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Receiver specifications of the MIRAS demonstrator

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ABSTRACT

Improved weather forecast and climate monitoring rely on a better quantification of soil moisture (SM) and sea surface salinity (SSS). The most accurate way to measure them from remote sensors is by using radiometers at L band, where the sensitivity of the brightness temperature is maximum and the atmosphere is almost transparent. In order to provide a global coverage of these basic parameters within a revisit time of 3 days and 30-50 Km spatial resolution, the SMOS mission was recently selected by ESA in the frame of the Earth Explorer Opportunity Missions. The sensor being considered for this mission (called MIRAS) uses the concept of 2D interferometric radiometry^{1,2,3}, and is scheduled to be launched in 2005. A demonstrator consisting of a number of receivers is now under development by CASA (E) by contract with ESA. This work presents the block diagram of a single receiver chain in the SMOS instrument and its main requirements in terms of mission tradeoffs.

Keywords: L-Band radiometry. Interferometric Radiometry. Passive Remote Sensing. Aperture Synthesis

1. INTRODUCTION

A baseline of an interferometer consists of two receivers and a complex correlator, as depicted in figure 1, in which G_k and G_j are the available power gains of the chains, $H_{nk}(f)$ and $H_{nj}(f)$ their normalized frequency responses, $F_{nk}(\theta, \phi)$ and $F_{nj}(\theta, \phi)$ the normalized voltage patterns of the antennas and b_k and b_j the analytic signals of the receivers' output waves^{1,3}. The antennas are positioned in the XY plane, at $Z=0$ and near the origin of coordinates. The usual spherical coordinates (r, θ, ϕ) and director cosines ($\xi = \sin\theta \cos\phi$, $\eta = \sin\theta \sin\phi$) are defined, and a thermal source is assumed to be located somewhere in the $Z>0$ hemisphere far away from the antennas. The basic measurement of the instrument is the cross correlation of the analytic signals b_k and b_j , given by:

Figure 1 Block diagram of a baseline in the SMOS interferometer

$$\frac{1}{2} \langle b_k(t) b_j^*(t) \rangle = k_B V_{kj} \sqrt{B_k B_j} \sqrt{G_k G_j} ; \quad B_k = \int_0^\infty |H_{nk}(f)|^2 df \quad (1)$$

here k_B is the Boltzmann constant, B_k and B_j are the noise equivalent bandwidth of the receivers and V_{kj} is the visibility function, related to the brightness temperature map of the source T_B by:

$$V_{kj} = \frac{1}{\sqrt{\Omega_k \Omega_j}} \iint_{\xi^2 + \eta^2 \leq 1} \frac{T_B(\xi, \eta)}{\sqrt{1 - \xi^2 - \eta^2}} F_{nk}(\xi, \eta) F_{nj}^*(\xi, \eta) \tilde{r}_{kj} \left(-\frac{u_{kj}\xi + v_{kj}\eta}{f_0} \right) e^{-j2\pi(u_{kj}\xi + v_{kj}\eta)} d\xi d\eta \quad (2)$$

where Ω_k and Ω_j are the equivalent solid angles of the antennas, f_0 is an arbitrary center frequency, and the baseline u_{kj} and v_{kj} are given by the projections over the XY axes of the normalized spacing between antennas: $u_{kj} = (x_k - x_j)/\lambda_0$; $v_{kj} = (y_k - y_j)/\lambda_0$ being λ_0 the wavelength at f_0 and (x_k, y_k) (x_j, y_j) the XY coordinates of the antennas. The term $\tilde{r}_{kj}(\tau)$ is the *fringe-washing function*, which takes into account receivers frequency response non similarities. It is given by:

$$\tilde{r}_{kj}(\tau) = \frac{e^{-j2\pi f_0 \tau}}{\sqrt{B_k B_j}} \int_0^\infty H_{nk}(f) H_{nj}^*(f) e^{j2\pi f \tau} df \quad (3)$$

The value of the fringe-washing function at the origin ($\tau=0$) depends on the degree of non-similarity of the receivers frequency response. It is a complex number which will be represented by its amplitude and phase:

$$\tilde{r}_{kj}(0) = g_{kj} e^{j(\theta_k - \theta_j + \theta_{kj})} \quad (4)$$

For identical receivers $\tilde{r}_{kj}(0) = 1$. In general, however, the amplitude g_{kj} is smaller than unity. The term $\theta_k - \theta_j$ is the difference between the phase of the receivers' frequency response at the center frequency. The non-separable phase and amplitude terms θ_{kj} , g_{kj} are of second order effect, however, since the image to be recovered is highly low pass in the observation domain, they have non-negligible impact in the shortest baselines, which are the most significant.

