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A new calibration technique for interferometric radiometers

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ABSTRACT

This paper presents a new method for receiver calibration of interferometric radiometers. It is based on the distributed noise injection mechanism and it consists of fully calibrating the baseline error terms of the central antennas (which share the same noise source) while keeping only separable error calibration for the distant ones. This improves the accuracy of the shortest baselines, which are the most significant ones due to the smoothness of the brightness temperature to measure. Simulations show that, compared to previously reported methods, the improvement on the radiometric resolution can be as high as 3.9. The robustness against frequency response mismatch between receivers is improved by a factor 2.9

Keywords: Calibration. Interferometric Radiometry. Passive Remote Sensing. Aperture Synthesis

1. INTRODUCTION

One of the key geophysical parameters for climate modeling is soil moisture, which can be monitored from space using passive remote sensing at 1.4 GHz. The most promising concept is aperture synthesis using interferometric techniques, which was first proposed by NASA in the 80's in a 1-D configuration (ESTAR)¹. ESA is now developing a space borne 2-D instrument (MIRAS)², aiming at providing global coverage of soil moisture. One of the still open issues is the instrument calibration, which has to be performed periodically on board to compensate for system drifts. The purpose of this paper is to present a new method of receiver calibration (not including the antenna errors) for large interferometric radiometers such as MIRAS.

Calibration is the process of obtaining the errors of an instrument with respect to an ideal one by means of measuring known standards. In the case of a total power radiometer, which measures the brightness temperature of a spot, calibration is carried out by measuring thermal noise emitted from a source at known temperature. An interferometric radiometer measures a complete map of brightness temperature in a single snapshot. Thus, from a fundamental point of view, perfect calibration is not possible without the measurement of a known scene. However, since this is usually not available during the normal operation of a space-borne instrument, approximate methods of periodic calibration must be devised. One of them is the injection of correlated noise, although, obviously, this procedure cannot deal with antenna errors.

2. DEFINITION OF TERMS

1. General equations

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, which may or not include frequency conversions, $F_{nk}(\theta, \phi)$ 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.

The antennas are positioned in the XY plane, at $Z=0$ and near the origin of coordinates. The usual polar 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 self- and cross-correlation of the output signals are given by:

$$\begin{aligned} \frac{1}{2} \langle |b_k(t)|^2 \rangle &= k_B T_A B_k G_k + k_B T_R B_k G_k ; \quad \frac{1}{2} \langle |b_j(t)|^2 \rangle = k_B T_A B_j G_j + k_B T_R B_j G_j \\ \frac{1}{2} \langle b_k(t) b_j^*(t) \rangle &= k_B V_{kj} \sqrt{B_k B_j} \sqrt{G_k G_j} \end{aligned} \quad (1)$$

Figure 1. A baseline

where k_B is the Boltzmann constant, T_{Rk} and T_{Rj} are the equivalent noise temperature of the receivers, B_k and B_j are their noise equivalent bandwidth $\left(B_k = \int_0^\infty |H_{nk}(f)|^2 df\right)$, T_{Ak} and T_{Aj} are the antenna noise temperatures and V_{kj} is the visibility function, related to the brightness temperature map of the source T_B by³:

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

where Ω_k and Ω_j are the equivalent solid angles of the antennas, f_0 is an arbitrary center frequency, and u_{kj} and v_{kj} are 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.

Finally the term $\tilde{r}_{kj}()$ is the *fringe-washing function*, defined as:

$$\tilde{r}_{kj}(\tau) = \frac{e^{-j2\pi f_0 \tau}}{\sqrt{B_k B_j}} \int_0^\infty H_{nk}(f) H_{nj}^*(f) e^{j2\pi f \tau} df = \frac{1}{\sqrt{B_k B_j}} \mathcal{F}^{-1} [H_{nk}(f + f_0) H_{nj}^*(f + f_0) u(f + f_0)] \quad (3)$$

where $u()$ is the unit step function and $\mathcal{F}^{-1}$ means the inverse Fourier Transform. 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:

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

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, and the non-separable phase term θ_{u_k} is of second order effect, although it may have non-negligible impact in the calibration procedures⁴.

