of -99 dBm. Another difference is that the two power levels for the out-of-band blockers and the 3 MHz blocker are at -26 dBm instead of -23 dBm making it slightly easier to meet spec. with a -97 dBm desired signal.

Spurious Response Frequencies:

When a spurious response frequency is selected the blocking requirement is relaxed to -49 dBm at the frequency where the blocker is applied. As with the other GSM standards the spurious response exceptions can be reassigned.

12 inband frequencies may be selected with a maximum of three adjacent spurious response exceptions.

24 out-of-band spurious response frequencies are allowed with a maximum of three adjacent frequencies assigned to be spurious response exceptions

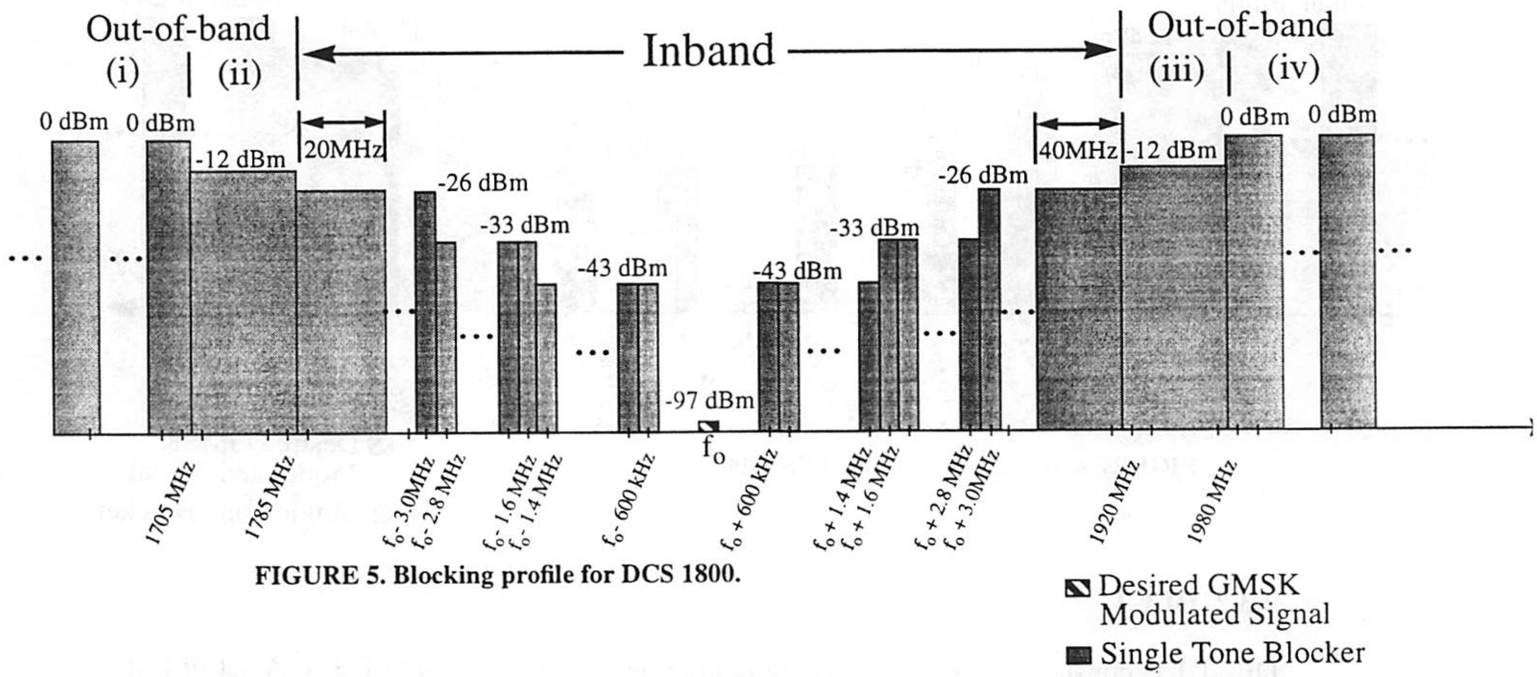


FIGURE 5. Blocking profile for DCS 1800.

2.3.1.4 PCS 1900 (Mobile Station Receiver)

Figure 6 is the blocking profile for PCS 1900. The blocking requirements are very similar to DCS 1800 with the exception that the desired inband signal is at -99 dBm.

Spurious Response Frequencies:

When a spurious response frequency is selected the blocking requirement is relaxed to -43 dBm at the frequency which the blocker is applied. The spurious response exceptions can be reassigned for each channel.

12 inband frequencies may be selected as spurious response exceptions. There appear to be *no limit to the number of adjacent frequencies* which may be selected as an exception.

24 out-of-band spurious response frequencies are allowed. Again, there appear to be *no exceptions* to the number of adjacent bands allocated as spurious response frequencies.

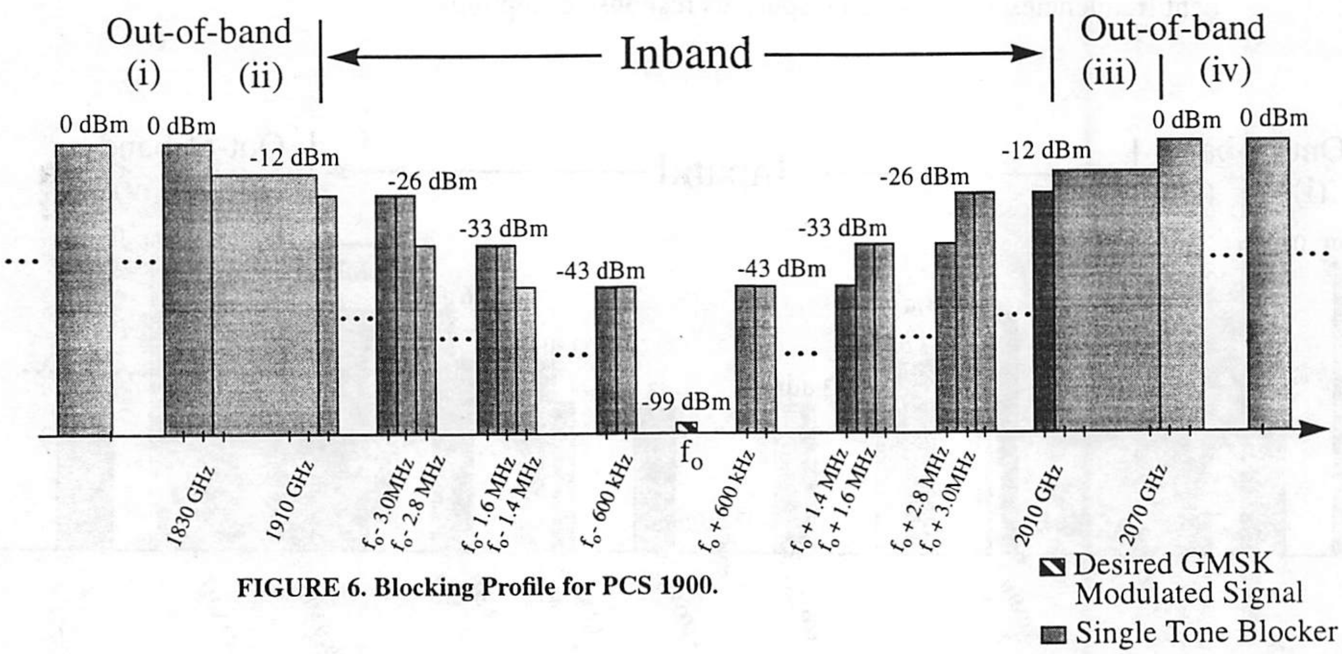


FIGURE 6. Blocking Profile for PCS 1900.

2.3.2 DECT

DECT has considerably easier blocking requirements as compared to GSM. A set of test conditions are given for both inband and out-of-band blocking signals. Later sections which outline the blocker level specifications almost exclusively refer to the GSM blocking requirements as typically the test that defines required performance of the receiver or any component.

2.3.2.1 Inband blocking requirements:

The receiver must maintain a 10^{-3} BER when a -73 dBm desired signal is applied to the receiver and a single blocker is applied to the input of the receiver. The blocker is a GMSK modulated signal of power level given in figure 7. The blocking requirements include a -83dBm Co-Channel blocker (The Co-Channel blocker is an interfering signal applied in the same band as the desired signal). All of the inband blocking tests are

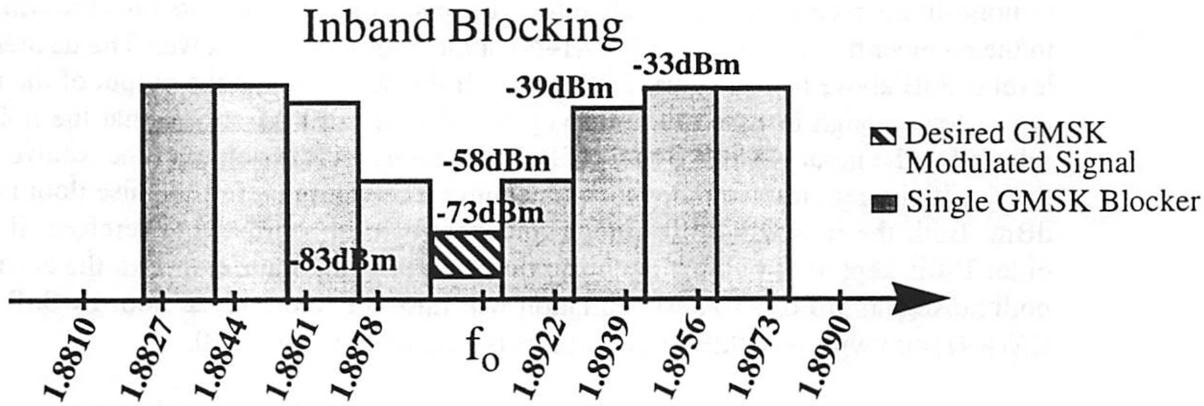


FIGURE 7. DECT Inband blocking requirements

repeated for each of the adjacent channels.

2.3.2.2 Out-of-band Blocking Requirements:

A desired -73dBm input signal is applied to the receiver in channel 4. Then a single unmodulated blocker (simple sinewave) is applied in each of the following bands with the signal strength indicated in figure 9. Since a 10^{-3} BER must be maintained, this maps to an approximate C/I ratio of 10dB at the output of the receiver.

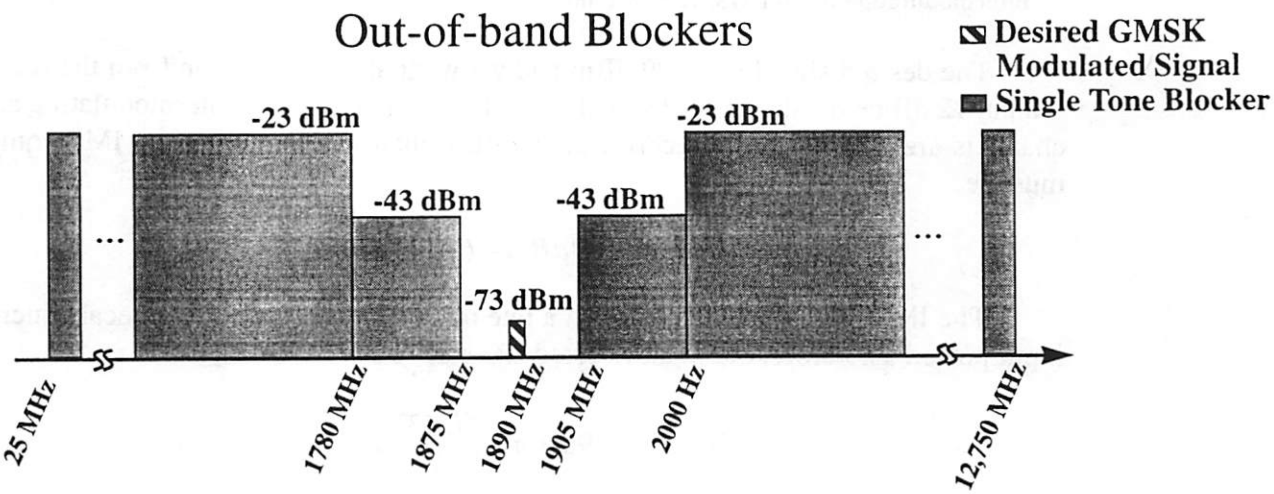


FIGURE 9. Out-of-Band DECT blocking requirements.

2.4 Intermodulation Requirements

2.4.1 GSM 900, E-GSM, DCS 1800, PCS 1900

The adjacent channel immunity test is performed by applying two unmodulated carriers with a power level of -49 dBm to the input of the receiver while a signal 3dB above the reference sensitivity is applied (-99 dBm for GSM 900, E-GSM, and PCS 1900. -97dBm for DCS 1800). The receiver must maintain a 10^{-3} BER or 9dB C/I at the output of the receiver while performing the adjacent channel test. However, this also includes the effects of noise in the receiver channel. Therefore, the distortion components plus the white noise in the receiver degrade the overall C/(I+N) at the output of the receiver. The desired signal level is 3dB above the sensitive requirement. If the noise floor at the output of the receiver is just low enough to pass the sensitivity test than it can be assumed that the noise floor referred to the input is 9dB (to meet 9dB CNR required at the output of the receiver) below the sensitivity requirement, then the maximum receiver input referred noise floor is at -111 dBm. Both the noise and 3rd order components are uncorrelated. Therefore, if the 3rd order IM is kept at or below the noise floor, then the total interference to the carrier from both noise and 3rd order intermodulation will raise the interference floor by 3dB and the C/(I+N) ratio will be at 9dB or better; this is illustrated in figure 10.

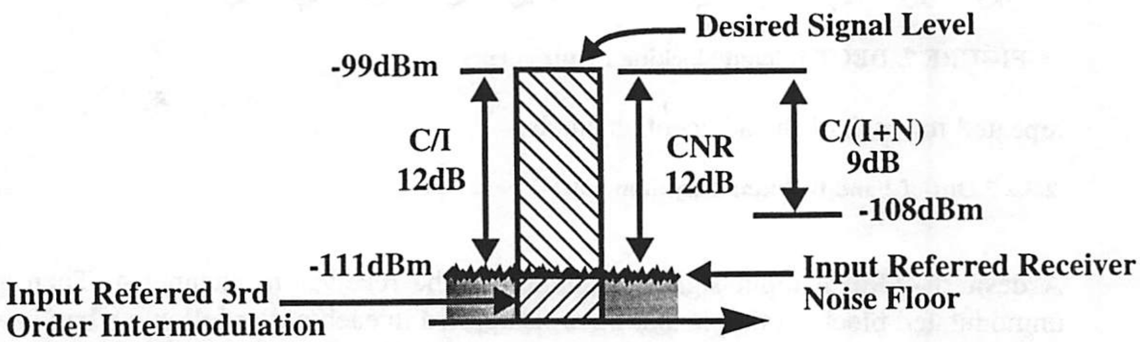


FIGURE 10. Maximum allowable input referred noise and distortion floors under the intermodulation test for GSM, PCS 1900.

The desired signal is at -99 dBm and we want all the distortion from the receiver to remain 12 dB below the desired signal or at -111 dBm. If the two intermodulating adjacent channels are applied to the receiver at -49 dBm then we know that the IM3 component must be,

$$IM3 = -49dBm - (-111dBm) \qquad \text{(EQ 1)}$$

The IM3 component decreases at a rate of 20dB/decade for every decade increase in input power. Therefore, the input referred IP3 can be expressed as,

$$IP3 = -49dBm + \left(\frac{IM3}{2}\right)dBm \qquad \text{(EQ 2)}$$

Which gives us a -18 dBm input referred IP3 or better required of the receiver to be compliant with the GSM standard.

2.4.2 DECT

Similar to GSM, the DECT standard outlines a set of conditions to test the intermodulation performance of the receiver. A desired carrier is applied to the receiver 3dB above the reference sensitivity or -80 dBm. Two adjacent channel signals are applied with a -46 dBm input power. Using the same procedure to calculate the input referred IP3 as in GSM we get an IP3 of,

$$IP3_{DECT} \geq -22dBm \tag{EQ 3}$$

3.0 Proposed Architectures

3.1 Wideband IF w/ Double Conversion

3.1.1 Description

Similar to the architecture used on the DECT receiver [12][13] we are currently considering the Wideband IF with double conversion (WIF) architecture (see figure 11). The requirements of the individual blocks of the Wideband IF approach have been completely specified in section 5.0.

The RF filter does a first order filtering of out-of-band signals. Then all of the potential inband signals are frequency translated to IF with a fixed frequency synthesizer. A simple low pass filter is used before the signal band is frequency translated using an IF channel select synthesizer. For both the GSM and DECT standards, the second mixer stage will modulate all of the channels to baseband. The second Local Oscillator will select the desired channel. The signal is then put through an anti-alias filter before passing through a high dynamic range oversampled ADC. A small image rejection calibration circuit may be used on the GSM receiver. While calibrating (possibly during start-up or between received frames in a TDMA system) an image tone will be synthesized using the local oscillators which are already present. This tone will then be injected into the input of the image rejection mixer and used to tune the quadrature phase of the LOs and possibly the gain between the signal paths.

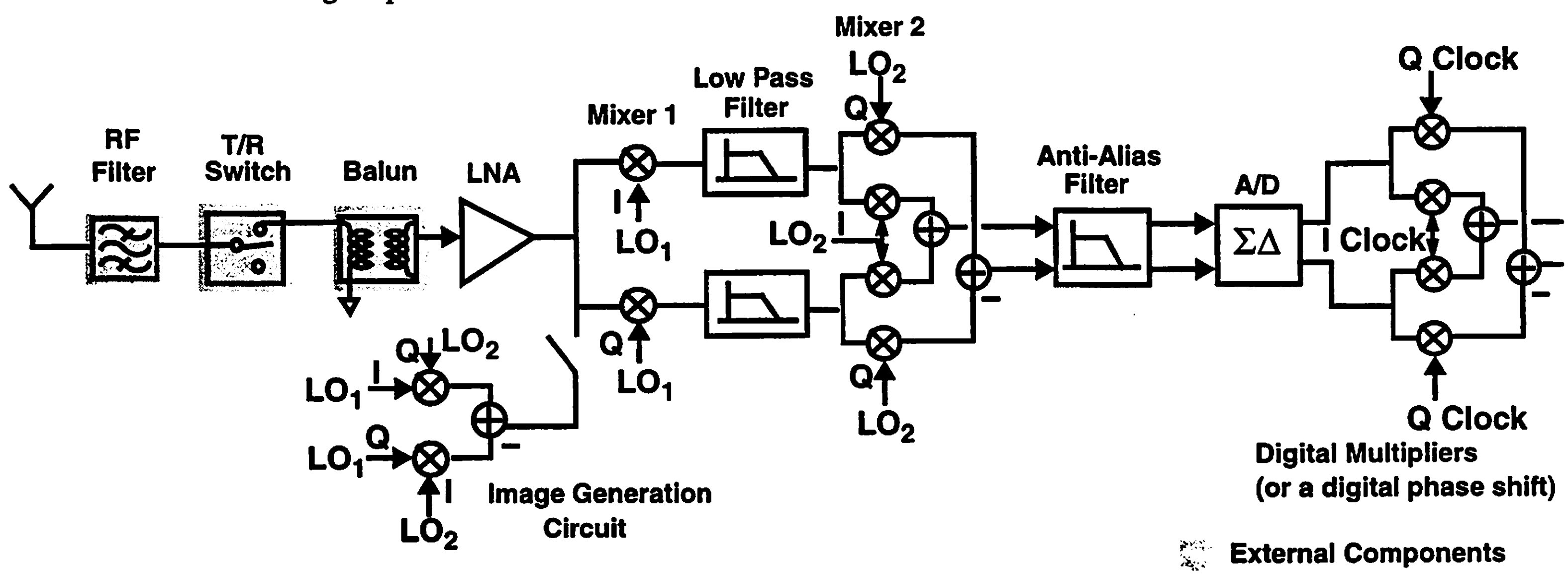


FIGURE 11. Wideband IF w/ Double Conversion Architecture

3.1.2 Pros and Cons of Wideband IF w/ Double Conversion

Pros

• The first, higher frequency synthesizer (LO1) is a fixed frequency synthesizer. Theoretically, it should have a superior phase noise performance over an integrated RF channel select synthesizer because a wide PLL loop bandwidth may be used to shape the phase noise contribution from the VCO. In addition, the integrated solution for a high frequency synthesizer is easier to implement in hardware when the frequency is fixed.

