On the Design of a Configurable UMTS/NAVSAT Transceiver

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ABSTRACT

This paper introduces a combined UMTS/NAVSAT receiver architecture. After a review of the state-of-the-art, a configurable UMTS/NAVSAT architecture is proposed. The investigated concept is based on a reconfigurable receive chain to prevent duplication of hardware, which will result in considerably lower costs. The biggest challenge in the design of NAVSAT receivers are the extremely tight noise figure (NF) requirements. Another key issue of integrated UMTS/NAVSAT receivers is the maximum tolerable UMTS transmit leakage injected into the NAVSAT receiver. An analytical derivation of the acceptable UMTS transmitter leakage for a certain ranging degradation is given. These values are compared with measured leakage values based on power amplifiers (PA) and surface acoustic wave (SAW) filters for UMTS.

I. INTRODUCTION

Navigation and location based services will be a key business driver in the field of mobile communication. Therefore, cellular phones and personal digital assistants (PDAs) will become the market leaders in the area of personal navigation applications. Market surveys forecast that the global GPS receiver market for automotive and mobile phone applications could reach around 55 million units in 2005. About 73% of this market, corresponding to around 40 million units, is expected to fall into the category of mobile phone applications. The planned global navigation satellite system (GNSS) modernization will undoubtedly further expand and improve applications for users in many fields by allowing combined use of such systems in hybrid receivers. These GNSS advances include the implementation of the Galileo system, now entering the development and validation phase under the cooperative management of the European Commission (EC) and the European Space Agency (ESA), as well as planned improvements in the U.S. counterpart, GPS. Taking all this into account, a combined solution of communication and navigation will be a key business in the future wireless field.

II. SYTEM OVERVIEW

A. Cellular

Table 1 shows the paired frequency bands for UTRA/FDD (UMTS Terrestrial Radio Access / Frequency Division Duplex).

Operating Band	UL Frequencies UE transmit, Node B receive	DL frequencies UE receive, Node B transmit
I	1920 – 1980 MHz	2110 –2170 MHz
II	1850 –1910 MHz	1930 –1990 MHz
III	1710-1785 MHz	1805-1880 MHz
IV	1710-1755 MHz	2110-2155 MHz
V	824 – 849 MHz	869-894 MHz
VI	830-840 MHz	875-885 MHz

Table 1: Frequency bands of the UTRA/FDD.

The UMTS air interface uses Wideband Code Division Multiple Access (W-CDMA), based on Direct Sequence Spread Spectrum (DS-SS). With DS-SS each user signal is spread by a user specific code. The most prominent advantage of DSSS-systems for cellular systems is its ability to eliminate the effect of multipath propagation by using a RAKE receiver in the mobile station. The choice of the user-specific codes employed for the spreading of the user signals greatly influences the overall performance of a CDMA system. The orthogonality among the spreading codes should be as large as possible. Otherwise the receiver will not be able to separate the different user signals due to multiple access interference (MAI).

The UMTS standard specifies a root raised-cosine (RRC) filter for pulse shaping, which determines to a large degree the spectral properties of the UMTS signals. The frequency response G_{rc} of the RRC-filter with roll-off factor α, transition type n (RRC for n=1, RC (raised cosine) for n=2) and chip duration T_C is defined by

$$G_{rc}(f)=\begin{cases} T_C & |f|<\frac{1-\alpha}{2T_C} \\ T_C\cos^n(\frac{\pi T_C}{2\alpha}\left(|f|-\frac{1-\alpha}{2T_C}\right) & \frac{1-\alpha}{2T_C}\le|f|\le\frac{1+\alpha}{2T_C} \\ 0 & |f|>\frac{1+\alpha}{2T_C} \end{cases} .$$

The pass-bandwidth of the above defined filter equals $(1+\alpha)/2T_C$, which results to 2.34 MHz for UMTS ($\alpha=0.22$, $T_C\approx260$ ns).

