ARCHITECTURE, SYSTEM, AND CIRCUIT CONSIDERATIONS FOR SDR RECEIVER FRONT-ENDS

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ABSTRACT

In this paper we will discuss the RF front-end architecture for software definable radio (SDR). The paper will discuss various architectural issues for a homodyne Zero-Intermediate Frequency (ZIF) multi-band receiver. The RF Receiver is capable of handling Wide-Band CSMA (WCDMA), GSM, and 802.11 Wireless LAN standards. The RF circuits of the receiver Low-Noise Amplifier (LNA), Mixer, and the IF Filters are presented.

1. INTRODUCTION

There are a growing number of wireless standards for commercial and defense application. Each standard serves different needs and even within standards, there are differences in the frequency bands they support. For instance, the latest WCDMA specification covers six different frequency bands). From the RF performance perspective, most application outline a set of conditions such as noise, inter-modulation, blocking, power output and the spectral PSD of the transmitter. Most RF IC's attempt to push the filtering function to a lower frequency, which requires a more aggressive dynamic range requirements on the Intermediate Frequency (IF) and the base-band circuits. the increased dynamic range requirements use more power. Although it would be ideal to have a receiver that could cover all these different standards and different bands, it is unrealistic to expect this will happen in the near future.

A more realistic goal is to focus on a programmable radio that covers a subset of the possible standards for specific purposes. Although SDR is usually associated with military applications, we will focus on commercial standards, namely two cellular standards – WCDMA and GSM – a data transfer standard – 802.11 – and a location standard – GPS. We should expect that a general purpose receiver is not going to be competitive in any particular standard when compared to a receiver specifically designed for that standard; we will have to take a hit in cost, current draw, ease of implementation.

2. ARCHITECTURES FOR SDR FRONT-ENDS

The basic problem for a radio receiver can be simply stated: gain a small signal up; while knocking down strong interfering and blocking signals; and do it all without saturating the receiver. If we were able to get high Q tunable filters for the front-end, this task would be relatively simple, but such filters do not exist.

Conceptually, the simplest architecture for an SDR receiver would be a low noise amplifier followed by a wide-dynamic range ADC. For GSM, this would require somewhere in the region of 20 bits dynamic range with a sampling rate of at least 2x the highest frequency of 2GHz. Unfortunately, Walden[1] showed that there are fundamental physical limits that preclude building such a wide dynamic range ADC at multi-GHz frequencies.

Table 1: GSM 1800, WCDMA, GPS Spec

	GSM	WCDMA	GPS
Frequency Band(MHz)	1710-1785 1805-1885	1920-1980 2110-2170	1575.42
Channel Spacing	200KHz	5MHz	--------
Required C/N (dB)	9	7.2	10.2
Input Noise (dBm)	-120.8	-107	-138.5
Sensitivity @ BER=$10e^{-3}$ (dBm)	-100	-110	-157.2
Max power dBm	-15	-25	-------
SNR(dB)	9		

The most familiar architecture for a radio receiver is the super-heterodyne. The super-heterodyne receiver was invented by Armstrong in the 1930s and has been in use ever since. The strength of the super-heterodyne is that it gets around the filtering and gain issue by successively moving to lower frequencies where higher Q filters are available. These filters remove successively more of the interfering signals and allow the signal to be gained up away from the noise floor so following stages contribute less to the noise figure. The problem for the super-heterodyne is the need to use all these filters which are off-chip passive devices with fixed bandwidths. For a single frequency band, single standard receiver with only one IF frequency, we require three such filters – one front end roofing filter, one image reject filter following the LNA, and one IF filter . If we need to support multiple bands, only the IF filter could be shared, so for each new band we add

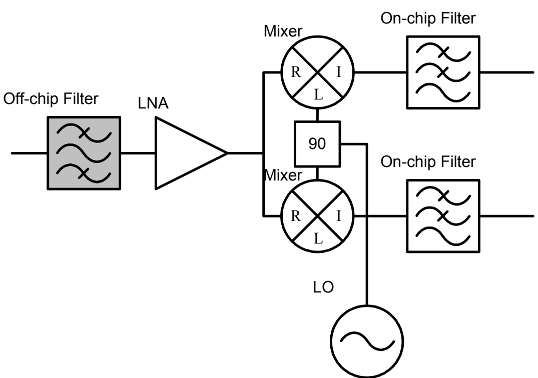

Figure 1. ZIF Receiver

two filters at the RF. Very soon this gets out of hand. Another problem for super-heterodyne receivers is the need for careful frequency planning and consideration of not just the signal and local oscillators, but all the harmonics and inter-modulations between them. For multi-band receivers this frequency planning quickly becomes a nightmare.