• Neither of the local oscillators are at the frequency of the carrier reducing the risk of LO mixing which degrades the overall receivers dynamic range. There may still be some leakage of the second oscillator to the IF input of the second mixer stage. However, there are no high gain elements in this leakage path (unlike a direct conversion receiver which may include the LNA as part of the LO leakage path). Also, unlike direct conversion, the DC offset produced at the output of the second mixer stage will have a relatively constant amplitude.

Cons

• The two stage mixing makes it particularly difficult to achieve a low noise figure, and low distortion receiver. *For GSM, it is particularly difficult to meet the 3MHz blocking specification with two sets of mixers in the receiver signal path.*

• The image-rejection mixer requires highly accurate phase and gain matching between the quadrature LOs and the signal paths. However, this hopefully would be addressed with a single-sideband mixer with the capability to inject an image tone for calibration purposes.

• The 3MHz blocking condition outlined in the GSM standard becomes particularly difficult to meet in the wideband IF receiver predominantly due to four effects which degrade the C/(I+N) while a blocker is present. 1) Because there is essentially no channel filtering between the receiver input and baseband the 3MHz blocker will now reciprocal mix with the phase noise of two mixer stages. 2) 3MHz blocker creates gain compression in all of the receiver components including the second mixer stage and baseband. 3) The 3MHz blocker will most likely cause the receiver noise floor to rise. Because there is an additionally mixing stage, the potential for the noise floor to rise more increases. 4) Because the signal is mixed to baseband with all of the blockers second order intermodulation may produce a DC component which falls within the desired signal band further degrading the carrier-to-interference ratio.

3.2 Low-IF Single Conversion

A second approach which was suggested by Jeff Weldon (shown in figure 12), is similar to the architecture used by UCLA for their direct conversion receiver. The key idea is to use a single-side band mixer to synthesize a local oscillator using two other oscillators which are most likely produced by two PLLs locked to a crystal. The Berkeley approach to frequency synthesis using a fixed frequency RF LO and a variable IF synthesizer which performs the channel selection could still be used in this approach. Unlike, the

UCLA receiver, we would frequency translate the carrier to a low IF (instead of baseband) using a single set of mixers to avoid the effects of 1/f noise and DC offsets.

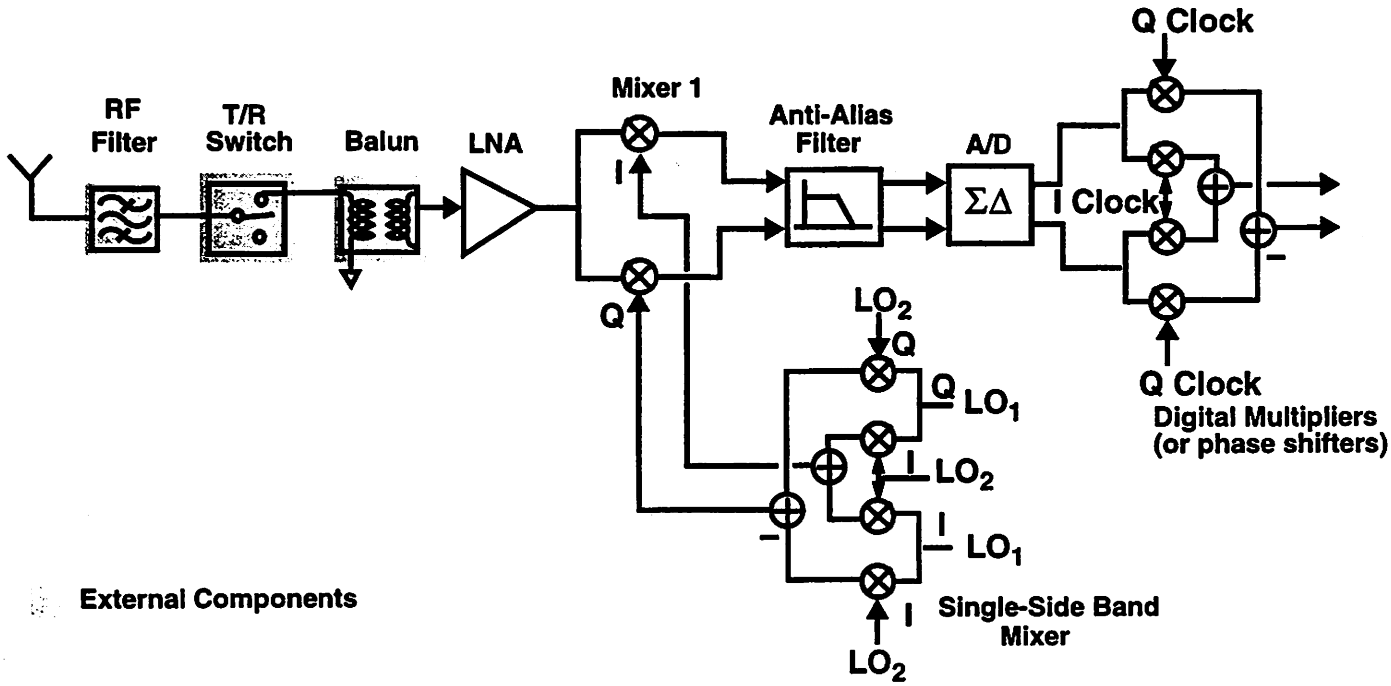


FIGURE 12. Low-IF single conversion.

3.2.1 Pros and Cons of Low-IF Single Conversion.

Pros

• Similar to Wideband IF, the Low-IF architecture can utilized a fixed high frequency LO leaving the channel-select function to be performed at the lower IF frequency.

• There is one less mixer stage in the receiver signal path compared to WIF making the receiver blocking, noise, and intermodulation performance considerable easier to meet aggressive specification like GSM.

Cons

• The local oscillator used by the one stage mixer for frequency translation is created using a single-sideband (SSB) mixer. Both the gain and phase mismatch going through the SSB LO mixer will generate an unwanted sideband at the output on the low side of LO1. This is similar to the image-rejection problem in the Wideband IF receiver.

One practical approach to address the unwanted sideband coming out of the image-rejection mixer is to ensure that it is far enough away in frequency from the desired sideband such that some attenuation in the unwanted band is provided by the RF filter. This may be accomplished by making LO2 high enough in frequency.

• The output of the SSB mixer must be able to drive the capacitive load of the receivers mixers. This node will in all likelihood be a high impedance node which must run at 1.9GHz. Harmonics of both LO_1 and LO_2 may be difficult to deal with in the SSB mixer.

The individual four mixers must be extremely linear and drive an output at high frequency into the mixer input. The SSB individual mixer cells will be the most challenging RF component to design in the Low-IF architecture.

• Practically speaking, the voltage gain of the LNA is limited to approximately 20 dB. In addition, the lower bound of the integrated inband noise referred to the input of the baseband section is approximately 20μV rms. To meet the 9dB receiver noise figure requirement for GSM a considerable amount of gain must be provided before the baseband circuits. It was estimated that a single mixer stage would need to provided a voltage conversion gain of 18dB (see section 5.2.2 and section 6.0) to overcome the noise introduced by the baseband and meet the noise figure requirement. It is somewhat impractical to design a mixer with 18dB of conversion gain and operate at 1.9GHz.

• If this receiver is designed as a low-IF system the ADC may become difficult to design with the required signal bandwidth and resolution. Direct conversion may be used to lower the signal bandwidth, however, this will have the LO leakage problems which plague zero IF receivers.

4.0 Estimate of Integrated Receiver Performance

The following section outlines a procedure that was used to both design and estimate the overall receiver performance for both the Wideband IF architecture and the Low-IF receiver system. The procedure to analyze the noise figure, intermodulation, phase noise requirements on the synthesizers and 1dB compression point were used to generate a set of block level specifications for each component in the receiver. The component level specifications are highlighted in section 5.0; a prediction of the overall receiver performance is given in section 6.0.

4.1 Noise Figure

Much literature has been written giving a detail description of methods to calculate the noise figure of a conventional discrete component receiver. In a typical multi-component receiver, the input and output impedance of the RF blocks are matched to 50Ω. However, modern receivers are becoming increasingly more integrated. In addition, baseband analog circuit designers evaluate a component's performance with respect to voltage gain and input thermal rms noise voltage rather than the power gain and thermal noise power spectral density. In contrast, RF circuit designers prefer designing components using power gain. However, this implies a knowledge of the input and output load impedance of a receiver block which may be difficult or awkward to use when analyzing an integrated receiver block. Therefore, some of the conventional definitions governing noise figure calculations can be confusing. The following is a suggested method to calculate both the sensitivity and the noise figure of an integrated receiver. To gain a better understanding of cascaded noise figure calculations and the relationship to the receiver sensitivity, some of the original definitions of noise figure by Friis (Friis equations) are first explored.

4.1.1 Sensitivity

The true definition of sensitivity is the minimum detectable signal (typically specified in units of dBm) at the receiver input such that there is a sufficient signal to noise ratio at the output of the receiver for a given application. From figure 13, we can see that depending on how the input signal power is interpreted, two different signal levels for the sensitivity may be obtained. The confusion now arises when the input of the receiver is matched to certain impedance; in the simplified example shown below this would be when the real impedance $R_{in} = R_s$. Is the sensitivity defined at V_s (the source generating signal) or is the sensitivity defined by the voltage across the input terminals of the receiver?

"Industry jargon" typically refers to an open-circuit voltage as "hard" and closed circuit voltage as the "soft"[5] definition of sensitivity. True radio-philes prefer the "hard" definition of sensitivity which is with an open circuit input to the receiver and a minimum detectable signal across the input terminals. However, most equipment including receiver

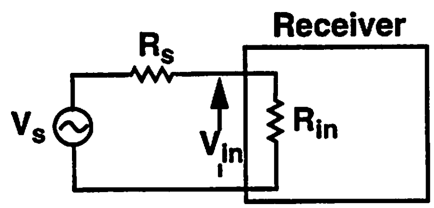


FIGURE 13. Input of the receiver with a source. Vin is the closed circuit voltage while Vs is the open-circuit voltage.

inputs are matched to a 50Ω environment, leading to the more typically used definition of sensitivity as the soft voltage across the input terminal of a block with a matched input impedance. Therefore, the actual sensitivity is defined as the available signal power (definition of available signal power will be given later) delivered to the input terminal of the receiver. The simple definition of sensitivity is the minimum signal power delivered to R_{in} such that a sufficient SNR may be obtained at the output of the receiver to maintain the BER required of the particular radio system. For the purposes of obtaining specifications for the GSM/DECT receiver we will use the "soft" definition of sensitivity which is the same definition used by most commercial standards. To further clarify the definition of sensitivity, assume we have a receiver where the input impedance is matched to a 50Ω source resistance and the receiver sensitivity is -113dBm, then the open-circuit voltage (Vs in figure 13) corresponding to this sensitivity is 1μV[5].

4.1.2 Receiver Noise Figure: Conventional Approach

A good approach to understanding the process of calculating receiver noise figure is to start with the original and definitive paper written by Friis in 1944 [6] which outlines the procedure to analyze the noise figure of a cascaded two port network. Starting as Friis did with a simple example of a source loaded with a 4 terminal device and an output circuit (Figure 14) we can quickly re-derive the noise figure equation. Using this model we now need to define a few terms as Friis did in his original paper.

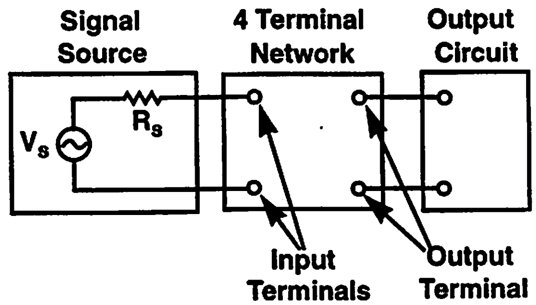


FIGURE 14. Simple Four terminal Network

For maximum power transfer from Vs to the input terminal of the network we need a matched impedance; the power delivered from the source to the input terminals is then $V_s^2/4R$. The power of the signal delivered to the input *under a matched condition* is defined as the ***available signal power*** which is defined as S_g. For a receiver, the available signal power for a sufficient SNR at the output of the receiver is what we defined as sensitivity above. Likewise, the available signal power at the output terminals of the network will be defined as S. Therefore, the ***available power gain*** G of the four terminal device is S/Sg. The available thermal noise power from the source resistance to the input terminals is defined as,

$$\frac{4kTR \cdot \Delta f}{4R} = kT \cdot \Delta f(watts) \qquad \text{(EQ 4)}$$

Note that the *available noise power* at the output of the source is due to the thermal noise source to the left of input terminals and not the noise generated by the input devices of the terminal. A useful number to remember which will aid in rapidly determining the available noise power delivered from the source (or input noise floor) of any receiver under the condition of a matched input impedance is 173.8 dBm/Hz (referenced to 1mW) or -186.8 dBV/Hz(referenced to 1V). Knowing the bandwidth of interest you can quickly calculate the available noise power at the receiver input in dBm using,

$$\text{Noise Floor (dBm)} \cong [-173.8 + 10\log(B)](dBm) \qquad \text{(EQ 5)}$$

or if 50Ωs is assumed, the noise floor in dBV is,

$$\text{Noise Floor (dBV)} \cong [-186.8 + 10\log(B)](dBV) \qquad \text{(EQ 6)}$$

where B is the signal bandwidth.

Next define N to be the available noise power at the output of the 4 terminal device. The noise factor is simply defined as the *available* signal-to-noise ratio at the signal source terminals to the *available* signal-to-noise ratio at the output of the network. A summary of the definitions used by Friis are given below.

kTB : Available noise power from the source
N : Available noise power at the output terminals of the network
Sg : Available signal power at the output of the source
S : Available signal power at the output of the network
F : Noise Factor
NF : Noise Figure, noise factor in dB NF=10log(F).
G : (Available signal power at the output)/(Available signal power at the input)

The noise factor for the 4 terminal network can then be expressed as,

$$F = \frac{\left(\frac{S_g}{kTB}\right)}{\left(\frac{S}{N}\right)} = \left(\frac{S_g}{kTB}\right)\left(\frac{N}{S}\right) \qquad \text{(EQ 7)}$$

which is straight from Friis paper. Using the fact that $G=S/S_g$, we can express equation 7 as,

$$F = \left(\frac{1}{G}\right)\left(\frac{N}{kTB}\right) \qquad \text{(EQ 8)}$$

From equation 8, we can see that the available noise power at the output is simply, N=FGkTB which includes the noise from the signal source. The available noise at the output due to the network only is then,

$$(F-1)GkTB(watts) \qquad \text{(EQ 9)}$$

Applying this same procedure of using available signal and noise powers the same argument can be used for cascaded networks. For example, if we have as Friis presents in his paper, network 1, cascaded with network 2 as shown in figure 15.

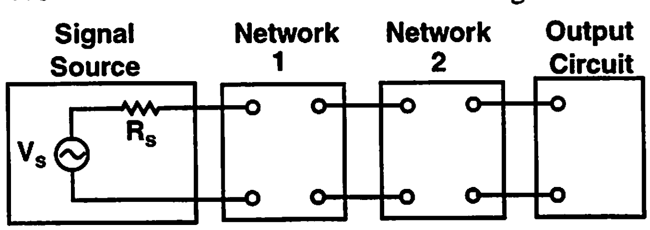


FIGURE 15. Cascade Network

The available noise power at the output terminals of network 2 is,

$$N_{12} = F_{12}G_{12}kTB \qquad \text{(EQ 10)}$$

substituting in the gain for the network 1 and network 2 we can express equation 10 as,

$$N_{12} = F_{12}G_1G_2kTB \quad \text{(EQ 11)}$$

the available noise power at the output of network 2 can be expressed as,

$$N_1 = F_1G_1kTB \quad \text{(EQ 12)}$$

simply multiplying by the gain in network 2 gives the available noise power at the output of network 2 due to noise sources in network 1 or

$$F_1G_1G_2kTB \quad \text{(EQ 13)}$$

from equation 9 we can see that the available noise power due to noise sources in network 2 only is,

$$(F_2 - 1)G_2kTB \quad \text{(EQ 14)}$$

The total available noise power at the output can now be expressed as the sum of the noise sources due to networks 1 and 2 reflected to the output (note that there is a slight error with this equation in Friis paper),

$$N_{12} = F_1G_1G_2kTB + (F_2 - 1)G_2kTB \quad \text{(EQ 15)}$$

Using equation 13, equation 14, and equation 15 we can solve for the overall noise factor of network 1 and 2 with the following result,

$$F_{12} = F_1 + \frac{(F_2 - 1)}{G_1} \quad \text{(EQ 16)}$$

Equation 16 can be generalized even further as the following expression for n networks in cascade [6],

$$F_n = F_1 + \frac{(F_2 - 1)}{G_1} + \ \dots + \frac{F_n - 1}{\prod_{i=1}^{n} G_i} \quad \text{(EQ 17)}$$

For passive filters, we often speak of the insertion loss rather than the power gain. In this case, the available signal power outputted by the source in figure 14 is reduced by the amount of insertion loss. For example, a network where the terminals are shorted together would correspond to a 0dB insertion loss. The noise figure is then equal to the insertion loss.

4.1.3 Integrated Receiver Noise Figure Calculation

Receiver noise figure calculations and specifications for different blocks are easily performed when there is a knowledge of the impedances moving down the receiver chain. With information on the input and output impedance of an individual component the available signal power gain may be determined which in turn facilitates the calculation of an arbitrary number of blocks which are cascaded. Further, because we are not matching the impedance at the output of most integrated receiver blocks it is difficult to assign a noise figure specification to an individual block without a precise knowledge of the impedance. Therefore, the following procedure was used to determine the noise specification of the individual blocks and the overall receiver.

The noise figure calculation for the entire receiver was separate into two parts. First, the integrated portion of the receiver was analyzed in terms of the equivalent noise voltages, and the component voltage gain. Then all of the noise components in the integrated portion of the receiver chain (all the blocks onchip) were referred to the input or interface between the last discrete components at the receiver front-end and the first integrated block, in our case that was input of the LNA. The integrated receiver noise sources referred to the input were then compared to the *available noise power* generated by the source impedance found on the board. Comparing the noise sources at the receiver input to the available noise power of the source can then be used to determine the noise figure of just the integrated portion of the receiver. Next the noise working back to the antenna was determined by simply utilizing Friis equation. A detailed explanation of the two step procedure used to determine the noise figure of the receivers studied in this project is given below:

1) On the front-end of the receiver, the insertion loss of individual discrete components were used to find the available signal power at the input of the chip. Because the input impedance of the LNA is matched to 50Ω, the available noise power at the LNA input can be computed in both dBm and dBV. This available noise power at the LNA input is converted to a noise voltage across the input impedance of the LNA.