In Fig. 1 the impulse responses of an RRC and an RC filter are shown. The RC response results due to the RRC-filtering in the transmitter and the receiver (matched filter). It is clearly visible, that only the RC response is ISI free (zero crossing exactly at multiples of T_C). An important issue for the design of the analog transceiver is the peak to average power ratio (PAR). Due to the QPSK-like modulation format of the UMTS user signals and the fact that several user and control signals are summed up before converting them to the analog/RF-domain, the PAR can easily reach 9 dB with maximum of about 14 dB for the downlink (DL) and 4 dB with a maximum of about 6.5 dB for the uplink (UL).

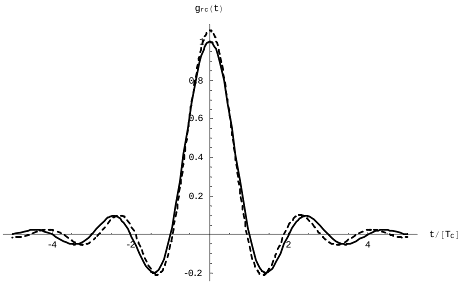

Fig. 1: RC and RRC (dashed line) impulse response.

B. NAVSAT

Fig. 2 shows the frequency bands of the Galileo signal structure.

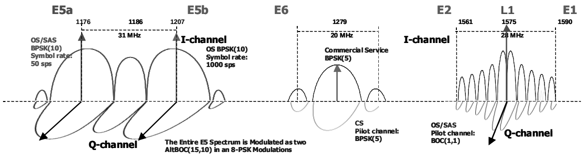

Fig. 2: Overview of the current status of the Galileo signal structure.

The following modulations are foreseen for the non-PRS Galileo signals:

- 2 AltBOC(15,10) modulated signals using an 8-PSK modulation. This modulation will be applied to the entire E5 band. It can be shown that the E5a and E5b can be tracked individually as Binary phase-shift keying with a chip-rate of 10.23 MHz BPSK(10).
- 2 BPSK(5): Binary phase-shift keying with a chip-rate of 5.115 MHz. This will be realized in a so-called Coherent Adaptive Sub-carrier Modulation (CASM) (this is required because of the third PRS signal on the same frequency).
- 2 BOC(1,1): Binary offset carrier with a chip-rate of 1.023 MHz, onto a square-wave with a frequency of 1.023 MHz. This is also known as Manchester coding. These two signals will also be part of a CASM.

In the receiver this can be treated as the following signals:

- E5b: BPSK(10) modulated signal with data modulated onto the spreading code.
- E5a: BPSK(10) modulated signal with data modulated onto the spreading code.
- E5a and E5b: AltBOC(15,10) modulated in QPSK with no data modulated onto the spreading code.
- E6: two independent BPSK(5) signals, one with data modulated onto the code and one without data.
- L1: two independent BOC(1,1) signals, one with data modulated onto the code and one without data.

For consumer applications in general and the transportation and tourism applications, which are the main focus of the GAWAIN project, in particular, the navigation signals transmitted at the L1 (1574.42 MHz) carrier will be the ones of highest commercial interest.

III. STATE-OF-ART RECEIVERS

A. Cellular

The homodyne receiver structure (also called zero-IF or direct-conversion architecture) depicted in Fig. 3 is the state of the art receiver structure for UMTS. The key advantage is the circumvention of the image signal because ω_{IF} is 0. As a result no image filter is required. This may also simplify the LNA (Low Noise Amplifier) design because there is no need for the LNA to drive a 50Ω load, which is often necessary when dealing with image rejection filters. Secondly, the IF-filter, which is usually an external SAW-filter, and the IF-amplifiers can be replaced by low-pass filters and baseband amplifiers that are amenable to monolithic integration.

This topology also entails a number of issues that do not exist or are not as serious in other receiver structures. Since in a homodyne topology the downconverted band extends to zero frequency, offset voltages can corrupt the signal and, more importantly, saturate the following stages. There are three main possibilities how DC offsets are generated. First, the isolation between the LO port and the inputs of the mixer and the LNA is not infinite. Therefore, a finite amount of feedthrough from the LO port to the mixer or the LNA input always exists. This LO leakage arises from capacitive and substrate coupling and, if the LO signal is provided externally, bond wire couplings. This leakage signal is now mixed with the LO signal, thus producing a DC component at the mixer output. This phenomenon is called self-mixing. A similar effect occurs if a large interferer leaks from the LNA or mixer input to the LO port and is multiplied by itself. A time varying DC offset is generated if the LO signal leaks to the antenna and is radiated and subsequently reflected from moving objects back to the receiver.