An architecture that has received some attention and would allow for the integration of a multiple standard receiver is the sub-sampling receiver. One example of such a receiver is the work of Shen et al [2]. The receiver works by sampling at a lower frequency than the RF frequency and relying on aliasing to bring the signal into the band of a switched capacitor filter working at the sampling frequency. Down conversion to baseband is achieved in successive sub-samplings with a decimation stage at the end of each to bring the signal down. Unfortunately, this scheme again runs into the problem of frequency planning that the super-heterodyne has. Also, strong blockers at multiples of the sampling frequency would be brought into band. Theoretically this could be controlled by using filters with high rejection at the particular problem frequencies, but for GSM, this would require filters with 120dB of rejection at RF frequencies which is just unrealistic.

The simplest receiver architecture is the direct-down conversion or Zero IF (ZIF) receiver (Figure 1). In this architecture the signal is mixed with a local oscillator signal which is at he same frequency as the desired signal. Both in-phase and quadrature components are needed for the local oscillator signal and two mixers at RF are used. The result is that the signal is brought down to DC and most of the gain and filtering is done at DC. This simplifies the front-end filtering requirements and also frequency planning becomes much simpler as only one main frequency is involved. Although it is one of the earliest architectures, the ZIF receiver suffers from many problems that have made its implementation difficult until recently. One of the most problematic is DC offset in the baseband chain. Any DC offset is gained up by the gain of the baseband chain, and, unless the signal modulation is such that AC coupling can be used, it is not possible to remove the DC offset easily. Another problem is coupling between the LO and RF ports in the mixers that causes the generation of a DC tone either from the LO mixing with itself, or a strong blocking signal self-mixing. With modern processes and very careful attention to detail on the part of the designer engineers and layout engineers, both these problems can be controlled and now DCR receivers are dominating the consumer radio world.

The simplicity of the ZIF architecture makes it a good candidate architecture for and SDR front-end implementation. The same is true on the transmit side, where many of the same benefits (and similar problems) hold true. In the remainder of this paper we will discuss the implementation of an SDR receiver front-end using the ZIF architecture.

3. SYSTEM LEVEL OVERVIEW

The most commonly used cellular standard in the World is GSM with about 70% market share. GSM is a so called 2G (digital cellular standard) with a follow-on 3G standard in WCDMA. GSM and WCDMA have been combined in one specification [3], and most WCDMA cellphones will also have to receive GSM channels, and this gives a strong need for a multi-mode or software defined radio.

Although the bandwidths of the channels are very different in WCDMA and GSM (5 MHz vs. 200 kHz), there are many similarities in the RF front-end requirements for the two systems. For GSM, the reference sensitivity is defined at -102dBm. For a 200kHz channel and GMSK modulation with a 9dB C/N requirement, this gives a maximum tolerable noise figure for the radio of:

$$NF = -102 - 9 - (-174 + 10\log(200k)) = 10dB$$

For WCDMA the situation is complicated by the spreading factor and processing gain. The reference sensitivity is defined for a 12.2kbps data stream, and this can be used to estimate the maximum noise figure given a required sensitivity of -117dBm and a required C/N of 7dB for the QPSK modulation used:

$$NF = -117 - 7 - (-174 + 10\log(12.2k)) = 9dB$$

For GSM, the required IIP3 is given for a required channel power of -99dBm and two interfering signals at -49dBm. With the noise at -111dBm (for 9dB C/N and a reference sensitivity of -102dBm), then to meet the standard, the third order intermodulation product of the two -49dBm signals needs to fall at -111dBm or less. This means that the IIP3 required is:

$$IIP3 = -49 + (-49 - -111)/2 = -18dBm$$

For WCDMA, the calculation is complicated by the processing gain, but the required IIP3 is -20dBm.

4. ANTENNAS AND FRONT-END FILTERING

The front-end filters are a problem for implementing an SDR. These filters are fixed by band, so to cover multiple bands requires multiple filters. Also, within bands, different standards require different filter performance. Many times the filter is actually designed to support a particular RF front-end.