A first-order approximation for equation (2) is obtained assuming the field of view sufficiently small to neglect the argument of the fringe-washing function and antennas with identical pattern. In this case, this equation is simply reduced to:

$$V_{kj} = V(u_k, v_j) \approx \tilde{r}_{kj}(0) \frac{1}{\Omega_k} \mathcal{F}_{\xi, \eta} \left[\frac{T_B(\xi, \eta)}{\sqrt{1 - \xi^2 - \eta^2}} |F_{nk}(\xi, \eta)|^2 \right] \quad (5)$$

where $\mathcal{F}_{\xi, \eta}$ stands for two dimensional Fourier transform in the variables ξ and η . That is, in the case of identical antennas and for small FOV, after correcting for the fringe-washing function at the origin to account for receivers' differences, the visibility function is related to the so called "modified brightness temperature", defined as $T_B(\xi, \eta) |F_{nk}(\xi, \eta)|^2 / \sqrt{1 - \xi^2 - \eta^2}$, by a two-dimensional Fourier transform. This gives a straightforward algorithm to perform the inversion of (2). From this point of view, it is clear that the receiver calibration procedure is essentially a means of measuring $\tilde{r}_{kj}(0)$. Nevertheless, two other effects, due to the specific hardware used in the receivers, have to be taken into account. This is addressed in the following paragraphs.

2. Separable Amplitude errors (noise temperature of the receivers)

If a digital 1bit-2level correlator is used, as it is foreseen in the ESA's MIRAS instrument, the measured magnitude is actually the *normalized cross-correlation*⁵. From (1), assuming that $T_{Ak} = T_{Aj} = T_A$, it is readily obtained:

$$\mu_{kj} = \frac{\langle b_k b_j^* \rangle}{\sqrt{\langle |b_k|^2 \rangle} \sqrt{\langle |b_j|^2 \rangle}} = g_k g_j \frac{V_{kj}}{T_A} \quad (6)$$

where

$$g_k = \sqrt{\frac{T_A}{T_A + T_{Rk}}} \quad ; \quad g_j = \sqrt{\frac{T_A}{T_A + T_{Rj}}} \quad (7)$$

are separable error terms that have to be retrieved also in the calibration procedure. They are univocally related to the receivers noise temperature and, for ideal noiseless receivers are unity. The antenna temperature T_A is measured with a well

calibrated total power radiometer using an antenna located in the center of the array. Thus its value is assumed known and equal for all the antennas.

3. Quadrature error

A complex correlator is actually implemented using a phase-quadrature down converter, that produces for each receiver an “in-phase” and a “quadrature” outputs (i_k and q_k respectively for receiver k). Quadrature error is produced by the error in the 90° phase shift of the Q-path local oscillator, as well as by different lengths (or group delays) between the I and Q paths of one receiver. It is also produced by sampling errors in the comparators⁶.

For a given baseline (k-j), two *real* normalized cross-correlations are actually measured ($i_k \otimes i_j$ and $q_k \otimes i_j$) which are related to the complex normalized cross-correlation given in (6) by:

$$\mu_{i_k i_j}^{\Delta} = i_k \otimes i_j = \text{Re}[\mu_{kj} e^{i(\phi_k - \phi_j)}] \quad ; \quad \mu_{q_k i_j}^{\Delta} = q_k \otimes i_j = \text{Im}[\mu_{kj} e^{i(\phi_k - \phi_j)}] \quad (8)$$

where ϕ_{ik} , ϕ_{ij} and ϕ_{qk} are the phase errors in the i path (k and j receivers) and in the q path (k receiver) and they should vanish for ideal receivers. The 90° phase shift of the LO has already been taken into account in (8), since $q_k \otimes i_j$ is expressed as the “imaginary part” of the complex correlation corrupted by the errors. Including (8) in (6), the following is obtained:

$$\begin{bmatrix} \mu_{i_k i_j} \\ \mu_{q_k i_j} \end{bmatrix} = g_k g_j g_{kj} \begin{bmatrix} \cos A_{kj} & \sin A_{kj} \\ -\sin B_{kj} & \cos B_{kj} \end{bmatrix} \begin{bmatrix} \text{Re}[\hat{V}_{kj}] \\ \text{Im}[\hat{V}_{kj}] \end{bmatrix} \quad (9)$$

where

$$A_{kj} = \theta_{oj} - \theta_{ok} + \frac{\theta_{qj}}{2} - \frac{\theta_{qk}}{2} - \theta_{kj} \quad ; \quad B_{kj} = \theta_{oj} - \theta_{ok} + \frac{\theta_{qj}}{2} + \frac{\theta_{qk}}{2} - \theta_{kj} \quad (10)$$

being $\theta_{ok}-\theta_{oj}$ the total phase difference between receivers (including LO phase errors), and $(\theta_{qj} \pm \theta_{qk})/2$ the total quadrature errors. These are related to the previously defined phase terms (4), (8) by:

$$\theta_{ok} = \theta_k + \frac{1}{2}(\phi_{ik} + \phi_{qk}) \quad ; \quad \theta_{oj} = \theta_j + \frac{1}{2}(\phi_{ij} + \phi_{qj}) \quad ; \quad \theta_{qk} = \phi_{ik} - \phi_{qk} \quad ; \quad \theta_{qj} = \phi_{ij} - \phi_{qj} \quad (11)$$

In (9) the *corrected* visibility $\hat{V}_{kj}$ has been introduced and it is defined as:

$$\hat{V}_{kj}^{\Delta} = \frac{V_{kj}/T_A}{\tilde{V}_{kj}(0)} \quad (12)$$

Using (5) it is apparent that, for the small FOV and identical antennas approximation, this “corrected” visibility is directly the 2D Fourier transform of the modified brightness temperature normalized to T_A . Fourier Transform inversion of $\hat{V}_{kj}$ gives, then, a first order approximation of the modified brightness temperature map of the source.

From (9) it follows that a baseline has three calibration parameters: One amplitude ($g_k g_j g_{kj}$), and two phases (A_{kj} and B_{kj}).

3. INJECTION OF CORRELATED NOISE. CENTRALIZED DISTRIBUTION

1. Correlation measurement

If, for each baseline, the antennas are substituted by sources of correlated noise at equivalent temperature T_c , this is equivalent to locate a point source at equal distance from the receivers and assume omnidirectional antennas. In this case, from (2) the “calibration” visibility is given simply by

$$V_{kj}^c = \tilde{V}_{kj}(0) T_c \quad (13)$$

where the superscript “c” stands for “calibration”. The normalized cross-correlation (the magnitude that is really measured) comes from (6), using the above calibration visibility, and taking into account that now $T_A = T_c$:

$$\mu_{kj}^c = g_k^c g_j^c \tilde{V}_{kj}(0) \quad (14)$$

The amplitude error terms are now given by:

$$g_k^c = \sqrt{\frac{T_c}{T_c + T_{R_k}}} \quad g_j^c = \sqrt{\frac{T_c}{T_c + T_{R_j}}} \quad (15)$$

Hence the measurement of the normalized complex cross-correlation when correlated noise is injected at the receivers' input, allows to retrieve the term $\tilde{\mu}_{kj}(0)$ of a given baseline, provided the noise temperature of the receivers and the source temperature are known. When quadrature error is considered, the measured real correlations are related to the error terms by a matrix equation, as in (9):

$$\begin{bmatrix} \mu_{i_k i_j}^c \\ \mu_{q_k i_j}^c \end{bmatrix} = \mathbf{g}_k^c \mathbf{g}_j^c \mathbf{g}_{kj} \begin{bmatrix} \cos A_{kj} & \sin A_{kj} \\ -\sin B_{kj} & \cos B_{kj} \end{bmatrix} \begin{bmatrix} 1 \\ 0 \end{bmatrix} \quad (16)$$

where $\mathbf{g}_{kj}$, A_{kj} and B_{kj} are the same as in (9). Since, in general, $\mathbf{g}_k^c \neq \mathbf{g}_k$ and $\mathbf{g}_j^c \neq \mathbf{g}_j$, it follows that cross-correlation measurements when noise is injected is not enough for a complete baseline calibration.