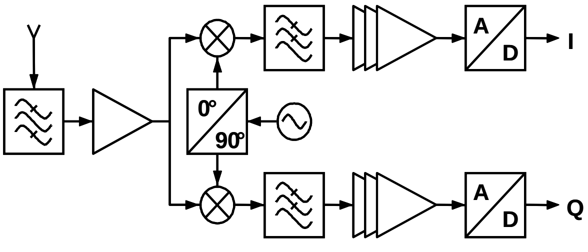

Fig. 3: Direct conversion receiver architecture.

Large amplitude modulated signals that are converted to the baseband section via second order distortion of the IQ mixers may also lead to time varying DC offsets. In order to prevent this kind of DC offset, a large second order Intercept Point (IP2) of the IQ mixer is necessary.

A UMTS compliant receiver needs approximately 80 dB of gain. Most of this gain is contributed by the baseband amplifiers. That means that even small DC offsets (in the range of several mV) at the mixer outputs may lead to DC levels sufficient to saturate the ADCs. The natural solution for DC offset cancellation is high-pass filtering. This approach is only possible because of the wideband nature of the UMTS signal. Other critical issues for the zero-IF receiver topology are IQ mismatches and flicker noise. Especially the latter one is highly problematic, if very low noise figures are required and/or the bandwidth of the wanted signal is in the same order of magnitude as the bandwidth in which the flicker noise is dominant.

B. NAVSAT

State-of-the-art GPS receiver front-ends are based on a low-IF architecture (see Fig. 4). This architecture comprises all benefits for a low-power, high-integration solution in complementary metallic oxide semiconductor (CMOS) technology while circumventing the aforementioned problems of a zero-IF architecture, such as dc-offsets, flicker noise and second harmonic distortion. The bandwidth of the RF front-end is approximately 4 MHz, to include the two main lobes of the Galileo signal as well as the main lobe of the GPS C/A code with two side lobes.

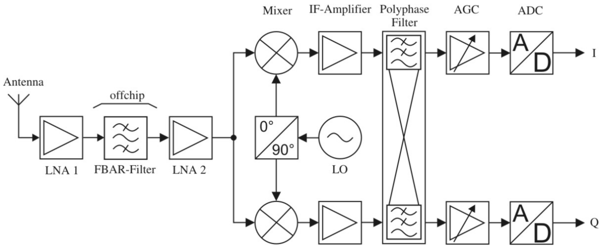

Fig. 4: Low-IF Front-End Architecture.

One major problem arises by the effect of self-mixing. The local oscillator (LO) and the LO-signal which is coupled into the RF cause a fluctuating DC-offset in the baseband. As the main slope of the GPS-signal lies around 0 Hz it would be severely degraded. The chosen low-IF architecture prevents this disadvantage by fixing the navigation signals at an IF of 3.5 MHz. Another advantage compared to the zero-IF topology is that the flicker noise between DC and 1 MHz is not a concern. Compared to a receiver with more than one mixer stage no further external filtering is required.

The limitation of the IF bandwidth is performed by a polyphase filter stage. The subsequent AGC further amplifies the signal to a detectable level and guarantees the optimal duty cycle of the three-bit ADC. A three-bit quantization reduces the SNR degradation to less then 0.7 dB. The target for implementation of the receiver front-end is an overall noise figure of 2 dB combined with low power consumption.

IV. PROPOSED CELLULAR/NAVSAT RECEIVER ARCHITECTURE

Within GAWAIN, a combined Zero IF/Low IF–Receiver, with a single reconfigurable front end, is proposed. With this concept it is possible to receive UMTS-signals with the Zero IF configuration and the Galileo signals with the Low-IF configuration with only one reconfigurable front end. Converting the received GNSS signal down to a low IF largely circumvents the distortions due to flicker noise. The structure of the receiver is shown in Fig. 5. Depending on what type of signal is received, the appropriate LO frequency is used and the analog and digital signal processing building blocks are properly configured.