One solution is to use multiple filters and multiple front-end LNAs. The receiver after the LNA can be combined. This solution rapidly becomes unmanageable for multiple standards in multiple bands. In future MEMs devices may offer some hope as they will allow matching elements to be switched in with low loss. This would allow the implementation of tunable antennas, which would enable the antenna to present a low impedance only at the frequency band of interest. This band would move depending on the antenna matching and also match into the LNA well only at the frequency band of interest.

5. LOW NOISE AMP

In the ZIF architecture, the components beyond the front-end filtering are usually integrated on-chip. This means that the LNA acts as an interface to the board and so must be matched properly. In the future if tunable antennas are available, this matching will be taken care of at the antenna. Variable matching can also be included on chip by switching in components or even using varactors driven from a DAC (but noise and linearity can be a problem here).

A triple mode LNA with broad and robust input and output matching has been designed. Both input and output are matched to 50 ohms. This stable LNA designed for 3G WCDMA 2.1 GHz, GPS 1.6 GHz, and GSM 1.8GHz standards.

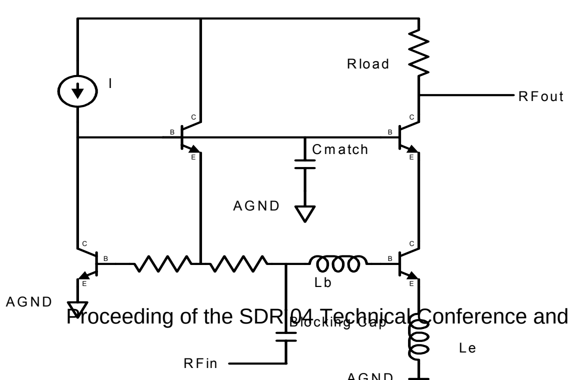

Figure 2. The multi-band, multi-standard LNA

The Cascode structure had been selected due to its high reverse isolation and the bandwidth flexibility. Reverse isolation is particularly important in ZIF receivers as the LO signal is at the same frequency as the RF and it is important that its level leaking through the front-end is minimized. Inductive degeneration is used at the emitter of the LNA for both matching and linearity purposes. LNA structure is shown in Figure 2.

The current was selected based on the required gain and linearity. The device size was properly selected for minimum achievable noise from the device. A new method [4] of matching has been introduced in which by adding proper impedance at the base of the common emitter of the cascode, an equivalent inductor will be seen from the RF input and this reduces the required matching input inductor while increasing the stability of the LNA. For GPS, a second amplifier would be added after this LNA to raise the signal even further above the noise floor. The total power consumption of the LNA including the PTAT current source is 22.41mW. The triple mode LNA extracted result is summarized in Table 2.

6. MIXER

The switches used in integrated mixers are inherently broadband. Unfortunately, we still have to have a broadband match into them to get maximum power transfer. This is best achieved using common base or common gate stages which are inherently broadband. The transformer based mixer described by MacEachern et al. [5] or a variation thereof also offers a useful way of coupling power into the mixer from the LNA. The LNA can receive its DC power through the inductive primary windings of the transformer or balun, and the secondary can be used to bias the mixer through a current sink at the common node of the secondary. The mixer is shown in Figure 3.

The mixer described in this paper uses an off-chip balun, but the intention is to integrate it at a later date. The balun drives into common base devices, which provide a match for the balun. The common base devices also provide a load for the LNA, which is on the other side of the balun, and some isolation from the switching quad of the mixer for the balun and LNA. The matching is controlled by the current mirror providing the current sink (the impedance looking into the common base stage is 1/gm) and switched capacitors or even varactors can be used to resonate out the inductance of the balun with a change in frequency. The results for the extracted mixer are shown in Table 1.