2) For the integrated receiver blocks (everything after the LNA), all the noise computations were made with respect to equivalent input noise voltages and resistances, where the equivalent noise resistance is defined as the noise resistance R_{eq} corresponding to rms noise voltage power spectral density where,

$$\frac{\overline{V^2}}{Hz} = 4kTR_{eq} \qquad \text{(EQ 18)}$$

In summary then, the noise figure of the receiver is calculated in two steps using the input of the chip as a boundary where the noise level, signal level, and the SNR are converted from available signal powers to rms noise voltages from which we get Req and the signal voltages as shown in the example receiver, figure 16. The overall noise contribution of the *integrated section of the receiver* is calculated by reflecting the equivalent noise sources along the receiver chain back to the LNA input, then comparing this value with the available noise power delivered to the LNA by the source resistance (50Ω) on the board. The available noise power due to a 50Ω output impedance from the last board component is *kTR50*, or the equivalent noise resistance is simply,

$$R_{board} = \frac{50\Omega}{4} = 12.5\Omega \qquad \text{(EQ 19)}$$

The equivalent noise resistance of each integrated receiver block reflected to the LNA input assumes the example Wideband IF architecture in figure 16,

$$R_{Integrated} = R_{LNA} + \frac{R_{Mixer1}}{Av_{LNA}^{2}} + \frac{R_{Mixer2}}{(Av_{LNA} \cdot Av_{Mixer1})^{2}} + \frac{R_{BB}}{(Av_{LNA} \cdot Av_{Mixer1} \cdot Av_{Mixer2})^{2}} \quad \text{(EQ 20)}$$

The noise factor of the integrated section of the receiver is then,

$$F_{Integrated} = \frac{R_{board} + R_{LNA} + \frac{R_{Mixer1}}{Av_{LNA}^{2}} + \frac{R_{Mixer2}}{(Av_{LNA} \cdot Av_{Mixer1})^{2}} + \frac{R_{BB}}{(Av_{LNA} \cdot Av_{Mixer1} \cdot Av_{Mixer2})^{2}}}{R_{board}} \quad \text{(EQ 21)}$$

where A_v is the voltage gain of the respective receiver components. With the noise factor for the integrated section of the receiver we can now use the more conventional definition to calculate the receiver noise figure with respect to the power of the discrete components at the receiver front-end. Again, referring to the example shown in figure 16, and using the results from equation 21, the overall receiver noise factor including the discrete components is,

$$F_{Receiver} = F_{RFfilter} + \frac{F_{TR}}{G_{RF}} + \frac{F_{Balun}}{G_{RF}G_{TR}} + \frac{F_{Integrated}}{G_{RF}G_{TR}G_{Balun}} \quad \text{(EQ 22)}$$

Where G is the power gain or in this case the insertion loss of the front-end components.

Remembering that the receiver sensitivity was defined as the minimum required available signal-to-noise ratio at the input of the receiver to get a sufficient SNR at the receiver output, an estimate of the receiver's sensitivity based on the overall noise figure and the noise floor at the frontend can now be made,

$$Sensitivity(dBm) = NF_{Receiver}(dB) + CNR_{output}(dB) + NFloor(dBm) \quad \text{(EQ 23)}$$

Where CNR_{output} is the required carrier-to-noise ratio at the receiver output to meet the minimum BER requirements of a standard and *NFloor* is the noise floor defined by equation 5.

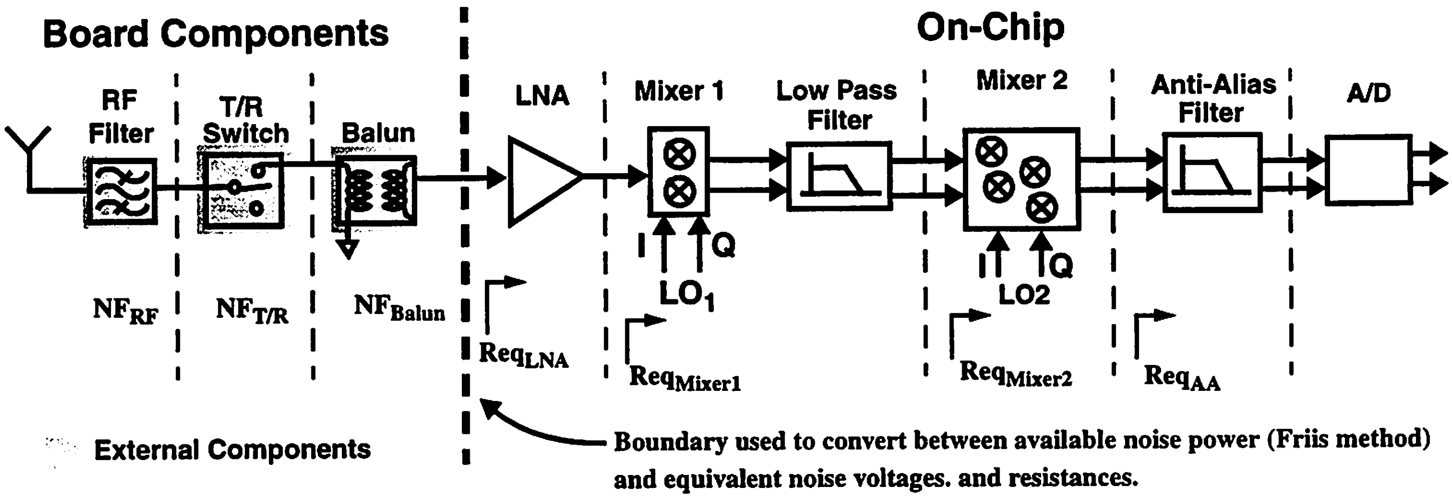


FIGURE 16. Model showing the boundary for the two step noise calculation.

4.2 Intermodulation

There are several methods for calculating the intermodulation performance of an individual block and of a cascaded chain of receiver components. Two methods for calculating the equivalent distortion performance of a number of cascaded receiver blocks as a function of the distortion performance of the individual components are outlined in this section.

4.2.1 3rd Order Intermodulation

A few useful relationships can be obtained by examining a simple plot of an individual component's intermodulation intercept point. Take the example of any generic component with an output 3rd intermodulation intercept point as shown in figure 17.

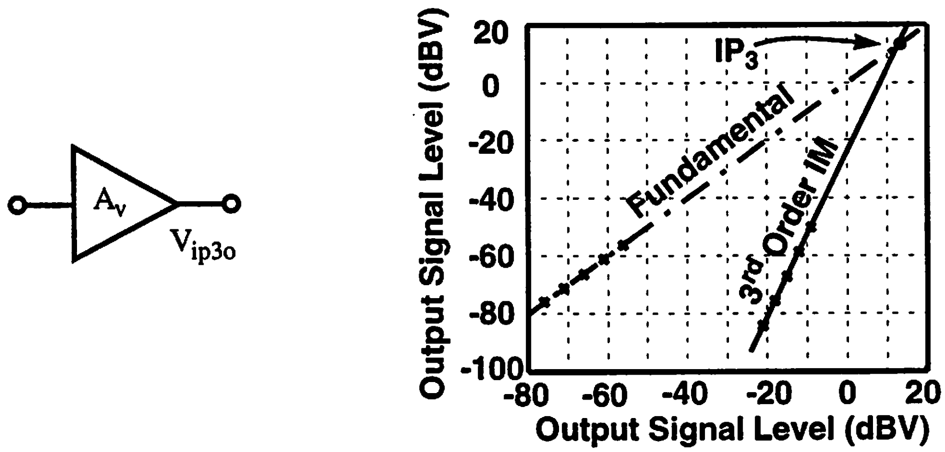


FIGURE 17. Simple Amplifier with voltage gain Av and Output Third order intercept V_{ip3o}

The plot in figure 17 takes a little examination to understand its meaning. The plot is of the output response both the linear and the 3rd order component as a function of the magnitude of the *output* signals which are intermodulating *at the output of the amplifier* (a two tone test). Both the x and y axis are the *output* signal levels in dBV. Therefore, given the output 3rd order intercept point, we can read both the linear component of the intermodulating signals and the 3rd order component produced by the two intermodulating signals as a *function of the output power in* dBV *of the two signals which are intermodulating*. A very useful expression that can determine the magnitude of the 3rd order response at the output or input of a receiver block as a function of the output or input IP3 respectively is [8],

$$V_{o3^{rd}} = \frac{V_{inter(o)}^{3}}{V_{IP3o}^{2}} \qquad \text{(EQ 24)}$$

where V_{o3rd} is the output 3rd order component generated by two adjacent channel interfering signals at the output of the amplifier of magnitude $V_{inter(o)}$, this is illustrated in figure 18. Likewise, at the input of the same amplifier we could write,

$$V_{i3^{rd}} = \frac{V_{inter(i)}^{3}}{V_{IP3i}^{2}} \qquad \text{(EQ 25)}$$

where V_{ip3i}, $V_{inter(i)}$, and V_{i3rd} are the input intercept point, the intermodulating tones, and the input referred third order component respectively, again this illustrated in figure 18.

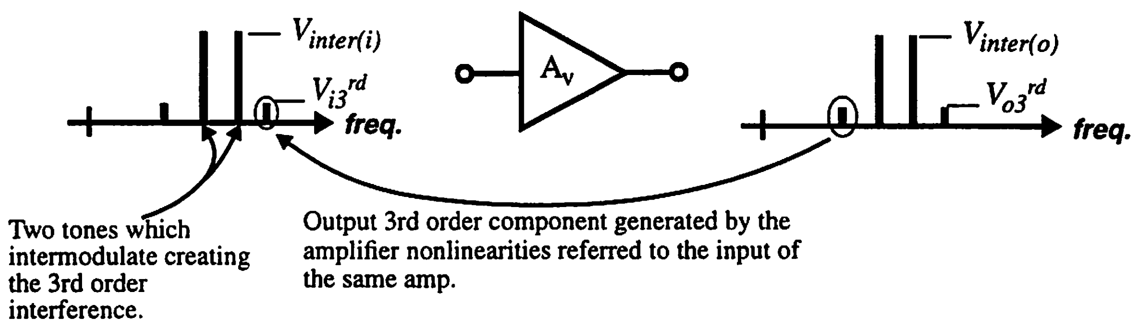


FIGURE 18. Representation of the adjacent channel interferers and the intermodulated 3rd order component which is created.

All variables in equation 24 are related to the input of the simple block shown in figure 17. Equation 24 and equation 25 can be extend to a more generalized expression to describe the distortion components generated by any order intermodulation at a given node in a receiver chain,

$$V_{dn} = \frac{(V_{inter})^{n}}{(V_{IPn})^{n-1}} \qquad \text{(EQ 26)}$$

4.2.2 Intermodulation for cascaded blocks

There are several methods for calculating both an intermodulation interferer at any stage in the receiver and the equivalent Intermodulation Intercept Point (IIP). Again, the example of the 3^{rd} order IM will be used to find the equivalent IIP3 of several cascaded blocks, although, this analysis could easily be extended to any order of intermodulation. The equivalent input or output intercept point of a three stage cascaded network will be

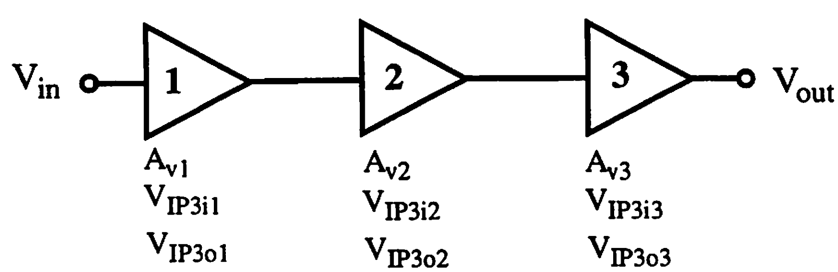


FIGURE 19. Intermodulation in a set of cascaded blocks.

determined, figure 19, as a function of the input or output intercept points of the individual blocks in the chain. A_{vn} is the voltage gain of the n^{th} block and V_{IP3in} and V_{IP3on} are the equivalent input and output voltage intermodulation intercept points respectively of the n^{th} block. Two methods to finding the equivalent intercept point at the output or input of a cascaded network are now explored.

The first and simplest approach is to reflect each of the individual intercept points to either the input or the output of the cascaded blocks and find the minimum term and approximate this as the intermodulation intercept point for the cascaded chain[7].

$$V_{ip3cascade} = min\left(V_{IP3i1}, \left(\frac{V_{IP3i2}}{A_{v1}}\right), \left(\frac{V_{IP3i3}}{A_{v1}A_{v2}}\right)\right) \qquad \text{(EQ 27)}$$

Equation 27 works well when trying to predict the intermodulation performance of a number of cascaded blocks when there is a "weak link" in the chain and one input or output intercept point dominants (much lower in the case of IP3) the cascaded IP3. However, when the individual IP3s contribute somewhat equally to the overall chains linearity performance then equation 27 is not a good approximation.

The second approach attempts to take into account the interaction of the intercept points between the cascaded blocks in the chain. In this approach, the assumption is that the distortion contribution from each of the blocks is uncorrelated, thus their distortion products are independent from block to block. If we write the total 3rd order distortion products at the output of the cascade chain shown in figure 19 we get [8],

$$V_{out}^{3rd} = A_{v3} \cdot A_{v2} \cdot V_{o1}^{3rd} + A_{v3} \cdot V_{02}^{3rd} + V_{o1}^{3rd} \qquad \text{(EQ 28)}$$

where V_{out}^{3rd} is the *total* 3rd order distortion of the cascaded configuration and V_{01}^{3rd}, V_{02}^{3rd}, and V_{03}^{3rd} are the output distortion contributions of each of the blocks. We can now reflect the output distortion to the input to find the equivalent input IP3 of the three cascaded blocks,

$$V_{in}^{3rd} = \frac{A_{v3} \cdot A_{v2} \cdot V_{o1}^{3rd} + A_{v3} \cdot V_{02}^{3rd} + V_{o3}^{3rd}}{A_{v1} \cdot A_{v2} \cdot A_{v3}} \qquad \text{(EQ 29)}$$

expressing each of the 3^{rd} order distortion components using equation 24 we have,

$$\frac{V_{in}^3}{(V_{iIP3cas})^2} = \frac{A_{v2} \cdot A_{v3} \cdot \frac{(V_{o1})^3}{(V_{IP3o1})^2} + A_{v3} \cdot \frac{(V_{02})^3}{(V_{IP3o2})^2} + \frac{(V_{o3})^3}{(V_{IP3o3})^2}}{A_{v1} \cdot A_{v2} \cdot A_{v3}} \qquad \text{(EQ 30)}$$

As noted previously, the assumption is that the distortion components generated from different blocks are uncorrelated which may not be necessarily be true and the potential exist for cancellation of the third order distortion from stage to stage. However, within an individual block the distortion is correlated between the input and output. Therefore, the output IP3 for an individual block may be reflected back to the input of the same block, or $V_{IP3in}=V_{IP3on}/A_{vn}$ for the n^{th} stage. Also, we can now express all of the output voltages in terms of V_{in} and the voltage gain A_v of a block. Equation 30 now becomes,

$$\frac{V_{in}^3}{(V_{iIP3cas})^2} = \frac{A_{v2}\cdot A_{v3}\cdot\frac{(A_{v1}\cdot V_{in})^3}{(A_{v1}\cdot V_{IP3i1})^2} + A_{v3}\cdot\frac{(A_{v1}\cdot A_{v2}\cdot V_{in})^3}{(A_{v2}\cdot V_{IP3i2})^2} + \frac{(A_{v1}\cdot A_{v2}\cdot A_{v3}\cdot V_{in})^3}{(A_{v3}\cdot V_{IP3i3})^2}}{A_{v1}\cdot A_{v2}\cdot A_{v3}} \tag{EQ 31}$$

Cancelling terms we get the familiar form of,

$$\frac{1}{(V_{iIP3cas})^2} = \frac{1}{(V_{IP3i1})^2} + \frac{{A_{v1}}^2}{(V_{IP3i2})^2} + \frac{(A_{v1}\cdot A_{v2})^2}{(V_{IP3i3})^2} \tag{EQ 32}$$

The total input referred IP3 for the cascaded configuration shown in figure 19 is,

$$V_{iIP3cas} = \frac{1}{\sqrt{1/(V_{IP3i1})^2 + {A_{v1}}^2/(V_{IP3i2})^2 + (A_{v1}\cdot A_{v2})^2/(V_{IP3i3})^2}} \tag{EQ 33}$$

A similar analysis reveals that for a two stage cascaded network the equivalent output IP3 can be expressed as,

$$V_{oIP3cas} = \frac{A_2\cdot V_{IP3o1}\cdot V_{IP3o2}}{\sqrt{(V_{IP3o2})^2 + (V_{IP3o1})^2}} \tag{EQ 34}$$

and the equivalent input IP3 of the same two stage network is,

$$V_{iIP3cas} = \frac{V_{IP3i1}\cdot V_{IP3i2}}{\sqrt{(V_{IP3i2})^2 + (A_{v1}\cdot V_{IP3i1})^2}} \tag{EQ 35}$$

Both equation 34 and equation 35 can be used recursively to obtain the equivalent IP3 at any node in a cascaded chain of receiver components. Using either equation 34 or equation 35 in conjunction with equation 24 the equivalent 3rd order distortion component which is seen as an interferer to the desired signal may be found at any point in the receiver chain. The 3rd order inference can then be added with the rms noise in the receiver chain to find the carrier-to-(noise-plus-distortion) ratio at any point in the receiver chain; this is done in section 6.0 to find the C/I ratio for the adjacent channel intermodulation test outlined in the GSM 5.05 specification[1].

4.2.3 2nd Order Intermodulation

One unique challenge associated with integrated radio receivers that downconvert the carrier to baseband without any channel filtering is the effect of second order intermodulation. This is a particular problem for all forms of GSM which have an aggressive blocking performance required of the receiver. The mechanism for second order intermodulation is outlined below with an explanation of the method used to find the required equivalent input IP2 of the baseband blocks. The effect of IM2 is quickly understood by examining a simple expression which relates the input and output signal of a block via a high order transfer function. First, assume looking into the baseband there is a non-linear transfer function relating the input and output signals by,

$$S_o(t) = a_1 S_i(t) + a_2 S_i^2(t) + a_3 S^3{}_i(t)\ldots \qquad \text{(EQ 36)}$$

where $S_o(t)$ is the output signal and $S_i(t) = S_i \cos(\omega_{bl} t)$ represents the input signal applied at ω_{bl} to any arbitrary block. Using a simple trigonometric relationship reveals that when the input signal passes through the second order non-linearity the following results,

$$a_2 \cdot (S_i \cos(\omega_{bl} t))^2 = a_2 \cdot S_i^2 \left(\frac{1 + \cos(2\omega_{bl} \cdot t)}{2}\right) \qquad \text{(EQ 37)}$$

From equation 37 it can be seen that the second order non-linearity creates a DC component. This is a particular problem when there is a weak desired signal which gets frequency translated to baseband in the presence of a strong adjacent channel blocker. The large blocker now creates an interfering component at DC which is at the center of the desired signal spectrum in the baseband blocks. Now it is desired to understand the relationship between the required equivalent input IP2 and the blocker specification given for a standard or application. From course notes given in [9], the relationship between the coefficients of the high order signal transfer function, HD2, and IM2 are given by.

$$HD2 = \frac{a_2 S_i^2 / 2}{a_1 S_i} \qquad \text{(EQ 38)}$$

$$IM2 \cong HD2 + 6dB \qquad \text{(EQ 39)}$$

Typically, the required IP2 performance may be inferred from a knowledge of the gain which precedes the baseband and blocking test which must be performed to comply to a standard. Again, similar to the required noise figure and phase noise performance, the required IP2 performance for the DCS 1800 mode of operation is far more aggressive than what is called for by DECT. Accordingly, as an example calculation the 3MHz blocking condition will be used to find the required IP2 performance in the DCS 1800 mode of operation.