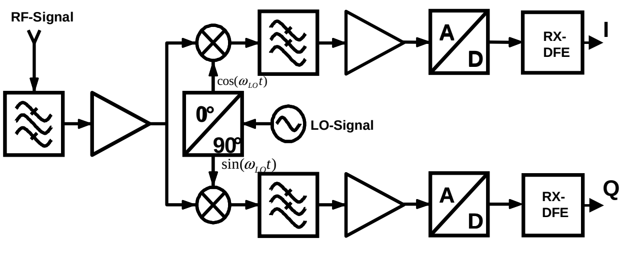

Fig. 5: Zero IF/Low IF Receiver Front-End.

The main tasks of the digital front end (DFE) are channel selection and decimation. A cascaded integrator-comb (CIC) filter [3] is used to decimate the ADC output by an integer factor. Channel selection filtering and the final integer decimation are realized by wave digital filters (WDFs) [4]. A finite impulse response (FIR) filter is used to correct the pass-band droop caused by CIC and WDF and for pulse filtering according to the different standards. To correct the pass-band group delay ripple, a configurable all-pass (AP) filter is used. For the baseband interface it is necessary to receive the signals by an integer factor of the symbol and so a fractional sample rate conversion (FSRC) must be implemented. For receiving GNSS signals the DFE must include a digital down conversion from low IF to baseband.

The ADC mode, the filter coefficients and the fractional decimation factor of the FSRC must be configured accordingly to enable signal reception in either UMTS or GPS/Galileo mode.

V. CODE TRACKING DEGRADATION DUE TO UMTS TRANSMIT SIGNAL LEAKAGE

The continuous transmission during an active UMTS connection has to be considered carefully in the design of a configurable UMTS/NAVSAT transceiver. The allowed spurious emissions in the UMTS standard [1] do not account for the frequency bands allocated for Galileo as described in section II. Table 2 shows the general minimum spurious emission requirements for operation in frequency band I, which is solely considered for the work within GAWAIN.

Frequency Bandwidth	Measurement Bandwidth	Minimum requirement
9 kHz ≤ f < 150 kHz	1 kHz	-36 dBm
150 kHz ≤ f < 30 MHz	10 kHz	-36 dBm
30 MHz ≤ f < 1000 MHz	100 kHz	-36 dBm
1 GHz ≤ f < 12.75 GHz	1 MHz	-30 dBm

Table 2: General spurious emissions requirements.

According to Table 2 3GPP compliant transmitters are allowed to have spurious emissions of up to -30 dBm measured within a measurement bandwidth of 1 MHz.

When the UMTS transmitter is located close to the Galileo/GPS receiver the ranging accuracy will be degraded. The achievable accuracy using a coherent DLL for the code tracking is given by

$$\sigma^2_{\text{celp}} = \frac{B_L(1-0.25B_L T)\int_{-\beta/2}^{\beta/2} G_s(f)\sin^2(\pi f\Delta)df}{C/N_0\left(2\pi\int_{-\beta/2}^{\beta/2} fG_s(f)\sin(\pi f\Delta)df\right)^2},$$

with B_L being the noise bandwidths of the DLL in Hz, T the Integration time of the correlators in s, Δ the Correlator spacing in s, $G_s(f)$ the Spectrum of the signal in s, β the two-sided bandwidth in Hz, and C/N_0 the Carrier-to-noise ratio in dB/Hz.

The spectral density for BOC modulated signals is given by

$$G_s(f) = \frac{1}{nT_c}\left(\frac{\sin(\pi f T_c)\sin(n\pi f T_c)}{\pi f\,\cos(\pi f T_c)}\right)^2 \quad \text{for n even,}$$

$$G_s(f) = \frac{1}{nT_c}\left(\frac{\sin(\pi f T_c)\cos(n\pi f T_c)}{\pi f\,\cos(\pi f T_c)}\right)^2 \quad \text{for n odd.}$$

Therefore, the variance of the range is inverse proportional to C/N$_0$ as can be seen in the equations above. Thus, the acceptable ranging degradation of 10 % can be translated into a C/N$_0$ degradation of 0.83 dB.

