

Juno Waves high frequency antenna properties

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Abstract

The Waves instrument aboard Juno is a sophisticated radio astronomy observatory investigating Jupiter's auroral radio emissions and plasma wave interactions. Waves records electrical field properties using two monopole antennas, which are connected to form a dipole. The receiving properties of the Waves dipole changes quite remarkably over the instruments frequency range from near DC to 40 MHz. In this contribution we outline Waves' electrical sensor properties above the quasi-static frequency range and provide detailed directivity pattern and insertion loss figures of the instrument for science application and data analysis.

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Key Points:

- High frequency properties of the Waves instrument aboard the Juno spacecraft
- Waves directivity takes arbitrary shapes over receive spectrum
- Waves foot point impedance introduces significant discontinuity loss

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11 **Abstract**

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 13 tory investigating Jupiter’s auroral radio emissions and plasma wave interactions. Waves
 14 records electrical field properties using two monopole antennas, which are connected to
 15 form a dipole. The receiving properties of the Waves dipole changes quite remarkably
 16 over the instrument’s frequency range from near DC to 40 MHz. In this contribution we
 17 outline Waves’ electrical sensor properties above the quasi-static frequency range and
 18 provide detailed directivity pattern and insertion loss figures of the instrument for sci-
 19 ence application and data analysis.

20 **1 The Juno mission and the Waves instrument**

21 Juno’s overarching science goal is to study Jupiter’s origin, internal structure, mag-
 22 netic and gravity fields, atmospheric composition and explore the polar magnetospheres.
 23 The gathered data will improve mankind’s understanding of the origin and evolution of
 24 our solar system and provide new insights into the Jovian system’s planetary formation
 25 and evolution. A description of Juno’s science goals and objectives can be found on NASA’s
 26 website, in NASA’s Planetary Data System and in Bolton (2018).

27 The Waves instrument itself is a sophisticated radio astronomy observatory inves-
 28 tigating Jupiter’s auroral radio and plasma wave emissions. Built as a wave spectrom-
 29 eter, Waves maps Jupiter’s electric and magnetic field. Jupiter’s polar magnetosphere
 30 is charted three-dimensionally by Waves for the first time.

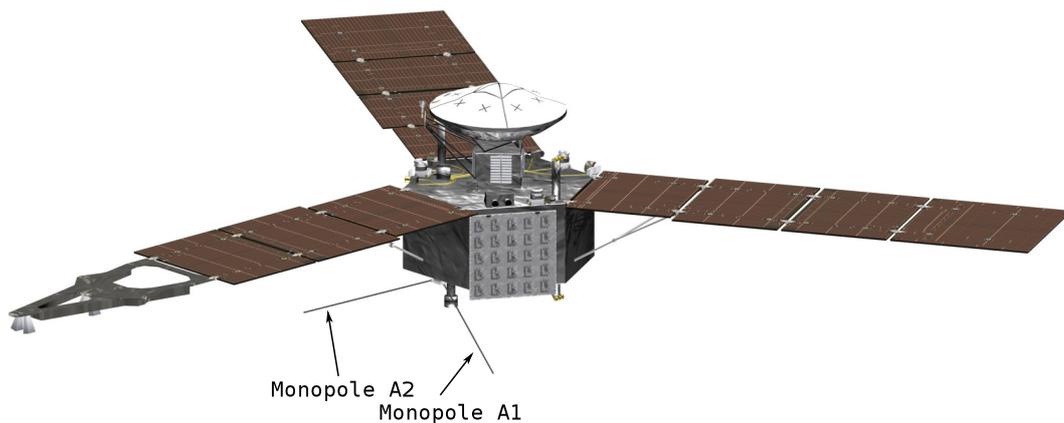


Figure 1. Artist rendering of the Juno spacecraft (©NASA/JPL-Caltech), outlining the Waves monopoles A1 and A2 operating as a dipole.