To find the required IP2 it will be first assumed that both the desired signal and the blocker are frequency translated to baseband without any filtering of the adjacent channel blocker; a fair assumption for any receiver which attempts to eliminate the IF filter. In addition, both the carrier and the blocker will see an equal gain by all components which precede the baseband. This gain will be denoted A_{vrf} which is the comprehensive gain between the antenna and the baseband blocks. In the DCS 1800 standard, under the 3MHz blocking condition, the relationship between the desired baseband signal, the 3MHz blocker, and the interference component generated by the second order intermodulation are as shown in figure 20.

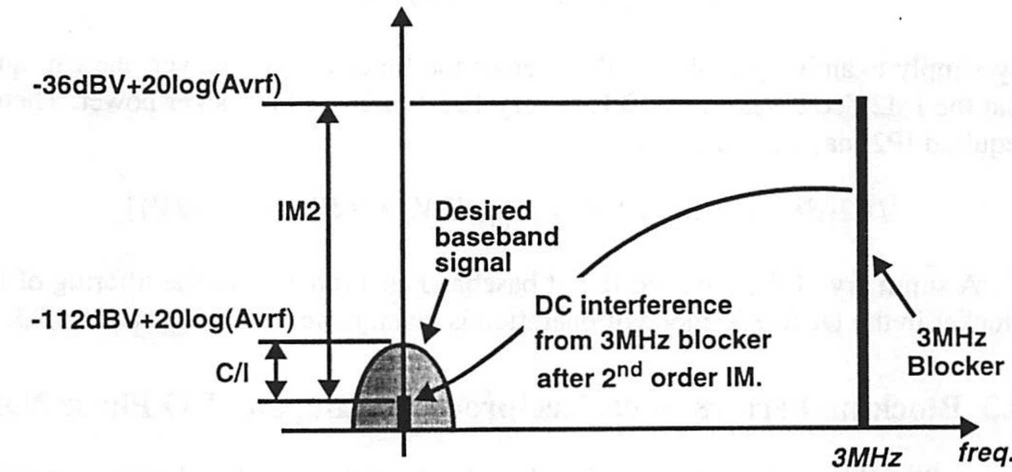


FIGURE 20. Signal spectrum after the second mixer in the wideband IF receiver during the GSM 3MHz blocking test. An interferer is created within the signal band by the 3MHz blocker passing through the baseband 2rd order nonlinearities.

The DC component which is generated by the second order intermodulation of the blocker can be found from equation 37. In equation 38, the numerator of the expression for HD2 is equivalent to the magnitude of the DC component produced by second order intermodulation of the blocker. To ensure that the second order interference has negligible degradation to the overall receiver C/I ratio under the 3MHz blocking condition, the DC component generated by the second order intermodulation must be at least 15dB below the desired baseband signal. Therefore, let S_{des} represent the power of the desired signal at the receiver input in dBV and S_{bl} the magnitude of the blocker at the receiver input, also in dBV. Then, to have negligible degradation of the C/I ratio from the DC component created by 2nd order IM,

$$a_2 S_i^2/2 \le S_{des}(dBV) + 20\log(A_{vrf})(dB) - 15dB \qquad \text{(EQ 40)}$$

The magnitude of the blocker at baseband is,

$$S_{bl}(dBV) + 20\log(A_{vrf})(dB) \qquad \text{(EQ 41)}$$

A closer examination will reveal the required HD2 under the conditions outlined above is simply the difference between equation 40 and equation 41 in dB. Therefore, when the interference is required to be 15dB below the desired signal, HD2 can be expressed as,

$$\begin{aligned} HD2(dB) = {} & S_{des}(dBV) + 20\log(A_{vrf})(dB) - 15dB \\ & - [S_{bl}(dBV) + 20\log(A_{vrf})(dB)] \end{aligned} \qquad \text{(EQ 42)}$$

or,

$$HD2(dB) = S_{des}(dBV) - 15(dB) - S_{bl} \qquad \text{(EQ 43)}$$

or the required IM2 performance of the baseband is,

$$IM2 = S_{des}(dBV) - 9(dB) - S_{bl}(dBV) \qquad \text{(EQ 44)}$$

By simply examining a plot of IM2 verses the blocker input power one can quickly see that the IM2 decreases by 10dB for every 10dB increase in blocker power. Therefore, the required IP2 may be expressed as,

$$IP2(dBV) = S_{bl}(dBV) + [S_{bl}(dBV) + 9dB - S_{des}(dBV)] \qquad \text{(EQ 45)}$$

A summary of the required IP2 at baseband as a function of the filtering of the 3MHz blocker in the DCS1800 mode of operation is given table 5 found in section 5.3.

4.3 Blocking Performance, Reciprocal Mixing and LO Phase Noise

Signals present in bands other than the desired channel will create spurious signals which fall within the signal band resulting in a degradation of the C/I (carrier-to-interference ratio) at the receiver output. Selectivity is a measure of a receivers immunity or ability to handle signals outside of the desired band. Intermodulation of one or more adjacent channel signals creating a spurious signal or interferer within the desired signal band is an example of one mechanism which degrades selectivity. The blocking performance of a receiver is another measure of the overall selectivity.

4.3.1 Mechanism of reciprocal mixing with the blocker

A Local Oscillator (LO) is used with a mixer in the receiver signal path to frequency translate the desired signal spectrum about the carrier to a lower frequency. Phase noise is a measure of the spectral purity of the local oscillators used in this operation. Figure 21 illustrates how undesired sideband energy from the local oscillator (phase noise) "reciprocal mixes" with adjacent channels or out-of-band signals which potentially manifest itself as an inteferer within the desired signal band. The oscillator must be designed such that under a worst case blocking condition, the reciprocal mixing of the blocker with the phase noise of the oscillator will produce an interference component far below the desired signal level.

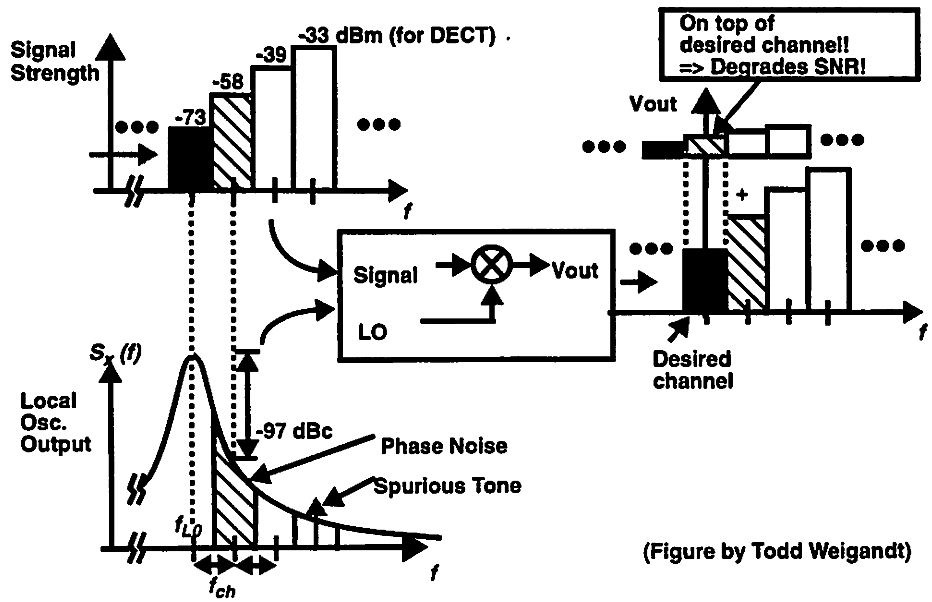


FIGURE 21. Reciprocal mixing of the blocker and phase noise

4.3.2 Calculation of required phase noise performance

One method to perform the phase noise calculation is to assume that the receiver channel is noiseless and the only interference produced within the signal band moving through the receiver chain is due to the phase noise reciprocal mixing with out-of-signal band blockers [10]. Typically, the RF standards specify the magnitude of the blocker (see section 2.0 for standards under consideration) along with the desired signal and a required BER which translates to a minimum C/I ratio at the output of the receiver. The simplest method to calculate the phase noise performance required of the oscillator is illustrated in Figure 22. Here, the phase noise is assumed to be flat across the band of interest at a certain offset from the carrier. The interference component that is then produced when the blocker mixes with the phase noise sidebands is then compared to the desired signal which mixes with the carrier energy. Based on the required C/I ratio at the output of the mixer, the blocker level along with the position in frequency relative to the desired signal and the desired signal level, the required phase performance in dBc/Hz may be estimated using,

$$PN(\Delta f_c)\left(\frac{dBc}{Hz}\right) = \tag{EQ 46}$$

$$S_{bl}(\Delta f_c)(dBm/dBV) - S_{desired}(dBm/dBV) - C/I_{min}(dB) - 10\log(BW))$$

where $PN(\Delta f_c)$ is the phase noise in dBc/Hz Δf_c away from the carrier, S_{bl} is the magnitude of the blocker in dBm or dBV while $S_{desired}$ is the magnitude of the desired carrier in dBm or dBV. C/I_{min} is the minimum required carrier-to-interference ratio and BW is bandwidth of the desired signal.

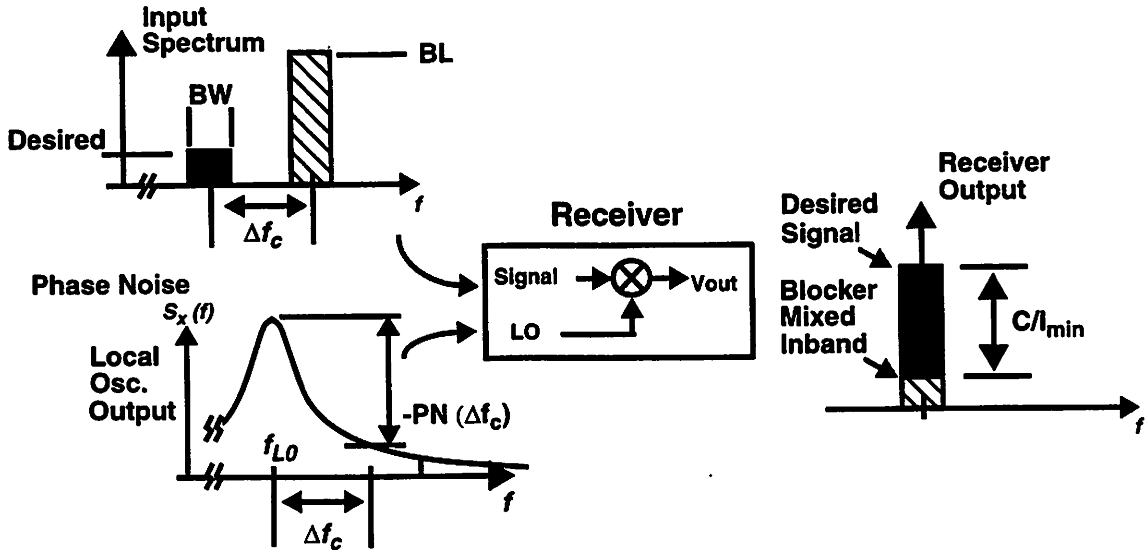


FIGURE 22. Simple calculation for required phase noise performance of the LO.

As stated before, equation 46 can be used to approximate the required phase noise performance of the local oscillators in the receiver assuming there are no other sources of interference in the receiver channel. However, practically speaking this is far from the true situation and the receiver noise contribution will further degrade the overall carrier-to-interference ratio at the output. Therefore, a better picture of the true C/I ratio at the output of the receiver should include the white noise added to the desired signal band as well as the effects of blockers reciprocal mixing with the phase noise and the effects of gain compression in the receiver signal path due to a large blocking signal which will be present.

The procedure to determine the receiver phase noise performance for this project is as follows. All of the equivalent input noise resistances were referred to the output of the receiver including the noise contribution at the input of the receiver (12.5Ω). Then at each mixer output, the power of the blocker signal reciprocal mixing with the LO phase noise creating an interferer within the desired signal band can be approximated by assuming the phase noise is flat across the band of interest. This gives the following expression,

$$10\log(\sigma^2_{Mixerout}) = [S_{bl}(\Delta f_c)(dBV) - [PN(\Delta f_c) + 10\log(BW)]](dBV) \quad \text{(EQ 47)}$$

where $\sigma^2_{Mixerout}$ is the power of interferer created inband. This interference source can then be referred to the output of the receiver along with all other interferers in the receiver chain including the noise contribution of the individual receiver components. This procedure is illustrated for the Wideband IF architecture shown in figure 23.

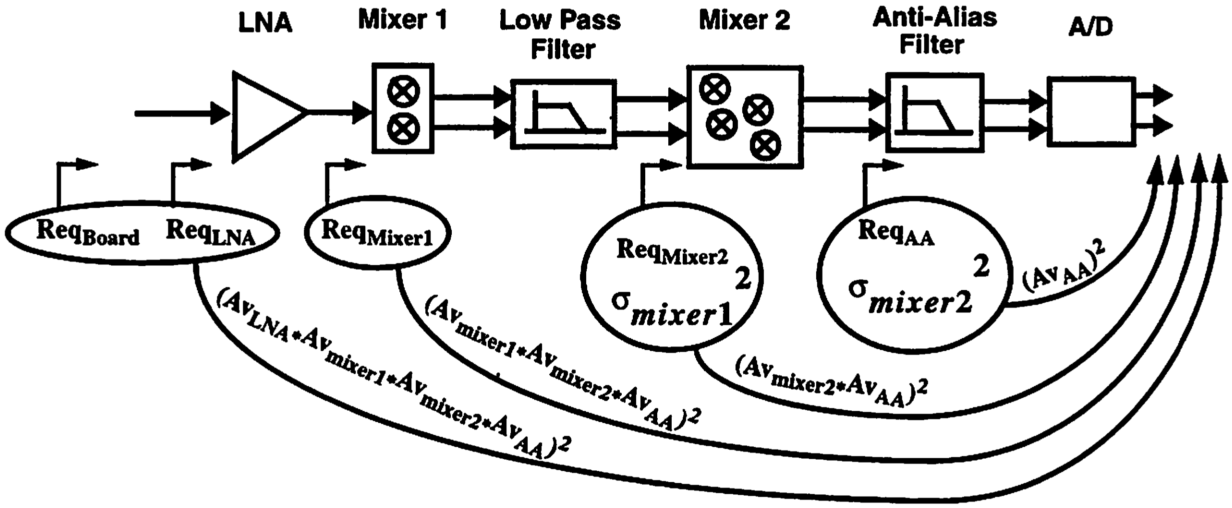


FIGURE 23. Sources of interference in the receiver signal path while an undesired blocker is present.

Treating the interference produced by the blocker reciprocal mixing with the phase noise of the LOs as a rms noise voltage source we can reflect all of the interference sources, both thermal and phase noise to the output of the receiver. The total interference can be expressed in terms of the equivalent input noise resistances (Req) of each block, the individual block voltage gain and the blocking interference created by the phase noise and blocker of each mixer. Defining σ^2_{out} as the total inband voltage interference power (both thermal and reciprocal mixing) at the output of the receiver we have,

$$\sigma^2_{out} = (Av_{Mixer2} \cdot Av_{AA})^2 \cdot \sigma^2_{mixer1} + (Av_{AA})^2 \cdot \sigma^2_{mixer2} + \sigma^2_{thermal} \qquad \text{(EQ 48)}$$

where $\sigma^2_{thermal}$ represents *all* of the thermal noise contributions referred to the output of the receiver which can be found using equation 20 and scaling the equivalent output noise resistance by 4kT to get the rms noise voltage. The receiver output C/(I+N) can be expressed as,

$$C/I_{output} = \frac{Av^2 \cdot S_{Desired}}{(Av_{Mixer2} \cdot Av_{AA})^2 \cdot \sigma^2_{mixer1} + (Av_{AA})^2 \cdot \sigma^2_{mixer2} + \sigma^2_{thermal}} \qquad \text{(EQ 49)}$$

where Av is the overall receiver voltage gain and $S_{Desired}$ is the rms voltage power of the desired signal. Next, if we make the simplifying assumption that the power of the blocker reciprocal mixing with the phase noise of the individual mixers contribute equally (both mixers have the same phase noise profile) and the thermal noise contributions of each receiver component has been determined, we can find the maximum interference allowed by the phase noise mixing with the blocker,

$$\sigma^2_{pn} = \frac{\left(\frac{Av^2 \cdot S_{Desired}}{C/I_{required}}\right) - \sigma^2_{thermal}}{((Av_{Mixer2} \cdot Av_{AA})^2 + (Av_{AA})^2)} \quad \text{(EQ 50)}$$

Using equation 50 in conjunction with equation 47, we can determine the required phase noise performance of the local oscillators to meet the blocking profile for a particular standard.

5.0 Receiver Component Specifications

With a knowledge of the procedure used to find the required performance of each individual block in the receiver chain from the previous sections, the specifications for two receiver architectures are given in this section. Each block in the receiver is specified with requirements for maximum equivalent input thermal noise resistance contribution (R_{eq}), the equivalent input 3[rd] order intercept *voltage*, the voltage gain A_v, and the maximum differential output zero-to-peak voltage. The equations given in section 4.0 were tabulated using microsoft excel. A complete receiver design was done for both the Wideband IF architecture and the Low-IF with single conversion system. It is probably worth noting that in the early stages of this project it was decided to use wideband IF rather than Low-IF. Therefore, the component level specifications have been considerably revised for the wideband-IF system (at least five revisions), while there is only one set of specs generated for the low-IF receiver.

By far the most difficult set of specifications to meet were those of the GSM standard. In particular, there exists a trade-off between the receiver noise figure and the blocking performance. The 3MHz blocker (-23dBm at the input) puts an upper limit on the amount of gain that can be used in the RF front-end making it more difficult to meet the noise figure requirement. Several iterations were performed in excel to obtain an initial set of specifications on the individual blocks.

A summary of the individual receiver component specifications are given in section 5.1 through section 5.4. Illustrations of the overall receiver performance based on the individual block specifications are given in section 6.0 for both the wideband IF architecture and Low-IF with single conversion system.

5.1 Low Noise Amplifier

The low noise amplifier can have either a differential or single ended input. The input must be matched to a 50Ω impedance. The LNA requirements are summarized in table 2. To accommodate the large range of signal levels associated with the desired carrier

TABLE 2. Specifications for the Low-Noise-Amplifier.

Arch.	R_{eq}	V_{IP3i}	$V_{o\text{-}pmax}$(Output)	A_v	P_{-1db}	DR	Power
WIF	< 11 Ω	> 100 mV	190 mV	22 dB	-27dBV	120 dB	min.
Low-IF	< 11 Ω	> 200 mV	188 mV	22 dB		120 dB	min.

in GSM, the LNA will need a bypass switch. Therefore, this will allow the LNA gain to switch between 22dB and 0dB voltage gain. Also, the output of the LNA will see the loading of two large switches used for the mixer adaptation algorithm and the transmit path tuning network.

5.2 Mixer Specification

As mentioned previously, a decision was made early in the project to implement the wideband IF with double conversion architecture for the GSM/DECT receiver. The wideband IF system performs a two step frequency translation of the carrier to baseband. Because the frequency translation is performed in multiple steps, a method to perform image-rejection is required. The dual conversion approach facilitates the implementation of an image-rejection mixer configuration which is similar to the weaver method introduced in 1956[11]. A collection of six mixers are used to implement the image-rejection function in the wideband IF system and modulate the carrier to baseband producing quadrature channels, as shown in figure 24.