To estimate the resulting degradation caused by the UMTS transmitter interference the following well-known equation can be used [2]:

$$\left(\tfrac{C_S}{N_0}\right)_{eff} = \frac{\frac{C_S}{N_0}\int_{-\beta_r/2}^{\beta_r/2} G_S(f)df}{\int_{-\beta_r/2}^{\beta_r/2} G_S(f)df + \frac{C_i}{N_0}\int_{-\beta_r/2}^{\beta_r/2} G_i(f)G_S(f)df}$$

C_s and G_s are the carrier power and spectrum of the Galileo/GPS signal, respectively. C_i and G_i are the carrier power and spectrum of the interferer, respectively. In this case this corresponds to the spurious emission requirements. To reduce the out of band interferences it is possible to apply filters with high enough out-of band suppression.

To obtain a first estimate of the tolerable interference we assume that the filter of the GNSS receiver is a "brick-wall" filter, i.e. the filter has 0 dB gain inside the operative band, and the interference from the UMTS terminal is wide-band with a constant power spectral density of n_I. Under the last assumption the power spectral density injected into the Galileo/GPS receiver by the UMTS transmitter can be derived in the following way:

$$\frac{C_{umts}}{N_0}\int_{-\beta/2}^{\beta/2} G_{umts}(f)G_S(f)df = \frac{n_I\beta}{N_0}\int_{-\beta/2}^{\beta/2}\frac{1}{\beta}G_S(f)df = \frac{n_I}{N_0}\int_{-\beta/2}^{\beta/2} G_S(f)df.$$

If we further assume that the RF-bandwidth β is that large that most of the signal spectrum is contained within the integration region, then the effective C$_S$/N$_0$ further simplifies to:

$$\left(\frac{C_s}{N_0}\right)_{eff} = \frac{\frac{C_s}{N_0}}{1+\frac{n_I}{N_0}} =: \alpha^2\frac{C_s}{N_0}$$

$$\Rightarrow \quad \frac{n_I}{N_0} = \frac{1}{\alpha^2} - 1 \quad \Rightarrow \quad \alpha^2 = \frac{1}{\frac{n_I}{N_0}+1}$$

The following Figures show the degradation of C/No (Figure 6) and the degradation of the standard deviation of the pseudo range measurement (Figure 7) as a function of the interfering UMTS power spectral density n_I.

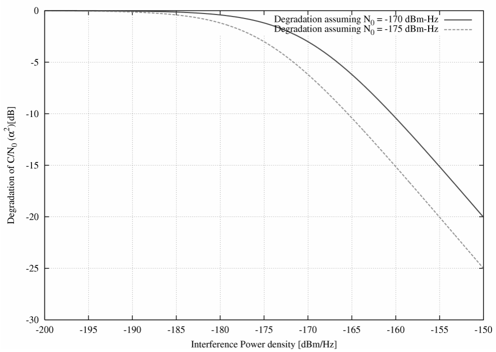

Fig. 6: Degradation of the carrier-to-noise ratio in dB as a function of the interfering UMTS spectral density (in dBm/Hz) for two values of assumed thermal noise density (No=-170 dBm/Hz and –175dBm/Hz).

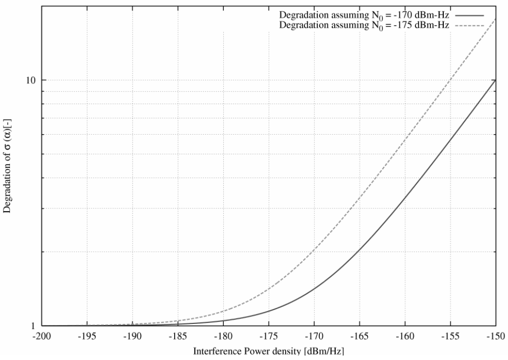

Fig. 7: Degradation of the pseudo-range accuracy as a function of the interfering UMTS spectral density (in dBm/Hz) for two values of assumed thermal noise density (No=-170 dBm/Hz and –175dBm/Hz).

Comparing the two figures above, we see that a degradation of for example 15 dB for C/N_0 will result in an increase of the standard deviation on the measurements (code and carrier) by a factor 6. To keep degradation of the standard deviation of the pseudo range measurement below 1.5 the UMTS transmit signal leakage into the Galileo/GPS band must not exceed the range of -175 dBm to -170 dBm.