2. Power measurement

An extra equation may be obtained by measuring the output power (before 1 bit A/D conversion) of the receivers. This can be done with a simple temperature compensated quadratic detector, which implies that all receivers should have this capability. The ratio of detected power when the input switch of receiver "k" is alternatively connected to the antenna (T_A) and to the correlated source (T_c) is:

$$Y_k = \frac{N_{\alpha_k}}{N_{A_k}} = \frac{k_B(T_c + T_{R_k})B_k G_k}{k_B(T_A + T_{R_k})B_k G_k} = \frac{T_c + T_{R_k}}{T_A + T_{R_k}} = \frac{T_c}{T_A} \left(\frac{\mathbf{g}_k}{\mathbf{g}_k^c} \right)^2 \quad (17)$$

where (7) and (15) have been used. The total amplitude term during measurement is then related to the total amplitude term during calibration by:

$$\mathbf{g}_k \mathbf{g}_j \mathbf{g}_{kj} = \mathbf{g}_k^c \mathbf{g}_j^c \mathbf{g}_{kj} \frac{T_A}{T_c} \sqrt{Y_k Y_j} \quad (18)$$

Thus, by measuring the power ratio Y in each receiver (two measurements per baseline), there is no need to know the equivalent noise temperature of the receivers, since the amplitude terms are related by a measured quantity.

3. Quadrature error measurement

According to (10), the reason for what A_{kj} and B_{kj} are different is because of the quadrature error. If a measurement of this can be independently performed, the number of unknowns will be reduced. The quadrature error of a given receiver can be measured by correlating directly its I and Q outputs. For a quadrature error-free receiver this correlation should be null. On the contrary, considering the receiver k having quadrature error θ_{qk} , (see (11)) it is straightforward to obtain:

$$\mu_{i_k q_k} = \sin \theta_{qk} \quad (19)$$

where the amplitude term of this normalized correlation must be 1 because the noise temperature of the receiver depends only on its RF part, which is common to the I and Q paths. Inserting (19) into (10), the following is obtained:

$$A_{kj} = \alpha_{kj} + \frac{1}{2}(\arcsin \mu_{i_k q_k} - \arcsin \mu_{i_k q_j}) \quad ; \quad B_{kj} = \alpha_{kj} + \frac{1}{2}(\arcsin \mu_{i_k q_k} + \arcsin \mu_{i_k q_j}) \quad (20)$$

where α_{kj} contains the in-phase separable and non-separable phase errors:

$$\alpha_{kj} = \theta_{oj} - \theta_{ok} - \theta_{kj} \quad (21)$$

4. Baseline calibration equations

Introducing the above equations in (16), the parameters α_{kj} and $\mathbf{g}_k^c \mathbf{g}_j^c \mathbf{g}_{kj}$ (denoted in the following as $\tilde{\mathbf{g}}_{kj}^c$) are readily retrieved:

$$\alpha_{kj} = \text{atan} \left(- \frac{\mu_{i_k i_j}^c \sin Q'_{kj} - \mu_{q_k i_j}^c \cos Q_{kj}}{\mu_{i_k i_j}^c \cos Q'_{kj} - \mu_{q_k i_j}^c \sin Q_{kj}} \right) ; \quad \mathbf{g}_k^c \mathbf{g}_j^c \mathbf{g}_{kj} = \tilde{\mathbf{g}}_{kj}^c = \frac{\mu_{i_k i_j}^c}{\cos(\alpha_{kj} + Q_{kj})} = \frac{-\mu_{q_k i_j}^c}{\sin(\alpha_{kj} + Q'_{kj})} \quad (22)$$

where $\mu_{i_k i_j}^c$ and $\mu_{q_k i_j}^c$ are measured normalized real correlations, and:

$$Q_{kj} = \frac{1}{2}(\arcsin \mu_{i_k q_k} - \arcsin \mu_{i_k q_j}) \quad ; \quad Q'_{kj} = \frac{1}{2}(\arcsin \mu_{i_k q_k} + \arcsin \mu_{i_k q_j}) \quad (23)$$

being also μ_{i,q_k} and μ_{i,q_l} obtained from measurements. It is worth noting that, if there were no quadrature error ($\theta_{qk}=\theta_{ql}=0$), from (19) and (20) it follows that $A_{kj}=B_{kj}=\alpha_{kj}$ and then equations (22) become simply:

$$\alpha_{kj} = \text{atan}\left(\frac{\mu_{q_k,l}^c}{\mu_{q_l,l}^c}\right) ; \quad \tilde{g}_{kj} = \sqrt{(\mu_{i,i}^c)^2 + (\mu_{q_k,l}^c)^2}$$

as can be obtained directly from (16).

Using the above procedure, the parameters α_{kj} and $\tilde{g}_{kj}^c = g_k^c g_j^c g_{kj}$ have been obtained out of three correlation measurements. Finally, introducing these values into (18) and (20), the baseline unknowns A_{kj} , B_{kj} and $g_k g_j g_{kj}$ are obtained. Since it has been assumed correlated noise as input, the above equations should be applied to baselines formed by receivers that share the same noise source. To fully calibrate the interferometer, all receivers must then be connected to the same noise source, which is only possible with a centralized noise distribution network and a single noise source.

4. DISTRIBUTED NOISE INJECTION

To avoid the electrical and mechanical problems associated with a centralized noise distribution network, a distributed noise injection mechanism (DNI) was proposed in a previous work⁷. In this case, the receivers of the interferometer are grouped in sets of antennas sharing the same noise source, as is shown in Figure 2 for the MIRAS geometry. The method was based on the assumption that the separable error terms are small enough to be neglected (namely $g_k \approx 1$ and $\theta_{ij} \approx 0$). In this case, if correlated noise is injected to sets of eight adjacent antennas, applying equation (16) a set of overdetermined equations is obtained⁷. Solving this system allows to compute the total in-phase (θ_0) and quadrature error (θ_q) of each one of the eight receivers and also the separable amplitude terms g_k and g_j . This procedure is repeated for all sets of eight antennas of the array. Overlapping sets of antennas are needed to establish the reference phase and to avoid the precise knowledge of all the noise sources.

1. Improved distributed noise injection method (i-DNI)

An improved calibration method, using distributed noise injection is presented here. It essentially consists of applying a centralized distribution approach (section 3 above) to the receivers that share the same noise source, particularly the central ones of the Y-shape array, and a distributed approach to the baselines from receivers of two different sets. This allows to calibrate all the separable and non-separable errors of the *baselines* formed by the receivers of a given set, and also the separable errors of the individual receivers that belong to different sets. As a consequence, without any further hardware requirements, the twelve central receivers (the central one is used for measuring the antenna temperature with a total power radiometer) and also the sets of eight belonging to a given set can be compensated for all the errors, while the rest are only compensated for separable ones. The residual error with respect to a full calibration of all baselines (which would imply a single distribution network for the whole structure) should not be important due to the fact that the brightness temperature to be measured is smooth, and the measured visibility function has large values for small baselines (signals of antennas close to each other) and almost zero for large ones.

For a given set of antennas, the procedure described in the previous section is carried out. For all baselines formed by those receivers, the calibration parameters $\alpha_{kj}=\theta_{0j}-\theta_{0k}-\theta_{kj}$ and $\tilde{g}_{kj} = g_k g_j g_{kj}$ are obtained for all pairs (k-j). Approximate values of the *individual* separable error terms θ_{0j} , θ_{0k} , g_j , g_k may be computed from the measured parameters only if the non-separable part is neglected. The estimation of this parameters is computed on the following basis:

$$\alpha_{kj} = \hat{\theta}_{0j} - \hat{\theta}_{0k} ; \quad \tilde{g}_{kj} = \hat{g}_k \hat{g}_j \quad (24)$$