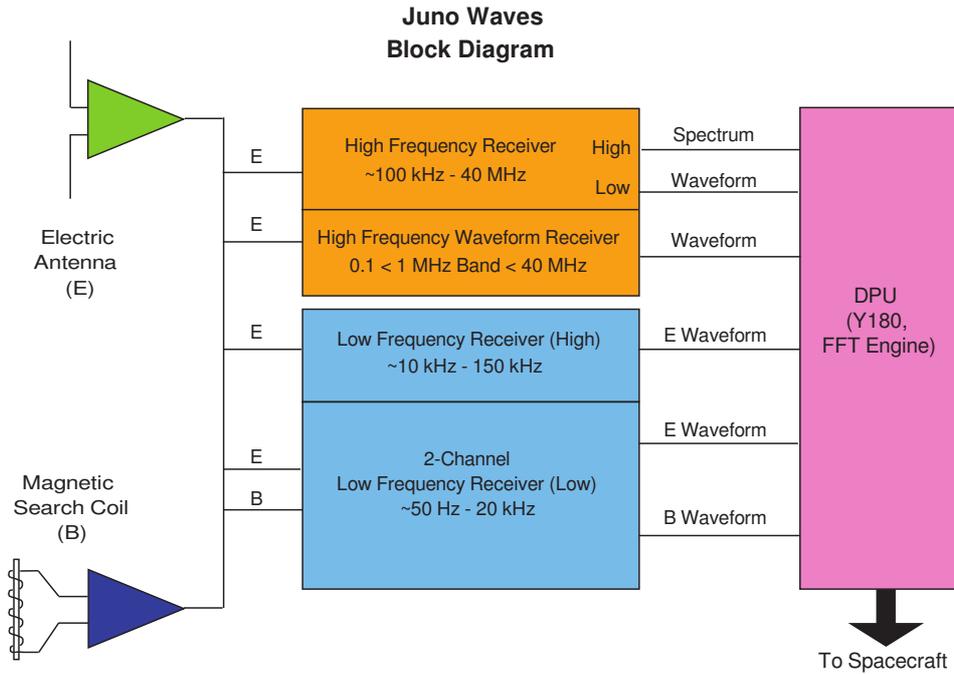


Figure 2. Waves instrument block diagram

31 Waves records electrical field properties using two monopole antennas (Figure 1),
 32 operating as a dipole, and one magnetic search coil, as sensors. The principle of such an
 33 instrument has been already explained many times, for example by Gurnett (1998) and
 34 in detail for Waves by Kurth et al. (2017). The instrument is in principle a spectrum and
 35 waveform receiver as can be seen in block diagram in Figure 2. Basically it consists of
 36 pre-amplifiers, filters, receivers for “low” and “high” frequency bands and signal process-
 37 ing units. The instrument is able to record and process electric field strength from 50 Hz
 38 up to approximately 40 MHz and magnetic fields from 50 Hz up to 20 kHz. The dynamic
 39 range capabilities are especially high due to the challenging demands between low field
 40 strength regions in “empty” space and high field strength regions in Jovian source re-
 41 gions, spanning at least 80 dB. A detailed description of the instrument’s receiver states
 42 and performance can be found in Kurth et al. (2017).

43 Since the spin stabilized S/C is normally pointing with its high gain antenna to Earth,
 44 the Waves dipole is presented to Jupiter and its radio source regions from different an-
 45 gles at any time. Results in section 4 show that for higher frequencies data recorded by
 46 Waves is highly dependent on the angular arrival direction of the incident wave. The di-
 47 rectivity performance of the Waves dipole changes quite remarkably over the instrument’s

48 frequency range from near DC and quasi-static up to 40 MHz. Above the quasi-static
 49 range the wavelength of any incident wave has a large impact on the surface current den-
 50 sity on the spacecraft leading to individual directivity pattern per frequency. When fea-
 51 tures on the S/C are in the geometric lengths of e.g. a quarter wave length then the re-
 52 sulting directivity pattern can exhibit pronounced features such as areas of high of “gain”.

53 In the absence of a multi-axial antenna system, source and direction finding inves-
 54 tigation have to rely on rotating dipole techniques, which are also directly benefiting from
 55 detailed antenna description. To accurately assess a wave phenomena recorded by Waves,
 56 it is necessary to acquire Juno’s spatial location and orientation for selected points in
 57 time, and look-up the dipoles antenna directivity, for the desired direction and frequency,
 58 in the provided 3D radiation tables. Additionally it is necessary to look-up Waves’ impedance
 59 mismatch from the insertion loss table at the desired frequency and factor both into the
 60 recorded spectra as archived in the NASA Planetary Data System.