VI. MEASURED UMTS TRANSMIT SIGNAL LEAKAGE

We performed a leakage measurement on a UMTS demonstrator to verify if the UMTS transmitter leakage power can be suppressed to a value below -175 dBm. Figure 8 shows the measurement setup. A signal generator (Rohde&Schwarz SMIQ) generates an UMTS signal, which is fed into the power amplifier (PA). To maximize the sensitivity of the spectrum analyzer (Rohde&Schwarz FSEA) a notch-and a SAW-filter were used. The notch-filter was tuned to the transmit frequency, whereas the SAW filter attenuated the harmonics of the transmit signal.

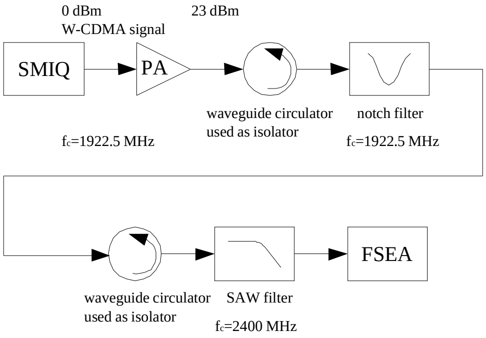

Fig. 8: Block diagram of the leakage measurement.

The measurement shows that the UMTS transmit leakage power level is around -133 dBm/Hz for all frequency bands except the E2-E1 band (see Table 3). In the E2-E1 band the level is around -129 dBm/Hz.

Operating Band	Galileo Bandwidth	UMTS noise
E5a	1164 MHz < f < 1289 MHz	-133 dBm/Hz
E5b	1189 MHz < f < 1214 MHz	-132.7 dBm/Hz
E6	1260 MHz < f < 1300 MHz	-133.4 dBm/Hz
E2 - E1	1561 MHz < f < 1590 MHz	-129.3 dBm/Hz

Table 3: Measured noise from the UMTS band I into the desired GALILEO bands.

Thus, a suppression of 37 dB to 42 dB is needed to keep the UMTS leakage power below -170 to -175 dBm/Hz. For the E2-E1 band the needed suppression increases to approximately 41 to 46 dB.

Fig. 9 shows the frequency response of a UMTS-SAW filter, the minimum guaranteed attenuation according to the vendor specification and the three GALILEO bands.

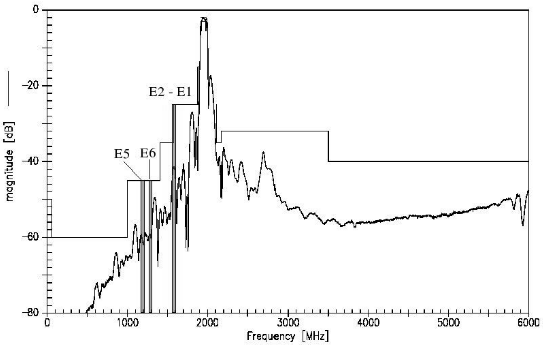

Fig. 9 Frequency response of the SAW transmitter filter

It is clearly visible that the attenuation of the SAW filter reaches a value of only 25 dB at the E2–E1 Galileo band edge. With this suppression the UMTS transmit leakage reaches a level of -154 dBm/Hz, which is more than the tolerable level of -170 dBm/Hz. However, the SAW-filter suppression for the other Galileo bands is high enough to keep the UMTS transmitter leakage below -170 dBm/Hz. Thus, special measures like additional filtering must be foreseen for the E2-E1 band.

VI. CONCLUSION

An important step into the market for Galileo is the timely availability of hybrid Galileo/GPS receivers in combination with wireless communications network positioning capabilities for consumer applications. We reviewed important signal format parameters of UMTS and Galileo and proposed a cellular/NAVSAT receiver architecture based on a zero-IF/low-IF receiver in combination with a digital front end. A first evaluation of the UMTS transmitter leakage into the NAVSAT receiver, which is one of the key issues for integrated cellular/NAVSAT receivers, has been presented. It shows that with a suitable SAW TX filter in combination with a standard UMTS transmitter the required leakage suppression should be feasible.

ACKNOWLEDGEMENTS

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