Establishing the above relations for all (or a number) of baselines, and using an optimization algorithm, the estimated receiver in-phase and amplitude terms can be retrieved. In the case of the amplitudes, taking logarithms, the following matrix equation is obtained from (24):

$$[y] = [M][x] \quad (25)$$

where $[y]$ is a column vector of $N(N-1)/2$ rows whose elements are the measured values of $\log(\tilde{g}_{kj})$, M is an $[N \times N(N-1)/2]$ sparse matrix having only ones and zeros (two ones per row) and $[x]$ is a column vector on N rows whose elements are

$\log(\hat{g}_k)$. The above equation can be solved using standard minimization procedures to obtain the approximate values of g_k ($k=1 \dots N$).

In the case of the phase equation, the procedure is not as straightforward. The minus sign that appears in the first of equations (24) makes the sparse matrix ill-conditioned. In particular, if a subset of N equations is used so as to have the same number of equations than unknowns, the resultant matrix has null determinant. This is a consequence of the indefiniteness of the reference phase. If one receiver is taken as a reference and assigned a 0 degree error phase, $\hat{\theta}_{01} = 0$, the rest of phase errors are directly estimated from (24): $\hat{\theta}_{0k} = \hat{\theta}_{01} - \alpha_{k1}$. This simple approach, however, may lead to errors between the differences of estimated errors and the true value given by (24). A least-square optimization algorithm is a better choice, still maintaining one of the receivers at zero phase error.

2. Phase/amplitude tracking: overlapping sources

If each subset of N receivers driven by a noise source has its own reference phase, the phase calibration of the visibilities formed by receivers from two different sets will have a constant error equal to the phase difference between the reference receivers. Phase tracking from one subset to another is possible using overlapping sources⁷. In this approach, all receivers belong to two subsets and are in fact calibrated twice, using two different noise sources. The phase error assigned to the receiver used as reference in the second subset is taken as the error term computed from the calibration of the same receiver in the previous subset. This allows all phase errors to be referenced to one single receiver, at the center of the array.

A similar procedure from the amplitude terms is used to compute the equivalent temperature of the different noise sources in terms of the one corresponding to the central antennas. In this way, there is only need to know this calibration source.

Assume a baseline (k - j) that has been calibrated using the known noise source T_{c1} . The total amplitude term $\tilde{g}_{kj}^c(T_{c1})$ has been obtained (22). If the same baseline is now driven by the noise source T_{c2} , an equation similar to (18) gives:

$$\tilde{g}_{kj}^c(T_{c2}) = \tilde{g}_{kj}^c(T_{c1}) \frac{T_{c2}}{T_{c1}} \sqrt{Y'_k Y'_j} \quad (26)$$

where

$$Y'_k = \frac{N_{c_k}(T_{c2})}{N_{c_k}(T_{c1})} \quad (27)$$

is the ratio of detected output powers when the two input noise sources are sequentially connected. Since the terms Y' are known from measurements and also the $\tilde{g}_{kj}^c$ ones, from (26), it is possible to retrieve T_{c2} from T_{c1} . Additionally, the overlapping approach allows to apply baseline calibration to a larger number of baselines in the same arm, improving the overall calibration of the instrument.

5. SIMULATION RESULTS

The better performance of the improved Distributed Noise Injection (i-DNI) method over the Distributed Noise Injection (DNI) one is only apparent when non-separable amplitude and/or phase errors are present. Since the central baselines are only a few percent of the total ones, the average rms values of the in-phase and quadrature errors, and the estimated receivers' noise temperature remain approximately constant. However, since these few baselines are the most important ones -the ones with the largest amplitudes- the improvement on the radiometric resolution is much more important. To assess this improvement, simulations have been performed for a Y-array with 15 antennas per arm spaced 0.89λ , with the parameters given in Table 1.