61 2 Model configuration and numerical computations

62 The S/C model and computational grid used to calculate the results presented in
 63 this contribution have already been described in detail in the companion paper (Sampl
 64 et al., 2016) covering the quasi-static part, i.e. the frequency range, where the observed
 65 wavelength is larger than the antennas geometrical dimensions. The applied model is di-
 66 rectly derived from the S/C manufacturer’s CAD and comprises approximately 12,500
 67 triangle mesh elements.

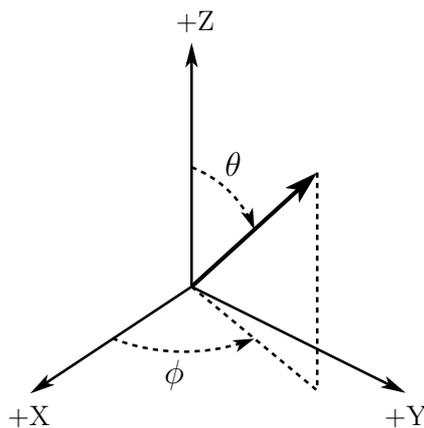


Figure 3. Definition of spherical coordinates θ (co-latitude) and ϕ (azimuth) in the spacecraft-fixed reference frame as used for the representation of antenna axes.

68 The full-wave solvers FEKO and Concept-II are used to solve the underlying in-
 69 tegral equation (c.f. equation (3) in Sampl et al. (2016)) for finding the spacecraft’s sur-
 70 face current distribution. From the current distribution the effective length of the dipole
 71 can be calculated, as well as other useful parameters such as the antennas radiation pat-
 72 tern and the terminal impedance. Both computational solvers yield identical results within
 73 the given digits, leaving any error marginalized. In the following only results from FEKO
 74 are presented and used in further post processing.

75 Figure 3 shows that the z-axis of the spacecraft coordinate system is co-aligned with
 76 the high gain antenna axis. The spacecraft x-axis is in the direction of the solar panel
 77 which includes the MAG boom at the end. The y-axis completes a right-hand orthog-
 78 onal system.

79 **3 Waves dipole quasi-static characteristics**

80 Sensor characteristics of electrically short dipole antennas are typically expressed
 81 in the present radio astronomy context using the effective length vector and the asso-
 82 ciated impedance matrix. The description of the antenna properties in this context in
 83 explained in detail by Macher et al. (2007); Macher (2014), where the antennas are treated
 84 as multi-port scatterer. The Waves’ electric sensor results for the quasi-static frequency
 85 range, up to 4–5 MHz, are already published in a companion paper (Sampl et al., 2016).
 86 The applied spherical spacecraft coordinate system definition is shown in Figure 3, and
 87 is identical as in Sampl et al. (2016).

88 For the sake of completeness we repeat here the results from Sampl et al. (2016)
 89 and Kurth et al. (2017), which include the base capacitance of the receiver hardware and
 90 stray capacitance from surrounding spacecraft structures.

$$91 \quad \mathbf{h} = \frac{C_A}{C_A + C_b} \mathbf{h}^o \quad (1)$$

92 Kurth et al. (2017) calculates the true length of the effective dipole in (1) result-
 93 ing in an overall effective length of $\mathbf{h} = 0.46$ m as opposed to the open port effective
 94 length of $\mathbf{h}^o = 1.46$ m. In (1) the dipole antenna capacitance of $C_A = 14.69$ pF is ap-
 95 plied together with a rounded base capacitance of $C_b = 32$ pF. The base capacitance
 96 is assumed to be the measured per port pre-amplifier input capacitance of $C_b = 22.15$ pF

97 plus an additional 10 pF for stray capacitances of the antenna foot points' structural as-
98 sembly and possible cable features.

99 Due to the findings in this paper, we conclude that the assumptions made for the
100 applied voltage divider in Kurth et al. (2017) are rather pessimistic. In Kurth et al. (2017)
101 the antenna foot points' surrounding structure is added to the base capacitance (addi-
102 tional 10 pF), despite the fact that the surrounding structure is already an integral part
103 of the antenna capacitance, calculated by the applied full-wave solver. Without the ad-
104 ditional 10 pF and calculating with the monopole capacitances of 28.50 pF, the total ca-
105 pacitance ratio stays at $C_A/(C_A + C_b) = 28.50/(22.15 + 28.50) = 0.562$ or 4.99 dB.
106 This results in an effective length of about $\mathbf{h} = 0.82$ m when calculating with the open
107 port effective length of $\mathbf{h}^o = 1.46$ m. Calculated with the geometric antenna length of
108 2.41 m, the overall loss is 9.36 dB.