If a digital 1bit-2level correlator is used, as it is foreseen in the MIRAS-SMOS instrument, the measured magnitude is actually the *normalized cross-correlation*^{2,3}:

$$\mu_{kj} = \frac{\langle b_k b_j^* \rangle}{\sqrt{\langle |b_k|^2 \rangle} \sqrt{\langle |b_j|^2 \rangle}} = g_k g_j \frac{V_{kj}}{T_A} ; g_k = \sqrt{\frac{T_A}{T_A + T_{Rk}}} ; g_j = \sqrt{\frac{T_A}{T_A + T_{Rj}}} \quad (5)$$

where g_k and g_j are separable error terms that have to be retrieved also in the calibration procedure, which are univocally related to the receivers equivalent noise temperature T_{Rk} and T_{Rj} . T_A is the antenna noise temperature, which is assumed to be equal for both receivers. It is measured with a well calibrated total power radiometer using an antenna located in the center of the array.

At this point, prior to analyze system and subsystem specifications in the following sections, some considerations must be pointed out. Radiometric accuracy ΔT versus the *standard deviation of the error* (parameter deviations from nominal value) shows in most cases a *linear dependence*⁴. This allows to represent the dependence of the radiometric error due to a given error as:

$$\Delta T|_j \approx S_j \sigma_{\text{error},j} \quad (6)$$

being S_j the *sensitivity* of the given error source to the radiometric accuracy. In particular, the sensitivity of all errors can be computed and reproduced in *Table 1* (sensitivity of the non-separable errors θ_{kj} , g_{kj} are specified taking into account the I-DNI calibration method^{5,6}):

Error	Sensitivity
Total separable amplitude error ($g_k g_j$)	0.59 K/%
Non-separable amplitude error (g_{kj})	0.28 K/%
In-phase error (separable) (θ_k, θ_j)	0.36 K°
Quadrature phase error (separable) (θ_q)	0.36 K°
Non-separable phase error (θ_{kj})	0.5 K°

Table 1. Sensitivities of receiver errors (antenna temperature of $T_A=200K$)

It is readily derived from *Table 1* that the required receiver specifications to achieve a good radiometric accuracy *without calibration* and/or characterization are beyond the state of the art of the technology. Hence, the calibration hardware and procedures must be devised as inherent parts of the receiver. It is, the receiver cannot be conceived independently of the calibration hardware and procedures.

2. SINGLE RECEIVER DESCRIPTION

Figure 2 gives a simplified block diagram of a single receiver which constitutes one of the signal channels in a baseline, as proposed for the SMOS-MIRAS demonstrator. As pointed out in the previous section, the receiver block diagram includes the SMOS-MIRAS foreseen calibration capability^{3,5}. The main outputs of the receiver are the in-phase, and quadrature components of the signal collected by the antenna. To do so, the receiver performs an I/Q down conversion to an intermediate frequency. Now, taking into account that the analytic signal $b(t)$ (*Figure 1*) can be written in terms of its I/Q components as $b(t)=i(t)+jq(t)$, the complex crosscorrelation is performed by means of real correlators as:

$$\langle b_k(t) b_j^*(t) \rangle = 2 \langle i_k(t) i_j(t) \rangle + 2 j \langle q_k(t) i_j(t) \rangle \quad (7)$$