Table 1.- Parameters used in the simulation of the performance of DNI and i-DNI methods

$T_A = 200 \text{ K (Earth)}$	$T_R = 80 \text{ K}$	$T_N = 300 \text{ K}$
$T_A = 0 \text{ K (sky)}$	$\sigma_{TR} = 10 \text{ K}$	$\sigma_{TN} = 25 \text{ K}$
$\sigma_{00} = 20^\circ$	$\sigma_{0q} = 5^\circ$	$S/N = 45 \text{ dB}$ (100 measurements of 0.3 s)

1. Non-separable phase and amplitude errors randomly added to the visibility samples

Figure 3 and Figure 4 compare the performance of the distributed noise injection method⁷ (DNI) with the improved distributed noise injection (i-DNI) method presented in this paper when non-separable errors are randomly added to the visibility samples. The standard deviation of the added errors has been taken as parameter. From them it can be stated that

the i-DNI method one improves the radiometric resolution by a factor in the range 2.4, for non-separable phase errors, and 3.4, for non-separable amplitude errors.

2. Phase/amplitude errors generated by numerical computation with non-ideal filters

Simulations have also been performed by including tolerances in the filters' parameters and computing the fringe-washing function at $\tau=0$ from (3). The non-separable amplitude and phase errors, along with separable in-phase errors come directly from (4). The computation has been made on the basis of the breadboard developed by Matra Marconi Space in the frame of an ESA contract⁸. It includes the following hardware for each receiver: a spurious reject filter, an image rejection filter, a low-pass filter after I/Q demodulation, and a DC block before the samplers to remove possible offset terms.

Table 2 summarizes the standard deviations of the parameters involved in the simulations along with the radiometric resolution achieved by both methods. It should be pointed out the strict filters tolerances required to satisfy the 0.5 K radiometric resolution, even for the i-DNI method.

Figure 3.- Impact non-separable phase errors on the radiometric resolution, in-phase, quadrature and receivers noise temperature for the DNI method ($\Delta T1$) and the i-DNI method ($\Delta T2$).

Figure 4.- Impact non-separable amplitude errors on the radiometric resolution, in-phase, quadrature and receivers noise temperature for the DNI method ($\Delta T1$) and the i-DNI method ($\Delta T2$).

Table 2 Tolerances considered in the simulated filters and radiometric resolution

Filter Parameter	Tolerance (standard deviation)				
Spurious filter bandwidth (MHz)	3.01	2.39	2.08	1.80	1.53
Image filter bandwidth (MHz)	2.21	1.97	1.66	1.48	1.34
LP filter bandwidth (MHz)	0.36	0.28	0.24	0.21	0.18
HP filter bandwidth (MHz) (DC block)	0.34	0.29	0.26	0.23	0.19
Spurious filter central frequency (MHz)	2.52	2.01	1.73	1.56	1.41
Image filter central frequency (MHz)	1.08	0.85	0.72	0.64	0.59
Spurious filter ripple (dB)	0.35	0.20	0.14	0.11	0.08
Image filter ripple (dB)	0.70	0.44	0.30	0.22	0.17
LP filter ripple (dB)	0.10	0.07	0.05	0.04	0.03
Radiometric resolution [K]					
DNI method. $\Delta T1$	5.91	1.92	1.60	1.46	0.94
Improved DNI method. $\Delta T2$	3.36	0.76	0.62	0.51	0.35

6. CONCLUSIONS

A method for receiver calibration of large aperture synthesis radiometers using correlated noise injection has been presented. General equations for complete baseline-based calibration have been derived. They are the basis of a full receiver calibration if a centralized noise distribution scheme is used. For large systems, in which a large distribution network may not be feasible, an improved version of the distributed noise injection method has been devised. It combines the full baseline calibration for receivers sharing the same noise source and calibration of only separable errors for baselines formed by receivers of different sets.

Simulations show that the improvement on radiometer resolution when 2 degree non-separable phase error and 3% non-separable amplitude error are introduced ranges from 2 to 3.9. Furthermore, for the same filter tolerances, the radiometer resolution is 1.46K for the previous distributed network approach⁷ and 0.51K for the one presented here.

The method improves, thus, the radiometric resolution. If the radiometric resolution is set to remain constant, the improvement can be used to relax filters' tolerances, which are very strict anyway.

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