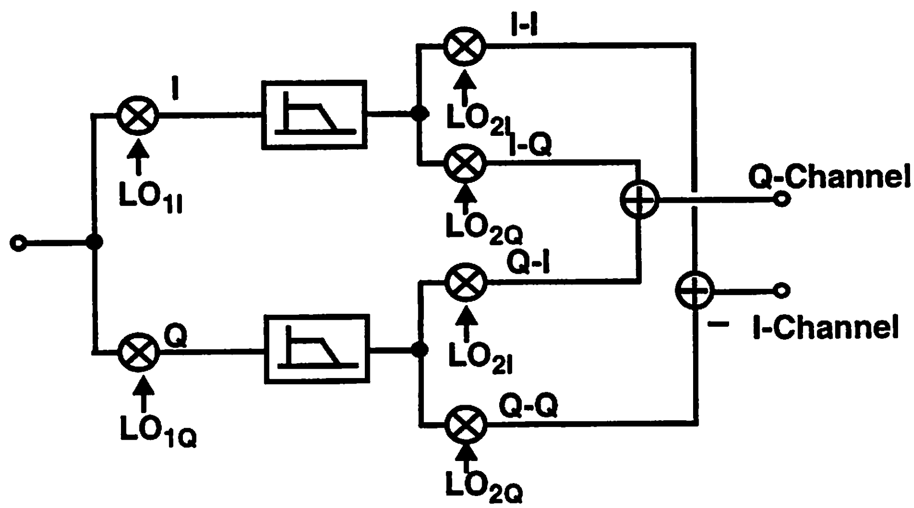


FIGURE 24. Weaver Method used to implement the mixing configuration of the wideband IF architecture.

Four mixers are required to produce a single channel at baseband. Developing a specification for the individual mixers inside the four mixer configuration requires a little understanding of how the noise and signal pass through the mixers. Shown in figure 25 is the four mixer configuration used to produce one channel at baseband. The individual mixers are specified with respect to the voltage conversion gain, the equivalent input noise resistance and maximum output swing for one of the four individual mixer cells shown in figure 25. A bit of analysis is required to translated the conversion gain and noise performance of an individual mixer cell to that of the gain and noise performance for the composite image-rejection mixer configuration.

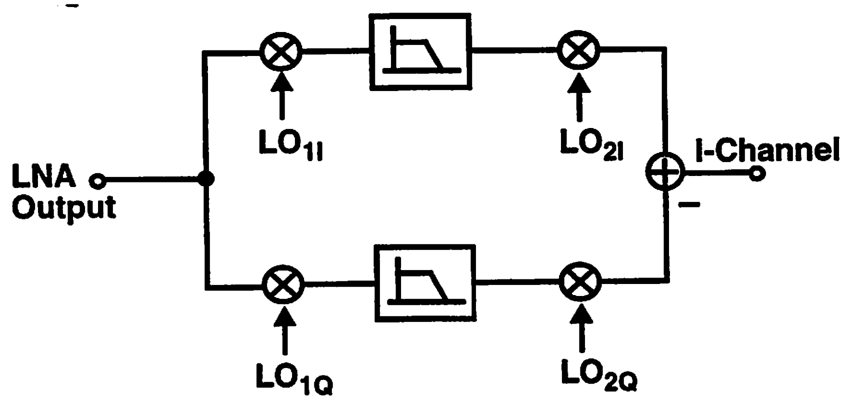


FIGURE 25. Single channel IR-Mixer used by the wideband IF receiver

In figure 26, a model is given for the transformation of the noise for the individual mixers to the noise produced by a single channel in the receiver. R_{eqmrf}, and R_{eqmif} are the equivalent noise resistances of the RF-to-IF mixers and the IF-to-Baseband mixers respectively. Both A_{vrf} and A_{vif} are the voltage conversion gains of the desired carrier from RF to the output of the IF lowpass filter, and the desired signal gain from the mixer input at IF to the mixer output just before the summation of the two signal paths taking into account the signal gain acquired when mixing from IF to baseband.

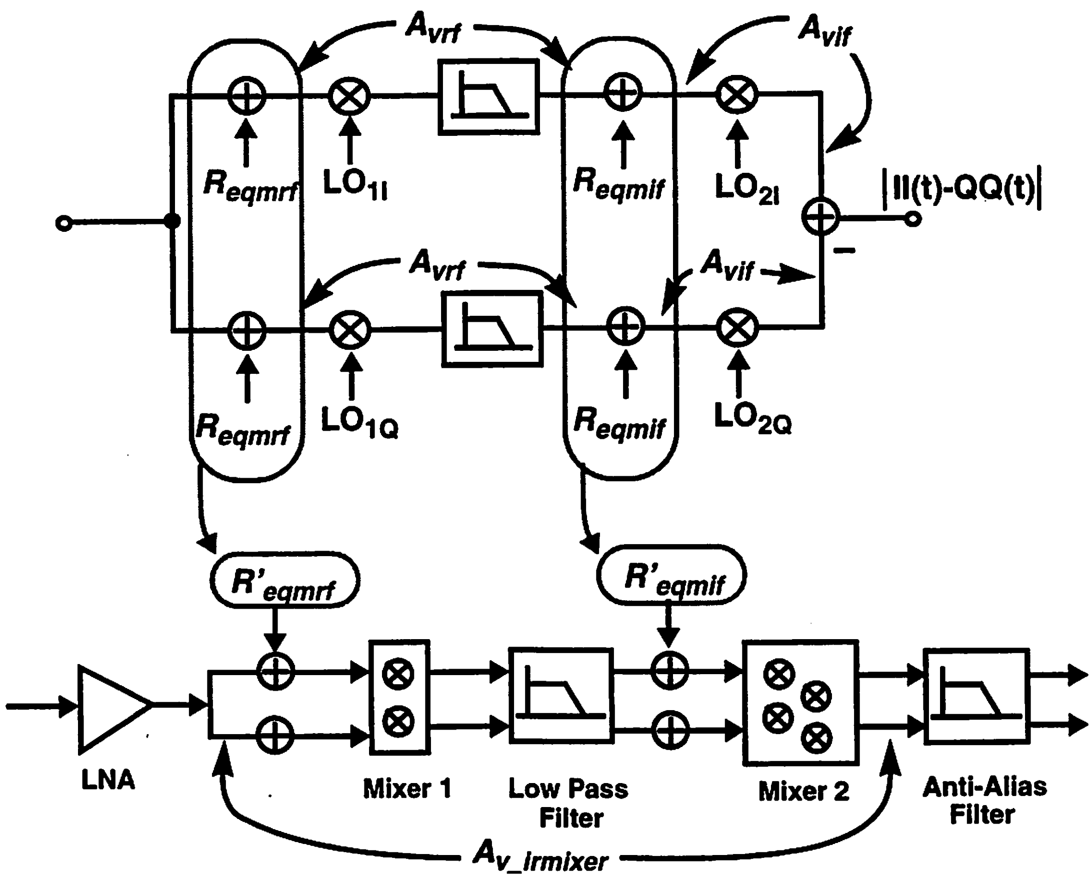


FIGURE 26. Model used to evaluate the noise and single gain in a single channel.

One way to understand the relationship between the signal gain and the effective noise produced at baseband by the mixers can be understood by applying a test signal at the image rejection mixer input and finding the transfer function of the signal through both paths and to the output at baseband. Likewise, the noise transfer function from each mixer can be referred to the output of one channel at baseband. First, a desired signal of mean squared voltage power Sg is applied to the input of the mixer. Before the signal in the two signal paths are summed to produce one channel, the desired signal power is,

$$S^{II}{}_D = (A_{vrf} \cdot A_{vif})^2 \cdot S_D \tag{EQ 51}$$

After the summation of the channels, there is an effective gain of 2x in the signal amplitude and a 4x increase in the signal power.

$$S^I{}_D = (2 \cdot A_{vrf} \cdot A_{vif})^2 \cdot S_D \tag{EQ 52}$$

Or,

$$S^I{}_D = 4 \cdot (A_{vrf} \cdot A_{vif})^2 \cdot S_D \tag{EQ 53}$$

Using a similar approach, the noise produced by the mixers, at the output of one of the two channels is,

$$N^{II} = ((A_{vrf} \cdot A_{vif})^2 \cdot 4kTR_{eqmrf} + A_{vif}{}^2 \cdot 4kTR_{eqmif}) \cdot B \tag{EQ 54}$$

where B is the signal bandwidth. Because the noise between the two signal paths is uncorrelated, the power of the noise adds when the two channels are added together. The total noise at the output of the mixer as a result of the noise inside the mixer is,

$$N^I = 2 \cdot [((A_{vrf} \cdot A_{vif})^2 \cdot 4kTR_{eqmrf} + A_{vif}{}^2 \cdot 4kTR_{eqmif}) \cdot B] \tag{EQ 55}$$

Similar to the SNR argument for a differential amplifier, the desired signal before summation is correlated between the two channels. Therefore, the amplitudes add and the power of the signal is increased by 4x after the summation. However, the noise power only increases by 2x passing through the summation circuit. Therefore, there is a net 3dB increase in the signal to noise ratio before and after the summation. The SNR at the output of the I (or the Q channel) channel can be expressed as,

$$SNR^I = \frac{4 \cdot (A_{vrf} \cdot A_{vif})^2 \cdot S_D}{2 \cdot [((A_{vrf} \cdot A_{vif})^2 \cdot 4kTR_{eqmrf} + A_{vif}{}^2 \cdot 4kTR_{eqmif}) \cdot B]} \tag{EQ 56}$$

or,

$$SNR^I = \frac{2 \cdot (A_{vrf} \cdot A_{vif})^2 \cdot S_D}{[((A_{vrf} \cdot A_{vif})^2 \cdot 4kTR_{eqmrf} + A_{vif}^2 \cdot 4kTR_{eqmif}) \cdot B]} \quad \text{(EQ 57)}$$

From equation 57 it is clear that the SNR or CNR of the desired signal increases by 3dB when the signal passes from before to after the summation of the signal paths at baseband. The question now arises of how to treat the noise produced by a single mixer to the noise of the entire radio channel. To use a spreadsheet such as excel to estimate the noise figure performance of the receiver it is desire to have both the signal and noise passing through the receiver see an equivalent gain. Therefore, one method of translating the equivalent input noise of a single mixer to that of the entire receiver is to simply use an equivalent noise resistance associated with each mixer stage with that of a noise resistance half the value of a single stand alone mixer. Therefore, in the receiver R_{eqmrf} and R_{eqmif} now become, $R'_{eqmrf} = R_{eqmrf}/2$ and $R'_{eqmif} = R_{eqmif}/2$ where, R'_{eqmrf} and R_{eqmif} are the equivalent noise resistances in the receive signal path; this is shown in figure 26. Figure 27 is a more detailed model showing the equivalent noise model for the entire receiver chain. It is easy to verify that this noise model for the image-rejection mixer actually works by finding the signal to noise ratio at the output of the model shown in figure 27.

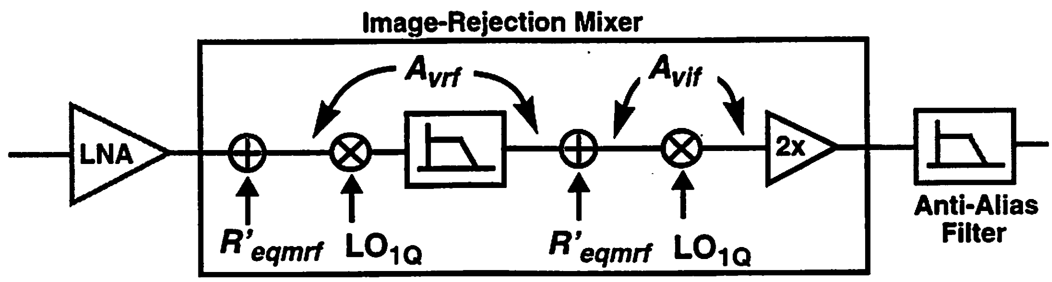


FIGURE 27. Model used to refer the noise of the individual mixers to the overall noise contribution in the receiver.

Assuming there is a signal of power S_D at the input of the mixer, now calculate the signal to noise ratio at the baseband output. The signal power at output of the I baseband channel is,

$$S_D^I = (2 \cdot A_{vrf} \cdot A_{vif})^2 \cdot S_D \quad \text{(EQ 58)}$$

while the noise power at the mixer output is,

$$N = [(2 \cdot A_{vrf} \cdot A_{vif})^2 \cdot 4kTR'_{eqmrf} + (2 \cdot A_{vif})^2 \cdot 4kTR'_{eqmif}] \cdot B \quad \text{(EQ 59)}$$

and the SNR at the baseband input is,

$$SNR^I = \frac{(4) \cdot (A_{vrf} \cdot A_{vif})^2 \cdot S_D}{(4) \cdot [(A_{vrf} \cdot A_{vif})^2 \cdot 4kTR'_{eqmrf} + (A_{vif})^2 \cdot 4kTR'_{eqmif}] \cdot B} \quad \text{(EQ 60)}$$

or,

$$SNR^I = \frac{(A_{vrf} \cdot A_{vif})^2 \cdot S_D}{[(A_{vrf} \cdot A_{vif})^2 \cdot 4kTR'_{eqmrf} + (A_{vif})^2 \cdot 4kTR'_{eqmif}] \cdot B} \quad \text{(EQ 61)}$$

replacing R'_{eqmrf} with $R_{eqmrf}/2$ and R'_{eqmif} with $R_{eqmif}/2$ in equation 61 results in,

$$SNR^I = \frac{(A_{vrf} \cdot A_{vif})^2 \cdot S_D}{[(A_{vrf} \cdot A_{vif})^2 \cdot 4kT(R_{eqmrf}/2) + (A_{vif})^2 \cdot 4kT(R_{eqmif}/2)] \cdot B} \quad \text{(EQ 62)}$$

which gives,

$$SNR^I = \frac{2 \cdot (A_{vrf} \cdot A_{vif})^2 \cdot S_D}{[(A_{vrf} \cdot A_{vif})^2 \cdot 4kTR_{eqmrf} + A_{vif}{}^2 \cdot 4kTR_{eqmif}] \cdot B} \quad \text{(EQ 63)}$$

which is identical to equation 57. Therefore, when estimating the noise figure of the integrated portion of the receiver using equation 21, $R_{mixer1} = R_{eqmrf}/2$ and $R_{mixer2} = R_{eqmif}/2$ should be used. Analysis of the CNR at the output of the receiver should also utilize half the equivalent input noise of a single mixer in the I and Q signal paths. Both section 5.2.1 and section 5.2.2 outline the required specifications of the *individual* mixers used in the image-rejection mixer configuration.

5.2.1 RF-to-IF Mixers (LO1)

All mixers should be doubly balanced. The mixer requirements are summarized in table 3. Note the gain is assumed to be the voltage conversion gain from the input to the output of an individual mixer.

The highest input frequency for the RF-to-IF mixers will be 1.9 GHz. The output frequency at IF output will range from 350 MHz to 400 MHz. The first mixer should have variable conversion gain from 0-10dB in 2dB increments.

TABLE 3. Specifications for the RF-to-IF Mixers.

Arch.	R_{eq}	V_{IP3i}	$V_{o\text{-}pmax}$ (output)	A_v	P_{-1db} (input)	DR	Power
WIF	< 900	> 1.0	600 mV	0-10dB	-5dBV	110 dB	min.
Low-IF	< 1500	> 1.0	270 mV	18dB	-	101 dB	min.

5.2.2 IF-to-Baseband Mixers (LO2)

Similar to the RF mixers, the IF mixers should be doubly balanced. The IF mixer requirements are in table 4. The mixer is specified only for WIF for obvious reasons. The IF-to-Baseband mixers will be required to operate on an incoming frequency ranging from 350 MHz to 400 MHz. The desired carrier will be downconverted to baseband (centered around DC). The conversion gain given in table 4 is the voltage gain of the carrier includ-

TABLE 4. Requirements for the IF-to-Baseband Mixers.

Arch.	R_{eq}	V_{IP3i}	$V_{o\text{-}pmax}$	A_v	P_{-1db}	DR	Power
WIF	< 5,000	> 1.6	1.7 V	7dB	-3 dBV	113 dB	min.
Low-IF	N/A	N/A	N/A	N/A	N/A	N/A	min.

ing the effects of frequency translation from the input of one mixer to the output just prior to the summation of the two channels.

5.3 Baseband Circuits

The lack of frontend and IF filtering associated with any highly integrated receiver generally places an unique and challenging set of required performance on the baseband blocks. In the wideband IF system, there is essentially no filtering of the adjacent channel blockers before the desired carrier is translated to baseband. Therefore, the baseband blocks have a very high dynamic range requirement. In addition, the baseband blocks of this receiver must meet the requirements of both the DECT and DCS1800 standard. At present, the frontend will downconvert both the DECT and GSM channel directly to baseband. Similar to the other components in the receiver, DCS1800 is the more challenging spec. to meet with the exception of the signal bandwidth required of the ADC in DECT. In all likelihood, a variable bandwidth anti-alias filter will be required to alternate between the DECT and GSM modes, this concept is shown in figure 28.

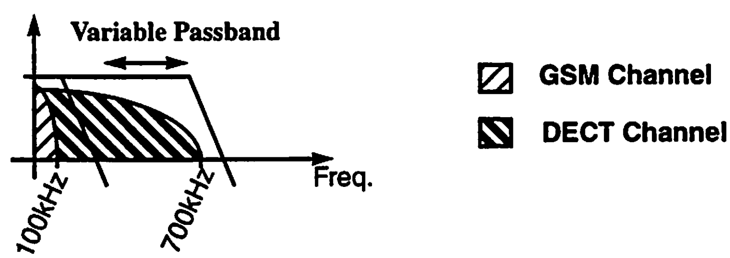


FIGURE 28. Bandwidth channel filtering requirements for dual-standard operation. Both the DECT and GSM channel are frequency translated to baseband.

The specifications for the baseband circuits are separated into two sections for the anti-alias filter and the ADC. The key requirements of the anti-alias filter are the equivalent input noise, the 3rd Order intermodulation intercept point, 2nd order intermodulation intercept point, and the anti-alias requirements for the subsequent sampled-data circuits.

The sampling frequency of the ADC will be 45.6MHz for DECT and 27MHz for GSM modes of operation. These sampling frequencies were used to determine the anti-alias filter requirements.

One of the key problems associated with any system which modulates the carrier to baseband are the effects of second order intermodulation. Both the mechanism of interference and the method to find the require IP2 of the baseband are outlined in section 4.2.3. The required IP2 of the baseband is a function of the magnitude of the blockers which are present. As usually, DCS1800 is the limiting factor in the required IP2 performance of the baseband, specifically the 3MHz blocking condition. Depending on the filtering at the output node of the second mixer the attenuation of the 3MHz blocker can relax the required IP2 of the baseband. From equation 45, it may be seen that every 10dB attenuation of adjacent channel interferers reduce the IP2 requirement by 20dB. Equation 45 was used to generated the required IP2 as a function of the filtering looking into the baseband blocks; from the output of the second mixer looking into the baseband. Table 5 lists the required

TABLE 5. Required IP2 vs. 3dB frequency at the output of the mixer.

3dB Freq.	S_{bl} (3MHz) @ Mixer Output	S_{des} @ Baseband	IP2 (dBV)
300kHz	- 18 dBV	-74 dBV	+ 47 dBV
500kHz	- 14.5 dBV	-74 dBV	+ 54 dBV
1MHz	- 4.5 dBV	-74 dBV	+ 74 dBV
5MHz	+2 dBV	-74 dBV	+ 87 dBV

IP2 performance as function of the 3dB frequency at the second mixer output. This analysis assumes a single pole response on the mixer output node. It becomes plainly obvious without any filtering at the mixer output the IP2 requirements for the baseband are particularly challenging.