As shown in Figure 2, the receiver has two measurement modes (vertical and horizontal brightness temperature) and two calibration modes (correlated and uncorrelated noise injection), which are selected by means of a four way switch. The signal is amplified, previously to the I/Q down conversion, by means of a Low Noise Amplifier (LNA). Subsystem specifications at the switch- LNA interface are critical, since determines the capability of the overall system to be calibrated to the required specifications. An isolator placed in front of the LNA avoids re-radiation of correlated noise generated by the amplifier to adjacent antennas⁷. A good return loss of the receiver, at the switch interface is also guaranteed. The signal is further amplified at the intermediate frequency by means of video amplifiers. Then is hard quantized (sign comparator) and sampled to give the I/Q samples (1bit/2 levels) to perform a bit-to-bit digital correlation. Main receiver parameters are: operating RF center frequency: $f=1413.5$ MHz; 3 dB cutoff frequencies: 1404-1423 MHz and an IF bandwidth: 8-27 MHz.

As shown in Figure 2, the receiver incorporates a power measurement system (PMS), which gives an estimation of the noise power at the system input (T_A+T_R) for calibration purposes^{5,6}. It operates as a simple radiometer, which is composed of a quadratic power detector followed by an integrator (low pass filter).

Once the receiver configuration has been described, the following sections are devoted to establish receiver specifications in terms of mission requirements. The separable phase and amplitude terms θ_k , θ_j , θ_q , g_k and g_j are estimated by the calibration process. Hence, subsystem tradeoffs are determined by the residual errors after calibration, and by those subsystem parameters that cannot be corrected by the foreseen calibration approach.

Figure 2. MIRAS-SMOS single receiver block diagram. Every baseline in the interferometer is composed of two receivers and a complex correlator, as depicted in Figure 1

3. RADIOMETRIC SENSITIVITY AND RELATED PARAMETERS

Main receiver specifications are determined by the required radiometric sensitivity (resolution), which is defined as the minimum temperature change at the input that can be detected at the output. There are two phenomena that limit the radiometric sensitivity: discretization and finite coverage of the spatial frequencies plane (u,v), and the signal to noise ratio. The radiometric resolution is obtained using the following equation^{3,8}:

$$\Delta T = \frac{\sqrt{3}}{2} d^2 \Omega_p \frac{(T_A + T_R)}{\sqrt{B \tau_{eff}}} \frac{\alpha_{LO}}{\alpha_F} \alpha_\omega \sqrt{N_v} \quad (8)$$

where (nominal values which are taken into account in the following calculations are given into parenthesis): T_A is the antenna temperature ($T_A=200$ K); T_R is the equivalent receiver's noise temperature, which includes antenna losses (T_R 169.6K, which is related to a noise figure NF=2.0 dB); d is the spacing between adjacent antennas in wavelengths ($d=0.89\lambda$, $f=1413.5$ MHz); $\alpha_{LO}=1$ for SSB receivers; $\alpha_{LO}=1.41$ for DSB receivers ($\alpha_{LO}=1$); $\alpha_\omega = \sqrt[4]{2} = 1.19$ for Gaussian filters; $\alpha_f = 1$ for rectangular filters ($\alpha_f = 1$); Ω_p elementary antenna equivalent solid angle ($\Omega_{eq}=1.7$ strad, related to a directivity D=8.7 dB); B is the noise bandwidth ($B=20$ MHz); $\tau_{eff}=\tau/Q$ is the effective integration time expressed as a

fraction of the real integration time τ . Q is a function of the oversampling ratio f_s/f_{max} and filter shape. ($f_s = 54\text{MHz}$, $B=8\text{-}27\text{MHz}$, rectangular filter, $Q=2.46$, giving for $\tau=0.78\text{ s}$ a $\tau_{eff}=0.32\text{ s}$); α_ω is a shape factor due to windowing, which is computed as:

$$\alpha_\omega = \sqrt{\sum_m \sum_n \frac{W_{mn}^2}{N_v}}; \quad N_v = 6N_{el}^2 + 6N_{el} + 1 \quad (9)$$

where N_{el} is the number of radiating elements per arm and N_v the number of visibility samples. This definition of α_ω makes it independent of the array size. Its value is 1, 0.52, 0.57, 0.54 and 0.45 for the rectangular, triangular, Hamming, Hanning and Blackman windows ($\alpha_\omega=0.45$, $N_{el}=27$ and $N_v=4537$). In order to evaluate the impact of different system and subsystem parameters it is more convenient to write this equations as:

$$\Delta T = \Delta T_0 \cdot DFC \cdot DFA, \quad \text{where } \Delta T_0 = \frac{(T_A + T_R)}{\sqrt{B\tau}}; \quad DFC = \sqrt{\frac{\tau}{\tau_{eff}}}; \quad DFA = \frac{\sqrt{3}}{2} d^2 \Omega_p \frac{\alpha_{LO}}{\alpha_F} \alpha_\omega \sqrt{N_v} \quad (10)$$