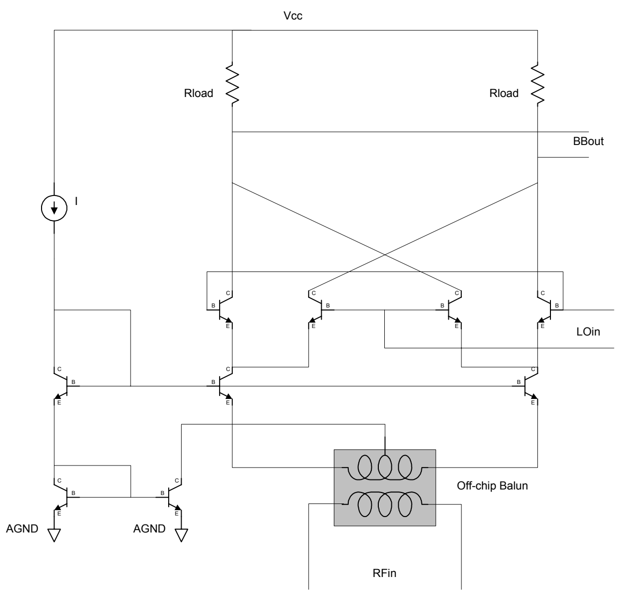

Figure 3. The multi-band, multi-standard mixer

7. FILTERING

The filtering in a ZIF receiver is provided by the baseband analog chain and digital. In a super-heterodyne receiver much of the filtering is passive with a relatively low insertion loss (5dB or less is not un-common) at a relatively high frequency (100MHz or more) so that the DC offset of the circuits is unimportant. For a ZIF receiver the filtering is active and hence relatively noisy (NF or 20dB or more) and as the filtering and necessary gain is happening at baseband, the DC offset and IIP2 of the filtering becomes important.

The filter needs to also gain the signal up for the ADC to detect, and also have variable gain. As the front-end will only have 30dB to 40dB of gain, there may need to be considerable gain in the baseband chain which may be a problem for DC offset. A better alternative, especially for continuous reception systems like WCDMA is to extend the dynamic range of the ADC so that the gain required in the analog domain is less and less gain changes are needed in the chain. This means that most of the filtering is pushed to the digital domain. In this case, the analog filter becomes more of an anti-alias filter rather than a channel filter.

Two types of filtering need to be considered: active RC and GmC. Active RC is common in super-heterodyne receivers and offers considerable linearity advantages over GmC, but it tends to have a limited frequency response and be noisier. GmC can be less noisy, but is also less linear than active RC. A combination of both may be a possibility (GmC out of the mixer for low noise, and active RC in later stages for better linearity with the gained up signal), but the performance of the two does not necessarily track and so there could be some movement in the poles in opposite directions causing some problems for the filtering function. For a loose anti-alias implementation, this may not be a problem and there may be ways in the baseband to fix things up.

Table 1. The performance of the individual blocks in the receiver

	LNA 1.6 GHz	LNA 1.8 GHz	LNA 2.1GHz	Mixer	Filter
Gain (dB)	17.76	17	15.94	19	0
IIP3	4.5	4.5	4.5	-4	15
IIP2	-	-	-	50	-
NF	1.2	1.1	1	8.5	27.3
VSWR1	1.28	1.3	1.4	-	-
Current(mA)	6	6	6	3.5	-
Power (mW)	-	-	-	-	71.28
Group delay(us)	-	-	-	-	0.6

A sixth order Butterworth filter is presented here with a NF of 26dB (which is sufficient for GSM and WCDMA). The filter is implemented as a cascade of three BiQuad sections. One BiQuad section is shown in Figure 4. The results for the filter are presented in Table 2 Sixth order filtering is probably overkill and generally third or fourth order filtering should be sufficient, but this work shows the capability of these filters. Switching between bandwidths is easily implemented by switching capacitors in and out and also by changing the resistors used to generate the transconductance.

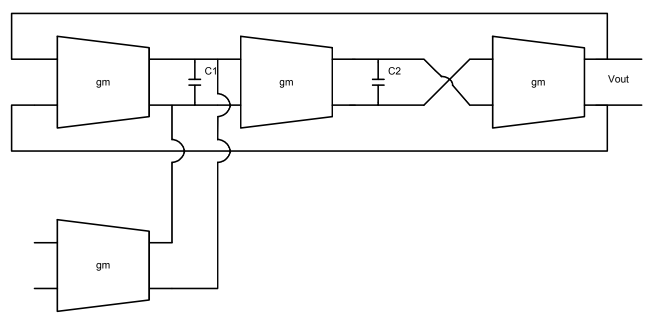

Figure 4. A single GmC Biquad section

8. ADC

The design of the ADC and the filtering are tied together. In general, the ADC has to have sufficient dynamic range that to convert all of the signal and the remaining blockers and interferers after the analog filtering to the digital domain. The dynamic range of the ADC must also be sufficient that the gained up noise of the front-end sits well above the quantization noise so that the quantization noise does not degrade the signal to noise ratio of the received signal.