5.3.1 Anti-Alias Filter.

The basic structure of the continuous time baseband blocks are shown in figure 29. As stated above, a single pole is used at the mixer output to reduce as much as possible the DC component generated by large adjacent channel blockers passing through the baseband 2nd order non-linearities. A variable gain amplifier is used after the second mixer with a nominal gain of 12dB +/- 3dB to compensate for any gain variation in the receiver frontend. The VGA used in the baseband differs from more traditional variable gain stages used in a super-heterodyne receiver systems. Typically, most of the adjacent channel energy is removed from an external IF SAW filter. After filtering the adjacent channel signals, the desired signal is isolated and a significant amount of gain may be added (the variable gain may be as high as 70dB) to reduce the dynamic range requirements of the subsequent blocks. In most integrated receiver systems (wideband IF being one them) both the desired signal and the adjacent channels are modulated to baseband without any filtering. Therefore, because large adjacent channel blockers may be present, only a moderate amount of variable gain may be used. In this implementation, the variable gain is only utilized to ensure that the largest signal present, whether that be a blocker or desired signal is at approximate the full scale of the ADC. Or in other words, the variable gain amplifier is used to compensate for any gain variation in the frontend, thus reducing the dynamic range of the ADC by approximately 1bit as compared to alternate implementation given in [14].

The requirements for the continuous-time filter are determined by the required antialiasing assuming a 14bit ADC is used for the Zero-IF wideband IF receiver. The antialias filter will be preceded by a variable gain stage which adjusts the amplitude of the largest signal present such that it is aligned in amplitude to the ADC full scale (700mV zero-peak differential voltage). In addition, the baseband front-end filter will partially attenuate all of the DCS1800 blockers at and above 3MHz. The composite configuration of the baseband is as shown in figure 29.

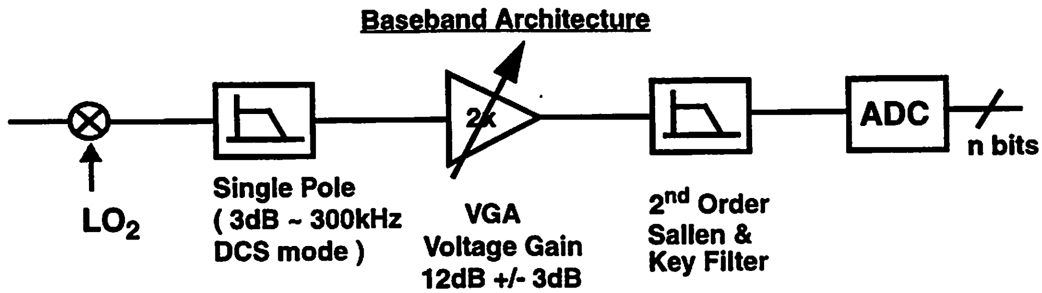


FIGURE 29. Baseband filter and ADC configuration.

The required performance of the anti-alias filter for both the low-IF and wideband IF approach is given in table 6. This spec. assumes that 25dB attenuation of the first out-of-band blockers from the frontend filter making the 3MHz blocker the largest blocker present in the baseband.

TABLE 6. Required Attenuation by the Anti-Alias Filter; assumes Wide-band IF with 16 bit ADC.

Filter Spec.	GSM: Wideband-IF	GSM: Low-IF	DECT
Required Passband	100kHz	100kHz-300kHz	700kHz
Anti-Alias Requirements	> 90dB @ 44.7MHz	> 90dB @ 44.8MHz	> 70dB @ 44.1MHz
Atten. of 600kHz blocker	-3 dB	N/A	N/A
Atten. of 1.6MHz blocker	-25 dB	> 10 dB @ 1.2MHz	N/A
Atten. of 1.78MHz blocker	~25 dB	N/A	0 dB
Atten. of 3.0 MHz blocker	> 41dB @ 3.0MHz	> 22dB @ 2.6MHz	N/A
Atten. of 3.56 MHz blocker	N/A	N/A	> 6 dB@ 3.56 MHz
Atten. of 5.34MHz blocker	N/A	N/A	> 12 dB @ 5.34 MHz

Table 7 lists the required noise, gain, maximum signal level and distortion performance required of the filter.

TABLE 7. Specifications for the Anti-Alias Filter.

Arch.	R_{eq}	V_{IP3i}	$V_{o\text{-}pmax}$ (Output)	A_v	P_{-1db}	DR (Input)	Power
WIF	< 20,000 Ω	> 7 V	600 mV	12dB+/-3dB	-5 dBV	92 dB	min.
Low-IF	130,000 Ω	> 6 V	535 mV	6 dB		88 dB	min.

5.3.2 Analog-to-Digital Converter (ADC)

The dynamic range requirements of the ADC are set on the high-end by the maximum signal level presented at the ADC input and on the low-end by the allowable input noise contribution from the ADC to the overall receiver noise figure. To obtain a receiver noise figure of less than 9dB, it was estimated that the ADC could have no more than

190kΩs of equivalent input noise resistance. A breakdown of the receiver noise contributors can be found in section 6.0.

A summary of the required ADC performance is given in table 8 and table 9. Note: that the required ADC signal bandwidth varies from 100kHz to 700kHz between the GSM and DECT modes of operation respectively.

TABLE 8. Specifications for the baseband ADC.

Arch.	R_{eq}	V_{IP3i}	$V_{o\text{-}pmax}$ (Output)	A_v	P_{-1db}	DR (Input)	Power
WIF	< 190,000	> 22 V	680 mV	0 dB	0 dBV	82 dB (GSM mode)	min.
Low-IF	–	-	-	-	-	-	-

Figure 30 and figure 31 illustrate all of the potential signal levels which include the desired signal, adjacent channel blockers as well as out-of-band blockers as they pass through the receiver for both GSM and DECT. The signal levels are given for the case of wide-band IF using the component voltage gain given in section 5.1 through section 5.4. The required dynamic range of the ADC for GSM is shown in figure 30 as the difference between the largest signal at the ADC input minus the input referred noise level of the ADC. From figure 31 it becomes quickly obvious that the signal levels for DECT fall within the range of possible GSM signals. Therefore, if the ADC can meet the dynamic range requirements for GSM, then it should also be able to meet the requirements of DECT with the exception that the signal bandwidth increases from 100kHz to 700kHz.

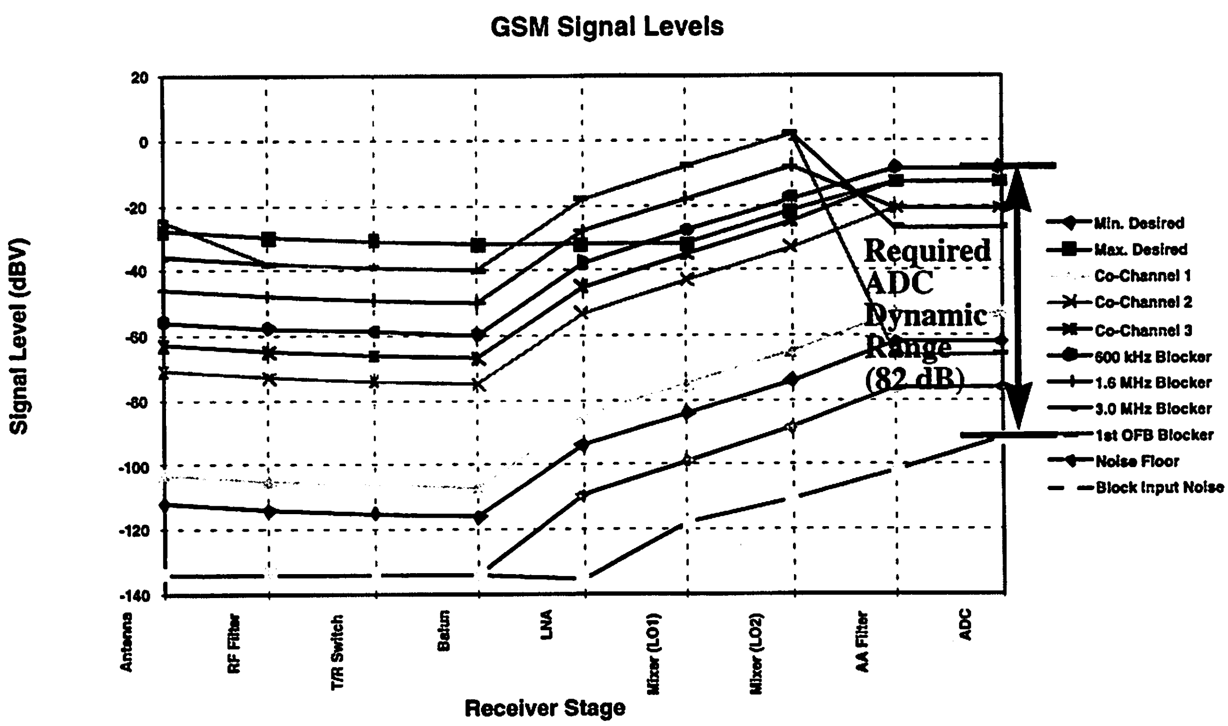


FIGURE 30. All possible GSM signal levels plotted at the output voltage of each receiver stage.

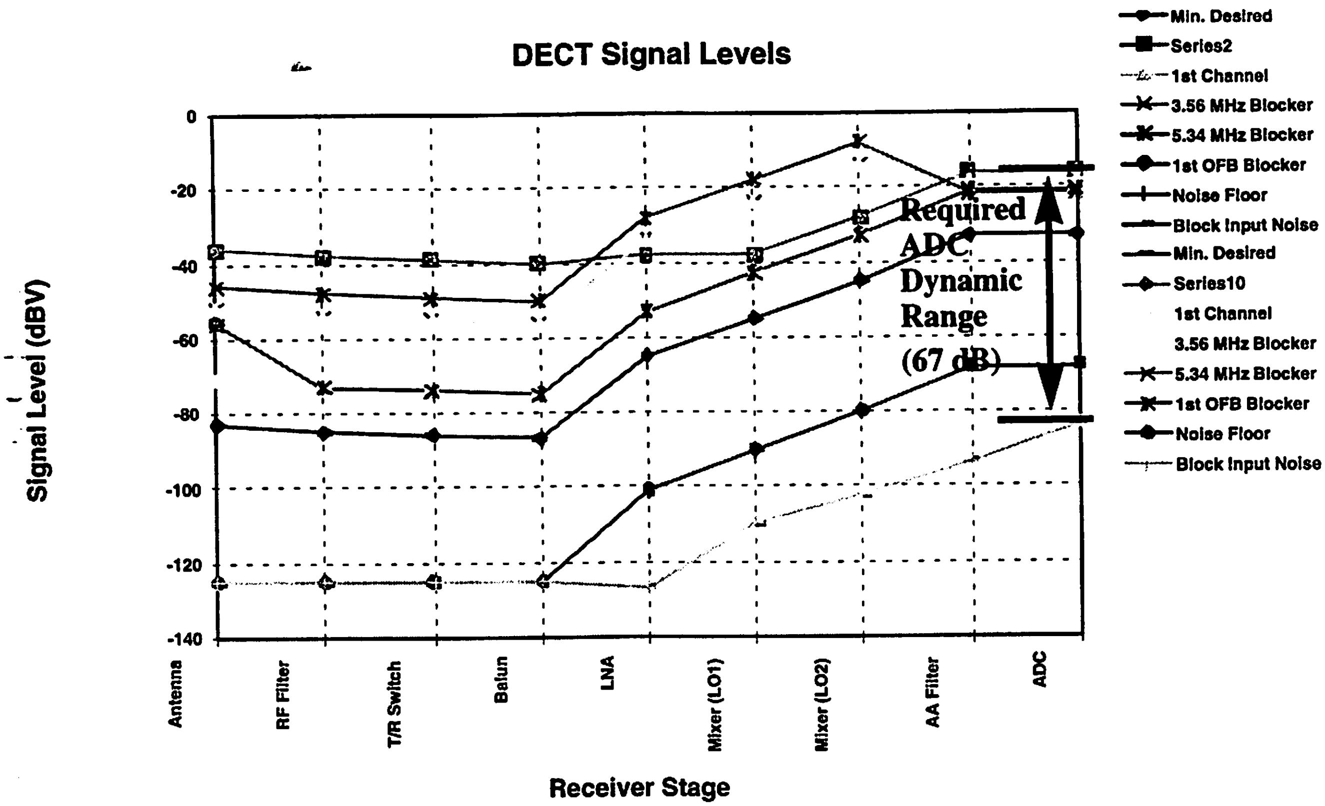


FIGURE 31. All possible DECT signal levels in dBV plotted at the output of each receiver stage.

Table 9 summarizes the dynamic range as well as the signal bandwidth requirements of the ADC when in both the DECT and DCS1800 modes of operation.

TABLE 9. Required ADC dynamic range between the GSM and DECT modes of operation.

	Signal BW	Dynamic Range (dB)
GSM	100kHz	82 dB
DECT	700kHz	67 dB

5.4 Frequency Synthesizers

By far, the most challenging block to integrate on to a single-chip receiver is the frequency synthesizer section with on-chip VCOs. This becomes even more challenging when the target application has both stringent phase noise requirements as is the case with all flavors of GSM. In addition, developing a frequency plan which implements the correct local oscillator frequency to implement multiple RF standards while striving for maximum reuse of the synthesizer hardware among the different target standards becomes particularly difficult.

5.4.1 Purpose of the Frequency Plan

A good frequency plan is crucial to achieving all the specifications of different standards with minimal amount of hardware and power consumption. The frequency plan determines how the frequency translation of the carrier is performed in both the receive and transmit paths. Therefore, the frequency plan determines the amount of hardware and power it takes to generate the reference frequency with the frequency synthesizer and has a significant impact on the overall synthesizer performance, namely, phase noise, spurious tones and the required tuning range.

The relatively narrow channel bandwidth coupled with the large 3MHz blocker of GSM defines the required synthesizer performance for a dual-standard receiver. Using the approach that was described in section 4.3, it was determined that the phase noise performance required of both LO1 and LO2 to meet the BER conditions with blockers present are as shown in table 10 for all flavors of GSM. It is probably worth noting that the 3MHz blocking test is relaxed in both DCS 1800 and PCS 1900 as compared to GSM or E-GSM. These phase noise requirements should be suitable for both the Wideband IF and the Low-IF receiver architectures. The phase noise requirements of DECT are significantly easier to meet.

TABLE 10. Required Phase Noise Performance of the Local Oscillators.

Offset from Carrier	Phase Noise : GSM 900	Phase Noise : DCS 1800	Phase Noise : PCS 1900
600 kHz	-128 (dBc/Hz)	-122 (dBc/Hz)	-125 (dBc/Hz)
800 kHz	-138 (dBc/Hz)	-132 (dBc/Hz)	-135 (dBc/Hz)
3 Mhz	-143 (dBc/Hz)	-137 (dBc/Hz)	-140 (dBc/Hz)

Both synthesizers must drive the mixer with greater than a 700mV zero-to-peak signal differentially. The required common-mode voltage for the mixer input has yet to be determined.

5.4.2 Procedure for developing the Synthesizer Frequency Plan

With all the available bands and standards for cellular and cordless applications, the first question is how many and which standards will be implemented to demonstrate the multi-standard operation while providing the feasibility of integrating the frequency synthesizer with a fully integrated transmit and receive path. At present, for practical reasons related to completing the project on time, it was decided to select one cellular and one cordless standard to demonstrate multi-mode operation. Originally, all flavors of the European and North American GSM were considered in addition to the Digitally Enhanced Cordless Telephone (DECT) standard. After careful consideration and several iterations of the frequency plan both DCS1800 and DECT were selected as the two standards to implement.

With a knowledge of the RF standards to implement, the next question to answer is how many RF paths will be required in the receiver? Fortunately, because the two standards under consideration are very close in frequency, in all likelihood only one RF path (LNA and IR mixer combination) will be required. This would be in contrast to selecting two standards with carrier frequencies spaced far apart. This would at a minimum require an additional LNA and PA.

With a knowledge of the number of required receive and transmit signal paths, the next design choice becomes the crystal oscillator. The ultimate objective is to develop a frequency plan for all standards while using only one external crystal reference oscillator. Additional desired properties which influence the choice of the crystal are the size and the phase noise performance of the just the external crystal oscillator. Currently, the available crystals on the market below 200MHz typically have a phase noise level below -145dbc/Hz at 50kHz offset frequency. With a low phase noise option added to the crystal a phase noise performance of -160dbc/Hz at 50kHz offset frequency may be obtained. Therefore, for this project a single crystal reference was used to implement the frequency synthesizers of both DCS1800 and DECT. In addition, a divided form of the crystal is used to generate the clocks used in the baseband blocks.

The last question to address before synthesizing a frequency plan is whether or not it is possible to achieve the required phase noise performance with a selected crystal using either the wideband PLL approach or a more conventional narrow band PLL. Also, a thorough understanding of the harmonics and intermodulation products of LO1 and LO2 located outside of the band of interest are essential.

With above the questions in mind, a frequency plan may be developed by iteratively going through the questions outlined above.

If DECT is included on this chip with any one of the GSM standards (EGSM, DCS1800, PCS1900), we need to have a reference frequency that is a multiple of 1.728MHz(channel spacing of DECT) and a multiple of 0.2MHz(channel spacing of GSM). The minimum value of such frequencies is 43.2MHz.

If the 43.2MHz reference is used, the frequency step of LO1 is 43.2M and the minimum IF range LO2 should be able to generate is 43.2M. To improve the image-rejection from the frontend filter, the IF should be at least 200MHz with an approximate 1.9GHz carrier. This implies that the divider ratio (N) to implement the LO1 is about 36 (1.6GHz/43.2MHz). The phase noise of the crystal and phase detector and the divider is amplified by N, e.g., 31dB. With a 31dB noise enhancement from the divider it is virtually impossible to meet the phase noise requirement for cellular application using a wideband PLL with an integrated VCO. Therefore, recently the crystal reference frequency was moved to 86.4MHz. With an 86.4MHz crystal, the divider ratio N is significantly reduced from 36 to 16 with a 400MHz IF. If the divider ratio is reduced to 16 the noise amplification of the crystal, phase detector, and dividers are reduced to 24 dB making it somewhat more possible to implement a wide band (8MHz loop BW) PLL for the first local oscillator (LO1) using an external low phase noise crystal. To first order, the phase noise from the first VCO will be shaped by the loop within the loop filter bandwidth. The total phase noise seen at the output of the local oscillator at 3MHz offset from the carrier is approximately -140dBc/Hz.

The narrow band PLL approach is used for LO2 to suppress the spurious tones generated by the loop. Therefore, with a narrow loop bandwidth, the phase noise from crystal and phase detector and the divider will also be suppressed by the loop filter at the output of the LO2 PLL. The overall phase noise profile of LO2 is dominated outside the loop bandwidth by the VCO. However, the phase noise requirements of the VCO for LO2 is relaxed by 12 dB because the VCO output is divided by 4 to get the IF frequency. The required tuning range of LO2 can be approximated as the crystal reference frequency divided by the IF frequency. For an 80MHz crystal and 400MHz IF, the tuning range is about 20%.

LO2 may be duplicated on the same chip to accommodate the required tuning of the second LO; this is yet to be determined.

The approach described above was used to synthesize four possible frequency plans which are shown in table 11.

Plan1 implements PCS1900 with 2 to 3 PLLs, depending on the tuning range of the LO2.

Plan2 implements DCS1800 with 2 to 3 PLLs. The IF frequency is the highest because DCS1800 has the widest signal BW(75MHz for receive band and 75MHz for transmit band). The tuning range is smaller with a higher IF frequency.

Plan 3 implements DCS1800 and DECT1900 with 4 PLLs. It does demonstrate the multistandard concept with a little bit more hardware while reducing the number of RF paths to 1.

Plan 4 implements DCS1800 and EGSM900. This plan uses the same LO1 to do either high side or low side mixing of the RF band. But it requires two RF paths.

In summary, plan3 was selected as the frequency plan to implement.