As it is readily seen, ΔT_0 gives the sensitivity of a traditional radiometer with the same parameters that the interferometer radiometer, and DFC and DFA are degradation factors. DFC is a degradation factor due to the type of correlation being performed which values DFC=1 for an analog complex correlator. DFA is a degradation factor due to array thinning. The nominal values for the parameters gives: $\Delta T_0=0.094\text{ K}$, $DFC=1.568$, $DFA=35.139$ and a snap shot radiometric sensitivity of $\Delta T=5.15\text{ K}$.

1. Receiver noise figure

All computations have been performed assuming a physical temperature of $T_0=290\text{ K}$. However, physical temperature T_{ph} impacts on T_{sys} through receiver equivalent noise temperature: $T_{sys}=T_A + T_R$. In this context T_R must be computed at the antenna plane:

$$T_R = T_{ph}(L-1) + T_{LNA}L \quad (11)$$

where T_{LNA} (80 K) is the low noise amplifier equivalent temperature (which depends on T_{ph}), and L ($L>1$) the interstage attenuation (including antenna losses). In order to evaluate the impact of T_{ph} on the radiometric sensitivity, T_R must be calculated at maximum and minimum operating temperatures (273 K to 323 K).

2. Antenna directivity

Antenna directivity impacts radiometric resolution through the equivalent solid angle Ω_p , which must be minimized while covering the whole field of view. However, the minimum attainable value for Ω_p is limited by the antenna spacing in the array (0.89λ is), and values is $\Omega_p=1.62\text{ strad}$ ($D_{max}=8.9\text{ dB}$), corresponding to an effective area of a circular dish of 0.89λ diameter. The nominal value assumed in the previous calculations for Ω_p was 1.7 strad ($D=8.7\text{ dB}$).

3. Integration time and pixel averaging

For the parameters of the proposed SMOS instrument³, every pixel remains in the field of view (at different positions) a time $\tau_{along-track}=114\text{ s}$. Hence radiometric resolution can be improved by means of a pixel averaging a factor :

$$DFT_{pixel\ averaging} = \sqrt{\frac{\tau_{along\ track}}{\tau_{snap\ shot}}} \quad (12)$$

Since the snap-shot integration time that was taken into account was $\tau_{snap-shot}=0.78\text{ s}$, the resulting radiometric resolution is $\Delta T_{pixel-averaging}=0.4\text{ K}$. However, it must be pointed out that averaging is actually performed over the geophysical parameter (by multiangular parameter retrieval), which does not depend on the observation angle, and radiometric resolution given by pixel averaging only gives a first approximation to the scientific mission tradeoff.

4. POWER, GAIN AND DYNAMIC RANGE REQUIREMENTS

1. Input dynamic range

The receiver measures the antenna temperature. For an antenna looking to an extended source, this is independent of the distance to the antenna, as long as the beam covers the whole source, which is basically the case of the MIRAS-SMOS elementary antennas. The input power is computed from the brightness temperature of the land (about 250 K) and sea (about 150 K). For all the system analysis, however, the noise power generated by the receivers themselves is effectively added to

the input. So the effective input power is the equivalent of $(T_A + T_R)$. In order to calculate the minimum input power absolute minimum values are considered for T_A and T_R : $P_{\min} = -103 \text{ dBm}$ ($T_A = 100 \text{ K}$, $T_R = 80 \text{ K}$, $B_n = 20 \text{ MHz}$). For calibrating purposes, correlated high temperature noise is injected to the receivers. The proposed value of the hot temperature for calibration is ten times 298K, as proposed in section 4.3. In this case, the maximum input power is: $P_{\max} = -90 \text{ dBm}$ ($T_{\text{HOT}} = 10 \times 298 \text{ K}$, $T_R = 80 \text{ K}$, $B_n = 20 \text{ MHz}$), which gives a moderate system dynamic range of 13 dB.