For ZIF receivers, it is desirable to have as much dynamic range in the ADC as is possible to handle the DC offset that is inherent in these receivers. For wide dynamic range, oversampling ADCs based on the sigma-delta modulators are the best approach. For the wider bandwidth standards this will require running at clock rates in the hundreds of MHz, but modern IC processes are very capable of this sort of performance.
Although GSM requires over 100dB of dynamic range, WCDMA requires less than 80dB, and 802.11 a similar amount. This means that it is possible to exploit the noise shaping inherent in the modulator to get the required dynamic range.

We illustrate this with a second order modulator. The dynamic range can be approximated [6] knowing the order of the modulator, the oversampling ratio (OSR), and the number of bits of quantization (N):

$$DR = 6.02N + 1.76 - 12.9 + 50\log(OSR)$$

For a clock rate of 200MHz and a bandwidth of 200kHz for GSM, the calculated dynamic range is over 140dB. For WCDMA at the same clock rate and with a 4MHz channel, the calculated dynamic range is 80dB which will meet the requirements. For a single bit quantizer and a 20MHz channel for WLAN, the calculated dynamic range is 45dB. This is not enough, so the solution is to raise the clocking rate or increase the number of bits or both. Using a 5 bit quantizer, the dynamic range is now 68dB. The linearity of the DAC in the feedback path will limit the dynamic range in this case although with careful layout and design it may be possible to get a DAC with 12-bit linearity. Other approaches include using a cascaded or MASH structure in the WLAN case and only switching in the cascaded stages when required.

9. TRANSMITTER

Direct upconversion is by far the best choice for the SDR transmit for much the same reasons as direct downconversion is used in the receiver. Unfortunately, the transmit has much tighter requirements than the receiver because it is broadcasting to the world. This makes things much more difficult for the transmit designers.

The problem with direct upconversion transmiters is that any DC offset in the baseband is converted to carrier at the RF and needs to be below a limit that is specified in the specification. Gain partitioning schemes that use any programmable gain in the baseband would change the DC offset with each change in gain and so the correction would also have to change. In the case of WCDMA, which is continuous transmit, the settling of the gain change and correction would be a problem. One solution is to move all of the gain to the RF. This is very power hungry, but we made the assumption that the SDR would trade power and performance for flexibility, so this increase current drain would, presumably, not be a problem.

The power amplifier is another bottle-neck. Linear power amplifier which are needed by WCDMA and WLAN systems are difficult to implement in a flexible manner. Switching PAs offer an alternative here, but require off-chip filtering to filter out the switching. This filtering would need to be tunable if it were not to become a further bottle-neck. Again, in future, MEMs devices may bring this tenability with low insertion loss.

10. CONCLUSIONS

In this paper we have presented a receiver for SDR for use with GSM, WCDMA and 802.11 radios. This receiver is based on a Zero IF architecture. We have presented results for the LNA, mixer, and filters. The ADC could be done using an oversampling architecture running at several hundred mega-hertz pushing most of the processing to the digital domain. On the transmit side, we suggest that a direct upconversion architecture is the best choice. In the transmit case, the power amplifier will become a bottle-neck to progress.

11. REFERENCES

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[1] Walden, Robert H. “Analog-to-Digital Converter Survey and Analysis”, IEEE Journal on Selected Areas in Communications, Vol. 17, No. 4, April 1999, pp. 539-550.
[2] Shen, D.H.; Chien-Meen Hwang; Lusignan, B.B.; Wooley, B.A., “A 900-MHz RF front-end with integrated discrete-time filtering” IEEE Journal of Solid-State Circuits,Volume 31 , No. 12 , Dec. 1996, pp. 1945 - 1954.
[3] 3rd Generation Partnership Project, 3GPP Release 5, www.3GPP.org
[4] Darbanain, N., Analysis, Design, and Implementation of multi-mode Low Noise Amplifiers, M.Sc. Thesis, Arizona State University, Spring 2004.
[5] MacEachern, L.A.; Abou-Allam, E.; Wang, L.; Manku, T.; ”Low voltage mixer biasing using monolithic integrated transformer dc-coupling“, Circuits and Systems, 1999. ISCAS '99. Proceedings of the 1999 IEEE International Symposium on ,Volume: 2 , 30 May-2 June 1999, Pages:180 - 183 vol.2.
[6] David A. Johns, and Ken Martin, Analog Integrated Circuit Design, John Wiley and Sons, New York, 1997. p. 544.