TABLE 11. comparison of frequency plan

comp	plan1	plan2	plan3	plan4
Standards	PCS1900	DCS1800	DCS1800/DECT	DCS1800/EGSM
Crystal	91.2M	80M	86.4M	84M
LO1	1459.2M/ 1550.4M	1280M/ 1360M	1382.4M/ 1468.8M	1344M/ 1428M
IF range (LO2)	390.8-450.8M(Xmit) 379.6-439.6M(Rec) (1.186X)	430-505M(Xmit) 445-520M(Rec) (1.21X)	412-429M(DECT) 327.6-367M(1.12X) 367.2-411.2M(1.12X)	366-412M(1.125X) 412-464M(1.125X)
PLL	2 IF range reduced to 1.09X if 3 PLL's	2 IF range reduced to 1.1X if 3 PLL's	4	3
RF paths	1	1	1	2

6.0 Predicted Receiver Performance Wideband-IF vs. Low-IF

Shown below in table 12 is a summary of the predicted receiver performance based on the LNA, mixer, and baseband specifications given in section 5.1 through section 5.4.

All of the raw excel spreadsheets used to produce the results given in this section can be found in the appendix.

TABLE 12.

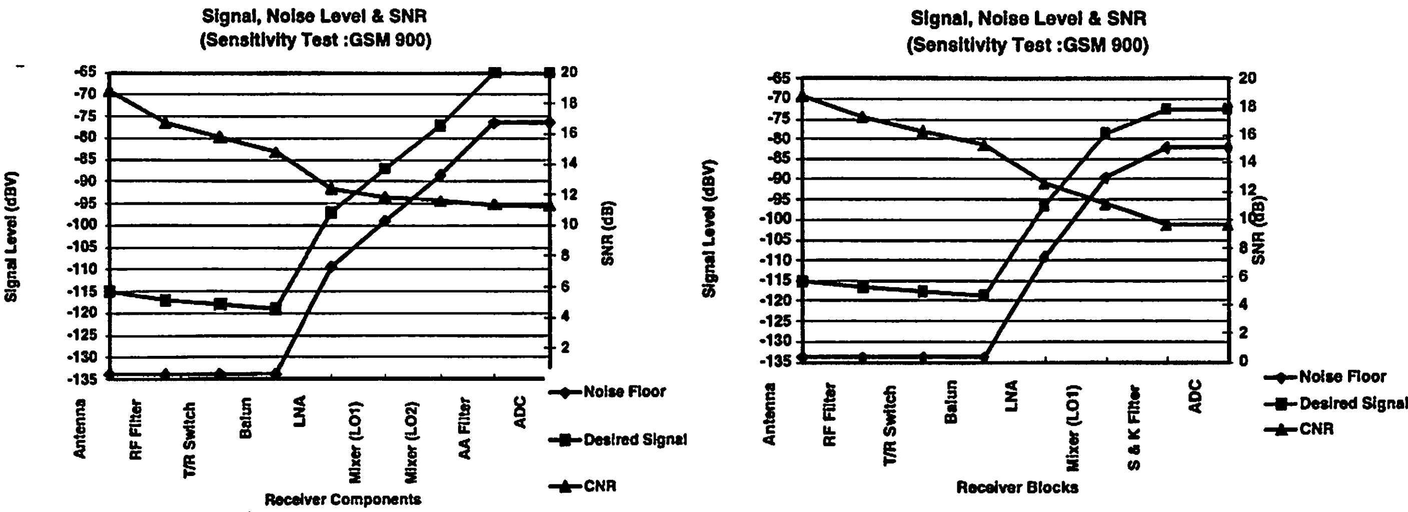
	WBIF	Low-IF / Single Conversion	Required GSM	Required DECT	Required GPS
Receiver Noise Figure	7.5 dB	9.15 dB	9.78 dB	20.29 dB	
Receiver Sensitivity	-104.3 dBm	-102.6 dBm	-102dBm	-83dBm	
Input IP3	-12 dBm	-8.6 dBm	-18 dBm	-26dBm	
P_{-1dB}	-8 dBV	~ -18.6 dBm	-8dBV		
Image-Rejection	80 dB				

Table 13 outlines some of the key results obtained from the excel worksheet used to predict the performance of the wideband IF receiver. The CNR, C/I, and C/(I+N) ratios throughout the receiver are tabulated for the sensitivity, intermodulation and blocking test in GSM. Note, the carrier-to-interference ratios shown in table 13 are at the output of each receiver block (ie. the number given in the ADC column is the carrier-to-interference ratio at the output of the receiver).

	Antenna	RF Filter	T/R Switch	Balun	LNA	Mixer (LO1)	Mixer (LO2)	Anti-Alias	ADC
Gain-Power	0	-2	-1	-1					
Voltage Gain (dB)	0	-2	-1	-1	22	10	10	12	0
Voltage Vip3 (block input)	100.00	100.00	100.00	100.00	0.1000	1.0	1.6	7	22.00
CNR (Sensitivity Test) (dB)	18.7794001	16.7794001	15.77940009	14.779	12.395	11.861421	11.59117154	11.386226	11.26786
Maximum Signal (dBV) Output	-13	-30	-31	-32	-18	-8	2	-9	-9
Location of Largest Blocker	100MHz OFB	fomax	fomax	fomax	3.0 MHz	3.0 MHz	3.0 MHz	600 kHz	600 kHz
Dynamic Range Required	89.1387196	87.1387196	86.13871964	85.139	98.681	91.7872385	94.34511048	77.309064	67.53183
C/I (Intermod Test Condition) (dB)	21.790944	19.790944	18.79094396	17.775	15.386	14.8059615	14.02517239	13.421647	12.51751
C/(I+N)dB (GSM 3MHz Blocking Test)	21.790944	19.790944	18.79094396	17.791	15.395	11.3990876	9.417042968	9.3537514	9.316251
Noise Figure	7.52308371	5.52308371	4.523083708	3.5231					
Input IP3 (dBm)	-11.79	-11.79	-15.79	-17.79					

TABLE 13. Key receiver performance under various test conditions. Wideband IF w/ Double Conversion

Both receiver architectures were designed to have an approximate noise figure of 9 dB with a reasonable intermodulation performance. The required dynamic range given in table 13 is set by the maximum signal minus the required noise floor to meet the sensitivity requirements. Shown in figure 32 is the desired signal level with a -102 (dBm) carrier

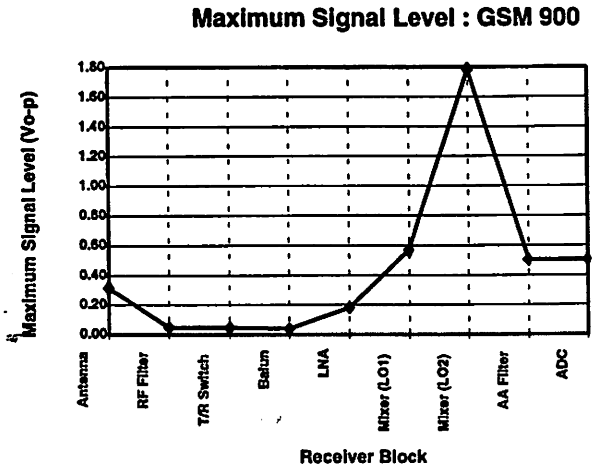


(a) (b)

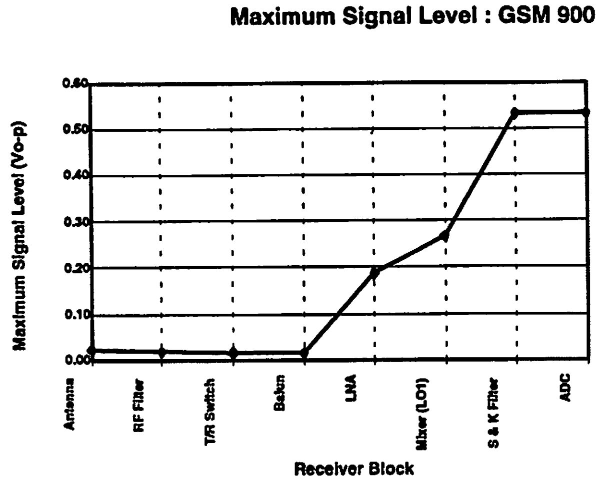
FIGURE 32. Noise floor, minimum signal level and carrier-to-noise ratio CNR in dBV for (a) Wideband IF (b) Low-IF with signal conversion.

applied to the receiver input (GSM 900 sensitivity test). The plots are given for both wideband IF and the Low IF systems. The desired carrier, noise floor, and carrier-to-noise ratios (CNR) are all given at each stage in the receiver chain. Both the signal level, and the noise floor rise moving from the antenna to the back-end of the receiver. However, the difference between the desired signal and the noise floor decrease when moving from the frontend to baseband. Obviously, at the input of the receiver the CNR ratio is the highest, however, as the signal moves down the receiver chain picking up noise, the carrier-to-noise ratio drops. The difference between the CNR ratio at the input and the output of the receiver is the receiver noise figure. From both table 13 and figure 32, it can found that the CNR at the output of the receiver is above the 9dB requirement under the GSM 900 sensitivity test.

With respect to noise and the required noise figure there exists a trade-off between Wideband IF and Low-IF w/ single conversion architectures. Practically speaking the voltage gain of the LNA is limited for stability reasons to about 22 dB while the noise performance of the baseband filter and ADC have a practical lower bound of $20uV$. Therefore, to meet a target noise figure for either receiver architecture, considerable voltage conversion gain must be obtained in the mixer stage to reduce the noise contribution from the baseband. A constraint on the voltage gain through the mixer section of the receiver is the 3MHz blocker in GSM, excess mixer gain will create prohibitively large signals at baseband. For the initial receiver design, the maximum signals propagating down the receiver chain are given in figure 33, and table 13.



(a)



(b)

FIGURE 33. Maximum signal levels throughout the receiver chain due to the blockers, (a) Wideband-IF (b) Low-IF with single conversion.

All signal levels both the blockers and the minimum and maximum desired carrier are shown in figure 34 for the wideband IF receiver with GSM 900 test signal conditions. The first out-of-band blockers are filtered by the RF ceramic filter. As the signal further propagates down the receiver chain, only gain is added to each of the signals without any filtering. At baseband, further attenuation is performed for both the out-of-band and inband blocks at the output of the mixer. Additional anti-aliasing and blocker attenuation is achieved through the Sallen & Key filter.

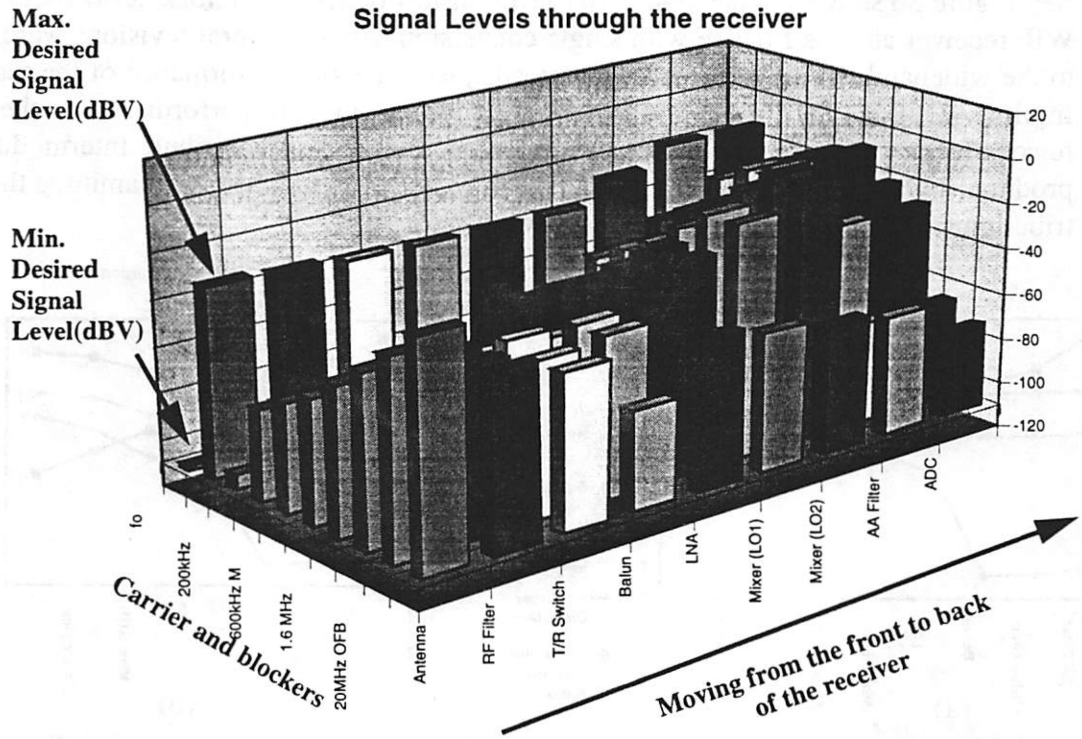


FIGURE 34. All the GSM signal levels in the receiver chain including both inband and out-of-band blockers.

Shown in figure 35 are simple pie charts which illustrate the partitioning of the noise contributors in the receiver. Figure 35 was created by referring all of the rms noise sources to the input of the receiver. The percentage contribution to the overall degradation of each component to the receiver CNR was then plotted.

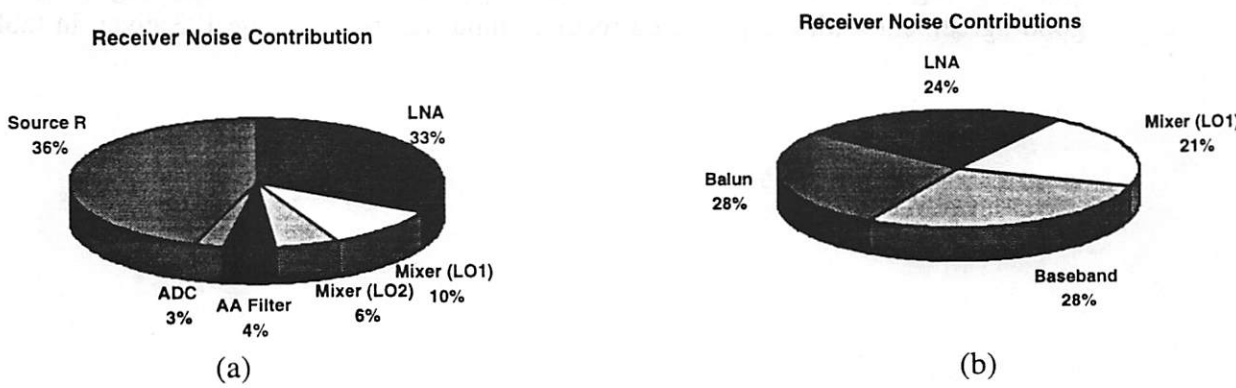


FIGURE 35. Breakdown in the input referred noise contributors from different receiver blocks in (a) Wideband-IF and (b) Low-IF with single conversion.

Note that in the case of Low-IF with single conversion the baseband contributes significantly more noise to the overall receiver than the WIF even though the first mixer gain has been increased from 10dB to 18dB.

The GSM intermodulation test was outlined in section 4.2. This distortion test is performed by applying two -49 dBm adjacent channel signals with a -99 dBm desired carrier. Figure 36 shows the desired carrier level, distortion level and noise level for both the WIF receiver and the Low-IF with single conversion. Again, several revisions were made to the wideband IF receiver to obtain a an adequate linearity performance of the receiver. Intuitively, the Low-IF receiver should have superior linearity performance as there are fewer blocks in the signal (one less mixer stage) path which contribute intermodulation products which degrade the C/I ratio. This can be better understood by examining the contribution of each stage to the overall receiver's linearity.

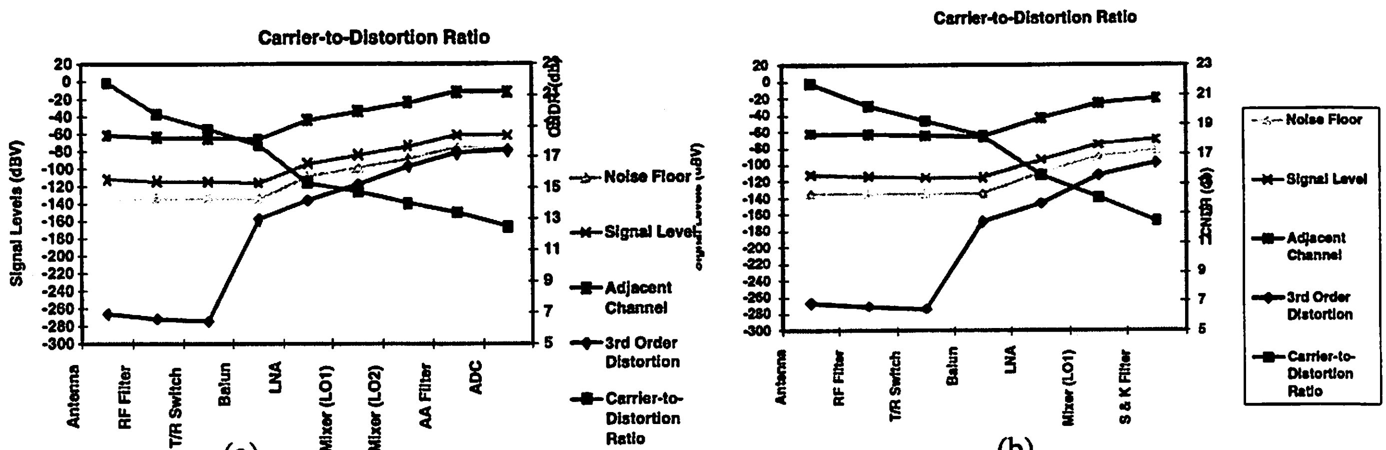


FIGURE 36. The desired signal level, noise floor, 3rd Order distortion components, adjacent channel signal strength all dBV are plotted with respect to the left axis for each stage down the receiver chain. The right axis shows the carrier-to-(noise + distortion) ratio for both (a) Wideband IF and (b) Low-IF with single conversion.

A predicted breakdown of all the interference components (both noise and 3rd order intermodulation) at the output of the receiver under the GSM 900 adjacent channel test is shown in figure 37. Notice that the percentage contribution of distortion for the WIF architecture is higher than in the Low-IF with single conversion approach. Figure 37 shows good agreement with the predicted receiver input referred voltage IP3 given in table 12.

We can see that the distortion contribution is significantly less for the Low-IF approach because the IP3 is much higher than that predicted for WIF.

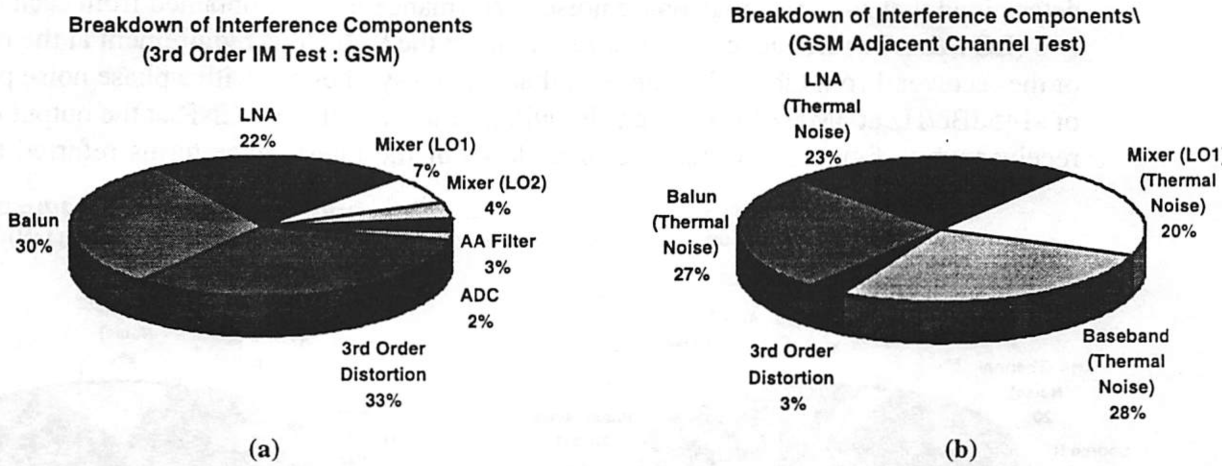


FIGURE 37. Contribution to the overall noise floor at the output of the receiver including the 3rd order distortion components created by two (-49 dBm) intermodulating tones (a) Wideband IF (b) Low-IF with single conversion.