2. Main path signal gain, dynamic range and linearity

The specification for this gain comes from the operation of the samplers. The main limitation of these components is the offset voltage, which gives an amplitude error⁹. In order to make this error contribution negligible the minimum input voltage at the comparator must be larger than 100 times the offset: $\sigma_{V_{\min}} > 100 V_{\text{offset}}$. The offset voltage of the comparators is about 1.5 mV. In this case, the required gain is about $G = 100 \text{ dB}$ (50 ohm system). It must be pointed out that the signal which is processed is a voltage signal. Hence impedance changes at the video stage, if any, must be accounted for. Taking this gain value, 13 dB of voltage dynamic range at the samplers input gives a voltage margin : $\sigma_{V_{\min}} = 158 \text{ mV}$ and $\sigma_{V_{\max}} = 708 \text{ mV}$.

Linearity of the main path is not critical due to the use of 1-bit 2 level digital correlators, which are not sensitive to amplitude variations. The only effect that can be important is the AM-PM conversion which produces a phase error at high input amplitudes, in particular during calibration. The specification of linearity for the main path is better given in terms of AM-PM conversion rather than 1dB compression point. In order to make this effect negligible, a residual phase error after calibration has been chosen to be 0.1° . This means that the phase difference between calibration (noise injection) and scene measurement due to AM-PM conversion should be lower than 0.1° . In other words: *In all the dynamic range of the receiver, the total phase difference due to AM-PM conversion must be lower than 0.1°* . Now, assuming 13dB dynamic range, this means that the slope of the curve phase/amplitude must be lower than $0.0077^\circ/\text{dB}$. The translation of this requirement to the 1 dB compression point or output backoff operation depends on the specific technology used.

At this point it is worth to mention, that a sampling skew at the comparators produces a phase error given by⁹: $\Delta\phi = 2\pi f_{\text{IF}} \Delta\tau$. A constant separable phase error is corrected by means of the calibration process, however, a random sampling time jitter affects the radiometric resolution. Its impact has been set to $\Delta T = 0.38 \text{ K}$, from Table 1 this constraints the separable phase error to $\sigma_\phi = 1.055 \text{ deg}$, and, since the intermediate frequency is $f_{\text{IF}} = 17.5 \text{ MHz}$, the sampling time jitter is constrained to $\sigma_{\Delta\tau} = 0.167 \text{ ns}$.

3. The power measurement system

Each receiver includes a power measurement system (PMS) for calibration purposes. The separable amplitude errors are due to the nonzero noise temperature of the receivers. Since they cannot be kept low enough, they have to be estimated from the measurement of the receiver noise temperatures and the antenna temperature^{5,6}. The residual error after estimation must then be low enough to guarantee a small contribution to the radiometric accuracy. The error in the measure T_R can be directly related to the separable amplitude error (the error associated to the measurement of T_A can be reasonably considered negligible in this context). Now, taking into account that the sensitivity of the total separable amplitude error is $S_{\Delta V} = 0.59 \text{ K/\%}$, (Table 1), that $T_A = 200 \text{ K}$, $T_R = 169 \text{ K}$ ($F = 2.0 \text{ dB}$), to have an impact of, for example, 0.25 K in radiometric accuracy, the standard deviation of the errors in the estimation of the receiver noise temperatures must be:

$$S_{\Delta T_R} = \frac{S_{\Delta V}}{\sqrt{2} (T_R + T_A)} = 2.21 \text{ K} \quad (13)$$

The receiver noise temperature is estimated by measuring the PMS output for two loads at different temperatures, called HOT/COLD. This approach is very well known. The estimated receiver equivalent noise temperature T_R , and its uncertainty due to the power measurement error, are given by:

$$T_R = \frac{T_H P_C - P_H T_C}{P_H - P_C} ; \quad dT_R = \frac{T_H - T_C}{[P_H - P_C]^2} P_H P_C \left[\frac{dP_C}{P_C} - \frac{dP_H}{P_H} \right] \quad (14)$$

where T_H and T_C are the HOT and COLD (known) input temperatures and P_H and P_C the corresponding PMS outputs. Now, taking into account the worst combination case and a constant PMS measurement relative error $\Delta P/P$, the relative error is then given by:

$$\frac{\Delta T_R}{T_R} = \frac{(1+B)(A+B)}{(A-1)B} 2 \frac{\Delta P}{P}; \quad \text{where } A = \frac{T_H}{T_C}; \quad B = \frac{T_R}{T_C} \quad (15)$$

Assuming $B=0.58$ and $A=T_H/T_C=10$, as stated when computing the receiver dynamic range, to have a measurement accuracy of 2.21K over 169K, the uncertainty in the power detector measurement is constrained to:

$$\frac{\Delta T_r}{T_r} = 6.48 \frac{\Delta P}{P}; \quad \frac{\Delta P}{P} = 2 \cdot 10^{-3} \quad (16)$$

Note that an increase in T_{HOT} increases system's dynamic range with a minimum improvement of the required accuracy (17% when $T_{HOT} \rightarrow \infty$). The analysis above assumes perfect knowledge of the hot and cold temperatures. Obviously, this cannot be true, but this error can be associated to the error in the characterization of the noise distribution network.

The specification $\Delta P/P$ stands for linearity of the PMS which determines its input voltage range and thus gain. Optimum signal power to drive the PMS depends upon circuit configuration. Again, in order to have a first approximation of the required gain a simple computation can be done simply taking into account typical values. To drive the power detector in its linear region, a typical power of -25 dBm is required. Assuming that the detector is 50 ohm matched, this gives a gain of $G=65$ dB for the PMS path.

The PMS output contains information about the total noise power collected by the antennas and is used to calibrate the separable amplitude terms, which depend on noise temperature of the receivers. To this end, the detector output must be integrated, which can be done, either using an analog filter, or by averaging the samples once sampled. If the first choice is considered, the output filter must have the correct bandwidth to produce the wanted integration time. The relation between the integration time and the bandwidth of an equivalent low pass filter is $B=1/2\tau$. Assuming the integration time the same as for the visibility samples ($\tau=780$ ms for snapshot), this gives a filter bandwidth of $B=0.641$ Hz. The sampling frequency must then agree with the Nyquist criterion, thus it must be higher than 1.28 Hz. Further averaging can be performed using the samples, thus increasing the effective integration time.

5. ANTENNA AND RECEIVER RETURN LOSS

Antenna return loss is a critical parameter of the LICEF unit since it determines the capability of the system to be well calibrated. It must be defined at the antenna-switch interface, and specified for the radiator, the receiver (at switch input) and the noise distribution network. Return loss has two main effects: Receiver phase and amplitude errors. A simplified model to study the effect of matching is given by Figure 3. The radiator and the noise source are defined through their open circuit voltages e_r and e_n , and their reflection coefficients, whereas the receiver is specified through its scattering parameters:

$$\Gamma_r = \frac{Z_r - Z_o}{Z_r + Z_o}, \quad \Gamma_n = \frac{Z_n - Z_o}{Z_n + Z_o}, \quad \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \quad (17)$$

The system is referred to a reference impedance $Z_o=50\Omega$. Errors produced by unmatching are only taken into account at the receiver input, due to the calibration process. Output match can be considered reasonably good, and reverse voltage gain $S_{12} \approx 0$ to consider that $\Gamma_{in} \approx S_{11}$. Moreover, any unmatching after the amplifiers affect both the measurement and calibration mode and its effects (both in amplitude and phase) can be included in the forward voltage gain term, S_{21} .