Figure 38 is a breakdown of each receive's block contribution to overall input referred voltage IP3. The individual terms of equation 32 are plotted in figure 38 as a percent contribution to $1/V^2_{IP3}$. Obviously, the component that takes a larger portion of this pie dominates the overall receiver linearity. As expected, the distortion performance of the second mixer limits the IP3 for the Wideband IF architecture. For the Low-IF receiver, the limitation on the receiver's linearity is related to the maximum input swing of the baseband filter. An ideal receiver design would show an equal contribution for all the receiver components to the overall IP3 and in this particular implementation the IP3 number given in section 5.0 for the LNA and Mixer (LO1) can probably be relaxed or traded-off to improve the overall power performance in either of the proposed architectures.

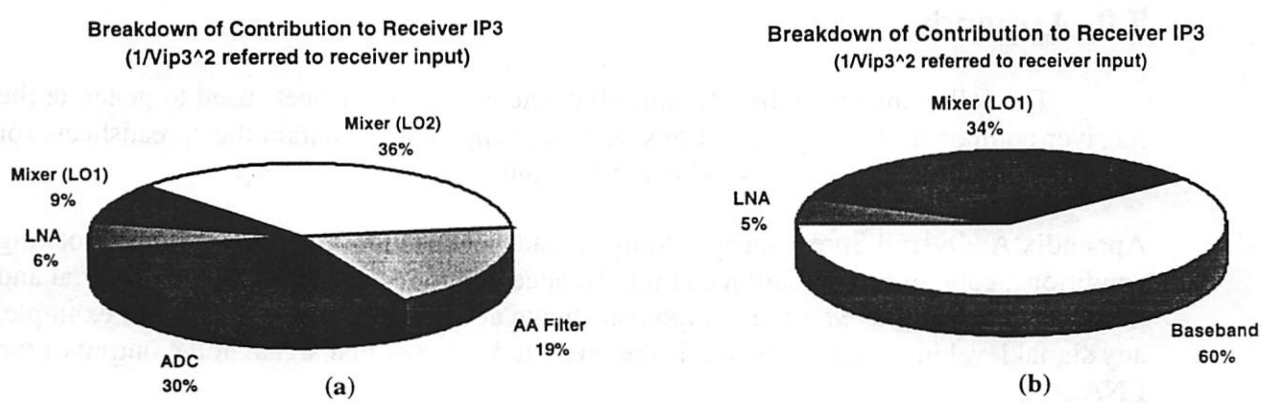
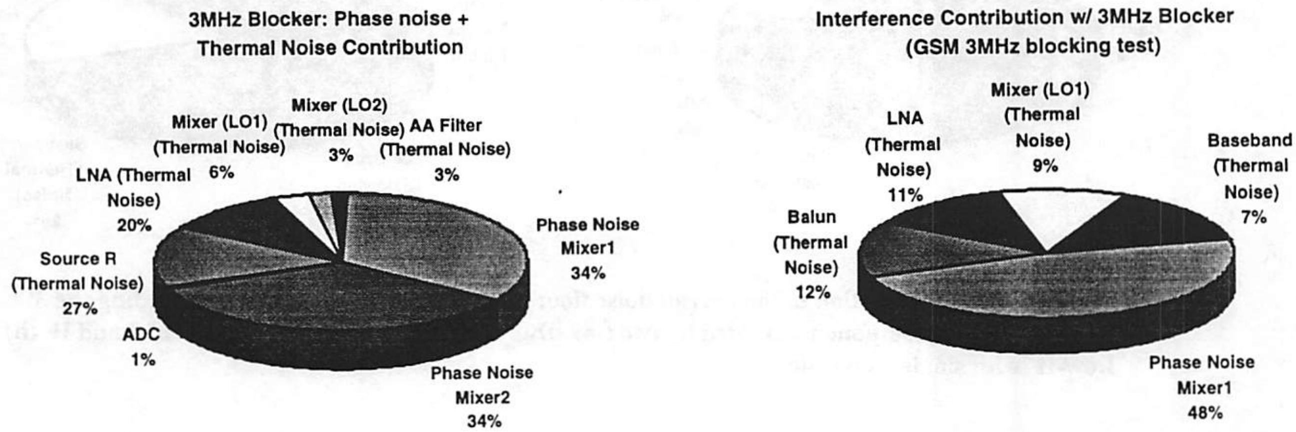


FIGURE 38. Individual contribution of each block to the overall receivers input referred IP3. The breakdown is with respect to each receiver block's V_{IP3} referred to the receiver input, (a) Wideband IF (b) Low-IF w/ single conversion

The phase noise performance of the overall receiver was estimated using the method described in section 4.3. Again, under the worst case blocking condition, the 3 MHz blocker at -23dBm with a -99 dBm desired carrier (GSM 900), the output carrier-to-noise ratio was estimated. With the predicted noise performance of the receiver chain it was determined that a -145dBc/Hz phase noise performance must be obtained from each of the two (LO1 & LO2) frequency synthesizers to meet the 9dB CNR requirement at the output of the receiver. From table 13 it can seen that using a synthesizer with a phase noise profile of -145dBc/Hz at 3MHz from the carrier will produce a sufficient CNR at the output of the receiver while figure 39 reveals the breakdown of the interference terms referred to the



(a) (b)

FIGURE 39. Distribution of interference referred to the output of the receiver under the 3MHz blocking condition of GSM. The reciprocal mixing of the 3 MHz blocker and the phase noise is shown as phase noise mixer1 & 2. (a) Wideband IF (b) Low-IF single conversion.

output of both of the proposed receiver architectures. Note that both of the proposed architectures, WIF and Low-IF, have comparable phase noise contribution which intuitive makes sense since both local oscillators reciprocal mix the phase noise with the blocker in equal amounts.

7.0 Appendix

The following appendix contains all of the excel spreadsheets used to generate the receiver component level specifications. All three appendices contain the spreadsheets for the wideband IF w/ double conversion receiver only.

Appendix A: Overall Spreadsheet.... Main spreadsheet used to find most signals, blocking conditions, gain, intermodulation and interference due to reciprocal mixing. All signal and noise levels are at the *output* of component shown at the top of each column. For example, any signal level in the LNA column is the predicted value of that signal at the output of the LNA.

Appendix B:Blocking Spreadsheet... Spreadsheet used to find the all possible signal levels propagating through the receiver. Contains signal levels for both GSM and DECT. Again, each column represents a signal level at the output of the component represented in that column.

Appendix C: Image Rejection Mixer... Signal, gain and noise levels as they pass through the image-rejection mixer.

7.1 Appendix A: Main receiver spreadsheet.

	Antenna	RF Filter	T/R Switch	Balun	LNA	Mixer (LO1)	Mixer (LO2)	AA Filter	ADC
Gain-Power	0	-2	-1	-1					
Gain-Av (dB)	0	-2	-1	-1	22	10	10	12	0
Gain (Av/Av)^2	1	0.630957	0.79432823	0.7943282	158.4893192	10	10	15.848932	1
Gain (Av/Av)	1	0.794328	0.89125094	0.8912509	12.58925412	3.16227766	3.16227766	3.9810717	1
Gain (Total)		-2	-3		19	29	39	51	51
Avail. Noise Power (dBm)	-120.7897	-120.7897	-120.7897	-120.7897					
Avail. Noise Level (dBV) (200kHz)	-133.7897	-133.7897	-133.7897	-133.7897	-109.4057457	-98.87172092	-88.8014715	-76.39853	-76.27816
Avail. Noise Power (watts) (200kHz BW)									
Minimum Signal Level Output (dBm)	-102	-104	-105	-106					
Minimum Signal Level Output (V^2) (dBV)	-115.0103	-117.0103	-118.0103	-119.0103	-97.01029996	-87.01029996	-77.01029996	-65.0103	-65.0103
Minimum Signal Level Output (Vrms^2)	3.15E-12	1.99E-12	1.58E-12	1.256E-12	1.99054E-10	1.99054E-09	1.99054E-08	3.155E-07	3.155E-07
Minimum Signal Level Output (Vrm)	1.78E-06	1.41E-06	1.26E-06	1.12E-06	1.41E-05	4.46E-05	1.41E-04	5.62E-04	5.62E-04
Minimum Signal Level Output (Vo-p)	2.5119E-06	2E-06	1.7783E-06	1.585E-06	1.99526E-05	6.30957E-05	0.000199526	0.0007943	0.0007943
Impedance	50	50	50	50	400	500	4000		
Vrms^2 (V^2/Hz) (Input Source Noise)					8.3E-19				
Req (Noise Equivalent Resistance) Input	12.5	12.5	12.5	12.5	9.2	4.50E+02	2.50E+03	2.00E+04	190000
Req (Noise Equivalent Resistance) Source									
Req (Noise Equivalent Resistance) Source					3439.218228	38892.18228	413892.2122	6876728.1	7066728.1
Vrms^2 (V^2) in 200kHz BW Noise	4.1675E-14	4.17E-14	4.1675E-14	4.168E-14	1.14664E-11	1.29667E-10	1.37992E-09	2.293E-08	2.356E-08
Block Noise Contribution (Receiver Input)				12.5	9.2	2.83930805	1.575525063	1.2619147	0.7564035
Noise Factor	5.65338251	3.567043	2.83340314	2.2506521					
Noise Figure	7.52308371	5.523084	4.52308371	3.5230837					
SNR (dB) (dBV/dBV) (GSM Sensitivity Test)	18.7794001	16.7794	15.7794001	14.7794	12.39544575	11.96142096	11.59117154	11.386226	11.26785
Adjacent Channel Interferer (dBm)	-49	-47	-46	-45					
Adjacent Channel Interferer (dBV)	-62.0103	-60.0103	-59.0103	-58.0103	-36.01029996	-26.01029996	-16.01029996	-4.0103	-4.0103
Adjacent Channel Interferer (Vrms^2)	6.29E-07	9.98E-07	1.26E-06	1.58E-06					
Desired Signal Level (dBm)	-99	-101	-102	-103					
Desired Signal Level (dBV)	-112.0103	-114.0103	-115.0103	-116.0103	-94.01029996	-84.01029996	-74.01029996	-62.0103	-62.0103
Desired Signal Level (Vrms^2)	6.29E-12	3.97E-12	3.15E-12	2.51E-12	3.97164E-10	3.97164E-09	3.97164E-08	6.295E-07	6.295E-07
Undesired Adjacent Channel (dBm)	-49	-51	-52	-53					
Undesired Adjacent Channel (dBV)	-62.0103	-64.0103	-65.0103	-66.0103	-44.01029996	-34.01029996	-24.01029996	-12.0103	-12.0103
Undesired Adjacent Channel (Vrms^2)	6.29E-07	3.97E-07	3.15E-07	2.51E-07	3.97164E-05	0.000397164	0.003971641	0.0629463	0.0629463
Undesired Adjacent Channel (Vrms)	7.93E-04	6.30E-04	5.62E-04	5.01E-04	6.30E-03	1.99E-02	6.30E-02	2.51E-01	2.51E-01
Output referred IP3				2.00	1.26	2.478159971	4.249664127		
3rd Order component at the output (Vrm^2)									
									104.3
Input referred IP3 (dBm)		100.00	100.00	100.00	5	-6			
Input referred Vip3 (per block)(rms)	100.00	100.00	100.00	0.0288551	0.1000	1.0	1.6	7	22
Input referred IP3 (dBV)					-10	0	2.041199827	8.4509804	13.424227
Block Contribution to IP3					100.00	158.49	619.10	323.45	518.98
IP3 referred to the Receiver Input (Vip3)	0.0575734	0.057573	0.03632637	0.0288551					
IP3 referenced to the Receiver Input (dBm)	-11.79	-11.79	-15.79	-17.79					
Output refered IP3(Vip3o)	100.00	79.43	89.13	89.13	1.26	3.16	5.06	27.87	22.00
Equivalent 3rd Order Output referred IP3	100	56.16749	55.4343167	59.29923	1.258924619	2.478159971	4.249664127	14.461789	12.084625
Total 3rd Order Volt. at output (Vrms)	4.99E-14	2.50E-14	1.77E-14	1.25E-08	1.57927E-07	1.29091E-06	1.38594E-05	7.551E-05	0.0001081
Total 3rd Order (Vrms^2)	2.49E-27	6.26E-28	3.14E-28	1.57E-16	2.49E-14	1.67E-12	1.92E-10	5.70E-09	1.17E-08
Total 3rd Order (dBV)	-266.03	-272.03	-275.03	-158.03	-136.03	-117.78	-97.17	-82.44	-79.32
CDR (dB) w/ Adjacent Channel Test	154.0206	158.0206	160.0206	42.0206	42.02058998	33.77174786	23.15478419	20.429481	20.429481
SNDR (dB) w/ Adjacent Channel Test	21.790944	19.79094	18.790944	17.774576	15.38601053	14.8059615	14.02517239	13.421647	12.517511
3 MHz blocker available power (dBm)	-23	-25	-26	-27					
3 MHz blocker (dBV)	-36.0103	-38.0103	-39.0103	-40.0103	-18.01029996	-8.010299957	1.989700043	13.9897	
3 MHz blocker (Vrms^2)	2.51E-04	1.58E-04	1.26E-04	9.98E-05	0.015811388	0.158113883	1.58113883	25.059362	
CPNR (Carrier-to-phase noise ratio) (dB)						14	14		
Phase Noise floor (dBV)						-98.01029996	-88.01029996		
Phase Noise floor (Vrms^2)	0	0	0	0	0	1.58114E-10	1.58114E-09	0	
Phase Noise Contribution receive output						29.92295013	29.92295013		
Phase Noise Response(dBc/Hz) w/ 3MHz						-143	-143		
Thermal noise + Phase noise (Vrms^2)	4.1675E-14	4.17E-14	4.1675E-14	4.168E-14	1.14664E-11	2.8778E-10	4.54219E-09	7.305E-08	7.368E-08
C/I(w/ 3MHz blocker)	21.790944	19.79094	18.790944	17.790944	15.39544575	11.39908764	9.417042968	9.3537514	9.3162513

7.2 Appendix B: Spreadsheet for signal levels both GSM & DECT

	Freq.	Antenna	RF Filter	T/R Switch	Balun	LNA	Mixer (LO1)	Mixer (LO2)	AA Filter	ADC
GSM Blockers(Worst Case GSM)										
Minimum Desired Signal (Sensitivity) (dBV)	fo	-112	-114	-115	-116	-94	-84	-74	-62	-6
Maximum desired Signal (dBV)	fomax	-28	-30	-31	-32	-32	-32	-22	-13	-1
Co-Channel - 1 Channel Away (dBV)	200kHz	-103	-105	-106	-107	-85	-75	-65	-53	-5
Co-Channel - 2 Channels Away (dBV)	400kHz	-71	-73	-74	-75	-53	-43	-33	-21	-2
Co-Channel - 3 Channels Away (dBV)	600kHz M	-63	-65	-66	-67	-45	-35	-25	-13	-1
600 kHz blocker (dBV)	600 kHz	-56	-58	-59	-60	-38	-28	-18	-9	-
1.6 MHz blocker (dBV)	1.6 MHz	-46	-48	-49	-50	-28	-18	-8	-21	-2
3.0 MHz blocker (dBV)	3.0 MHz	-36	-38	-39	-40	-18	-8	2	-27	-2
Out-Of-Band blocker 20 MHz away (dBV)	20MHz OFB	-25	-38	-39	-40	-18	-8	2	-66	-6
Out-Of-Band blocker 100MHz away (dBV)	100MHz OFB	-13	-35	-51	-67	-45	-50	-55	-58	-7
Noise Floor (dBV)		-133.7897	-133.7897	-133.7897	-133.7897	-109.4057467	-98.87172092	-88.6014715	-76.396525	-76.2781
Block Input Referred Noise level (dBV)		-133.7897	-133.7897	-133.7897	-133.7897	-135.1324658	-118.2382189	-110.796091	-101.76004	-91.9828
Largest Blocker present (dBV)		-13	-30	-31	-32	-18	-8	2	-9	-
Largest signal present (Vo-p)		0.31660298	0.04472136	0.039857954	0.035523439	0.178038939	0.56300856	1.780389391	0.5017819	0.501781
Location of Largest Signal		100MHz OFB	fomax	fomax	fomax	3.0 MHz	3.0 MHz	3.0 MHz	600 kHz	600 kHz
Required Dynamic Range GSM		120.7897	103.7897	102.7897	101.7897	117.1324658	110.2382189	112.7960909	92.760044	82.98280
DECT Blockers										
Minimum Desired Signal (Sensitivity) (dBV)	fo	-96	-98	-99	-100	-78	-68	-58	-46	-4
Maximum Desired Signal (dBV)	fo	-36	-38	-39	-40	-38	-38	-28	-16	-1
1 Channel Away (dBV)	1.78 MHz	-71	-73	-74	-75	-53	-43	-33	-21	-2
2 Channels Away (dBV)	3.56 MHz	-52	-54	-55	-56	-34	-24	-14	-18.3	-18.
3 Channels Away (dBV)	5.34 MHz	-46	-48	-49	-50	-28	-18	-8	-22	-2
First Out-of-band blocker 6MHz (dBV)	6 MHz	-56	-73	-74	-75	-53	-43	-33	-21	-2
Noise Floor (dBV)		-125.1387196	-125.1387196	-125.1387196	-125.1387196	-100.9547663	-90.42074052	-80.1504911	-67.945545	-67.8271
Block Input Referred Noise Level (dBV)		-125.1387196	-125.1387196	-125.1387196	-125.1387196	-126.6814854	-109.7872385	-102.34511	-93.309064	-83.5318
Largest Blocker present (dBV)		-36	-38	-39	-40	-28	-18	-8	-16	-1
Largest Blocker present (Vo-p)		0.022413774	0.017803894	0.015867737	0.014142136	0.056300856	0.178038939	0.56300856	0.2241377	0.224137
Location of Largest Signal		fo	fo	fo	fo	5.34 MHz	5.34 MHz	5.34 MHz	fo	fo
Required Dynamic Range		89.13871964	87.13871964	86.13871964	85.13871964	98.68148537	91.78723851	94.34511048	77.309064	67.53182

7.3 Appendix C: Spreadsheet used for the image-rejection mixer.

	LNA	Mixer (LO1)	Mixer (LO2)	Gain Stage	AA Filter
		Individual	Individual w/	(Includes	
			Shift to DC	Recombination)	
Gain-Av (dB)	22	10	-3	13	12
Gain (Av/Av)^2	158.4893	10	0.501187234	19.95262315	15.84893
Gain (Av/Av)	12.58925	3.16227766	0.707945784	4.466835922	3.981072
Req (Noise Equivalent Resistance) Input	11	900	2500	1250	2.00E+04
Minimum Signal Level Output (V^2) (dBV)	-97.0103	-87.01029996	-90.01029996	-77.01029996	-65.0103
Minimum Signal Level Output (Vrms^2)	1.99E-10	1.99E-09	9.98E-10	1.99E-08	3.15E-07
Minimum Signal Level Output (Vo-p)	2E-05	6.30957E-05	4.46684E-05	0.000199526	0.000794

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