1. Effect on receiver phase estimation

The measured voltage e_m at the receiver output can be written as:

$$e_m = \frac{1}{2} S_{21} \frac{1 - \Gamma_r}{1 - \Gamma_r \Gamma_{in}} e_r \quad (18)$$

Figure 3. Simplified model of the receiver. The antenna and the noise distribution network are modeled by means of their equivalent impedance and open circuit voltage.

where S_{21} is the receiver transfer response. If a pair of units "1" and "2" forming a single baseline are taken into account, the measured visibility is given by the complex cross-correlation of e_{m1} and e_{m2} :

$$\langle e_{m1} e_{m2}^* \rangle = \underbrace{\frac{1}{4} \langle e_{r1} e_{r2}^* \rangle}_{\text{ideal visibility}} \underbrace{S_{21}^1 S_{21}^{2*}}_{\text{receivers response}} \underbrace{\frac{1 - \Gamma_{r1}}{1 - \Gamma_{r1} \Gamma_{in1}} \left[\frac{1 - \Gamma_{r2}}{1 - \Gamma_{r2} \Gamma_{in2}} \right]^*}_{\text{error terms}}$$

Receiver response S_{21} is calibrated (in module and phase) by means of the noise injection method. However, the term originated by unmatching remains as an error. Since LICEF is not sensitive to amplitude due to the 1B/2L quantization, unmatching produces a phase error in the measured quantities, namely $(\phi_{er1} - \phi_{er2})$. This term can be associated to each single receiver, and it is given by:

$$\phi_{er} = \arg \left[\frac{1 - \Gamma_r}{1 - \Gamma_r \Gamma_{in}} \right] \cong \arg[1 - \Gamma_r] = \text{tg}^{-1} \left[\frac{|\Gamma_r| \sin \phi_r}{1 - |\Gamma_r| \cos \phi_r} \right] \approx |\Gamma_r| \sin \phi_r \quad (19)$$

It is assumed that the reflection coefficients are low enough. Therefore, unmatching produces a separable phase error term in each receiver ϕ_{er} . However, it must be pointed out that a constant phase error ϕ_{er} in all the receivers cancels out when computing the visibility, due to the conjugate term (eq.1) producing no error in the recovered image. Hence, only differences between the phases of the source reflection coefficient are of significance. Hence, if we consider that the dispersion of ϕ_r is restricted to a quadrant $(-\pi/4, \pi/4)$, the uncalibrated phase term ϕ_{er} is a zero mean error of standard deviation given by: $\sigma_{\phi_{er}} = 0.4328 |\Gamma_r|$. Antenna return loss has been specified to 25 dB. This gives a reflection coefficient $|\Gamma_r| = 0.056$, an uncalibrated phase error of $\sigma_{\phi_{er}} = 1.39$, which, taking into account *Table 1* gives a radiometric error of $\Delta T = 0.5$ K.

6. FREQUENCY RESPONSE AND INTER-RECEIVER SIMILARITY

The frequency response of the system is fixed by rejection to interference, which is mainly dominated by nearby L-band radars¹⁰. The interference rejection mask holds for the total receiver, not only the RF filter. This means that it should be applied to the *combined frequency response of RF and IF* eventually measured at the IF port. However, for the SMOS-MIRAS instrument system frequency response is basically shaped by the RF filter. A type of filter, compliant with the interference rejection is a Chebychev 8th order filter, with nominal pass-band ripple of 0.1 dB, and nominal bandwidth at ripple frequencies of 19 MHz.

Inter-receiver similarity is also a key parameter in an interferometric radiometer, through the fringe-wash term that produces the non-separable phase and amplitude error terms (θ_{kj}, g_{kj}) . These error terms are partially corrected by the calibration procedures (i-DNI method⁶). In order to set degradation of radiometric accuracy due to this terms to about 0.4 K, they must be restricted to 1.5% of amplitude error ($g_{kj} > 0.9850$) and 0.8 degrees of phase error (*Table 1*). Now, this specification in non-separable amplitude and phase errors must be translated into a filter similarity specification¹¹. *Figure 4* gives the allowed filter deviation mask, both in amplitude and in group delay. The inner lines correspond to 2/3 of all the RF filters, while the outer lines correspond to the maximum peak-to-peak deviation from the nominal values.

7. CONCLUSIONS

This paper presented the MIRAS-SMOS receiver block diagram. In order to establish a radiometric accuracy, system and subsystem specifications have been derived taking into account technologic limitations and mission tradeoffs. This has limited individual contributions to radiometric accuracy below 0.5 K.

Figure 3. Inter-receiver similarity. Amplitude and group delay deviation masks

8. ACKNOWLEDGMENTS

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