109 4 Waves directional performance

110 Previously published results (Sampl et al., 2016) discussed only the lower end of
 111 the receivers spectral capabilities. Since the instrument records also observations above
 112 some MHz, we now look at the reception properties of the antennas up to their upper
 113 end of 40 MHz. The analysis of the frequency range above the quasi-static regime is im-
 114 portant, because in-flight calibration which is well applicable for the determination of
 115 the quasi-static effective length vectors, is practically infeasible (Cecconi & Zarka, 2005;
 116 Vogl et al., 2004)) for frequencies above some MHz. The reason is that a large number
 117 of unknowns would have to be determined, since the effective length vector is a complex-
 118 valued direction-dependent quantity above the quasi-static frequency range and can not
 119 be simplified anymore as defined by Macher et al. (2007); Macher (2014) and applied for
 120 Waves as in Sampl et al. (2016).

121 In the following the choice for characterizing the Waves dipole is shifted from the
 122 effective length vector, where the angle of arrival was disregarded in the quasi-static regime,
 123 to color coded 3D radiation pattern representing antenna directivity D , which is propor-
 124 tional to the antenna gain G . Such a representation requires the calculation of individ-
 125 ual factors of gain or attenuation for discrete angles of wave incidence in the whole spa-
 126 tial domain.

127 Nevertheless, even above the quasi-static regime it is possible to characterize the
 128 antenna properties (in particular for the study of wave polarization) by effective length
 129 vectors. For non-quasi-static applications the representation by Sinclair (1950) is usu-
 130 ally preferred which defines the effective length vector \mathbf{h}^S normal to the direction of wave
 131 incidence. The relation between \mathbf{h}^S and the effective length vector definition \mathbf{h} which
 132 becomes direction-independent in the quasi-static limit (often used in low-frequency ra-
 133 dio astronomy) is $\mathbf{h}^S = \mathbf{e}_r \times (\mathbf{h} \times \mathbf{e}_r)$ (Macher et al., 2007; Macher, 2014).

134 Further, textbooks (Balanis, 2005) and literature (Trainotti & Figueroa, 2010) de-
 135 tail how to transform between antenna directivity D , antenna effective area A (the re-
 136 ceiving cross section) and the effective length vector \mathbf{h}^S at a given direction of incidence
 137 \mathbf{e}_r . Thus, we have the relation

$$138 \quad D = \frac{G}{\eta} = \frac{\pi \hat{\zeta}_0}{\eta \lambda^2 R_A} |\mathbf{h}^S|^2 = \frac{\pi \hat{\zeta}_0}{\eta \lambda^2 R_A} |\mathbf{e}_r \times \mathbf{h}|^2 \quad (2)$$

Table 1. The maximum receiving directivity D_{RM} of the dipole is stable over the quasi-static frequency range. Via the radiation resistance R_A and the dipole the open port effective length \mathbf{h}° , D_{RM} can be calculated via (2), when no further losses are involved.

freq (kHz)	Rheometry		Simulation		
	\mathbf{h}° (m)	\mathbf{h}° (m)	D_{RM}	D_{RM} (dBi)	R_A (Ω)
300	1.60	1.450	1.50	1.761	1.662E-03
1000		1.399	1.50	1.762	1.718E-02
2000		1.497	1.49	1.751	7.892E-02

139 where R_A is the antenna radiation resistance, η efficiency and $\zeta_0 = 120\pi \Omega$ the
 140 free space wave impedance. For the present application we can assume $\eta \approx 1$ ($G \approx D$).

141 In Table 1 a number of selected frequencies have been calculated using (2) to show-
 142 case maximum values in the quasi-static regime and interconnect rheometry results (elec-
 143 trolytic tank measurement using a scale model as receiver) in Sampl et al. (2016) with
 144 simulation results in the higher frequencies. The necessary input is available from the
 145 corresponding simulation results (Sampl et al., 2016), which provide the required antenna
 146 radiation resistance R_A . Since modern computational solvers yield the antenna direc-
 147 tivity D directly such a transformation is generally not necessary.

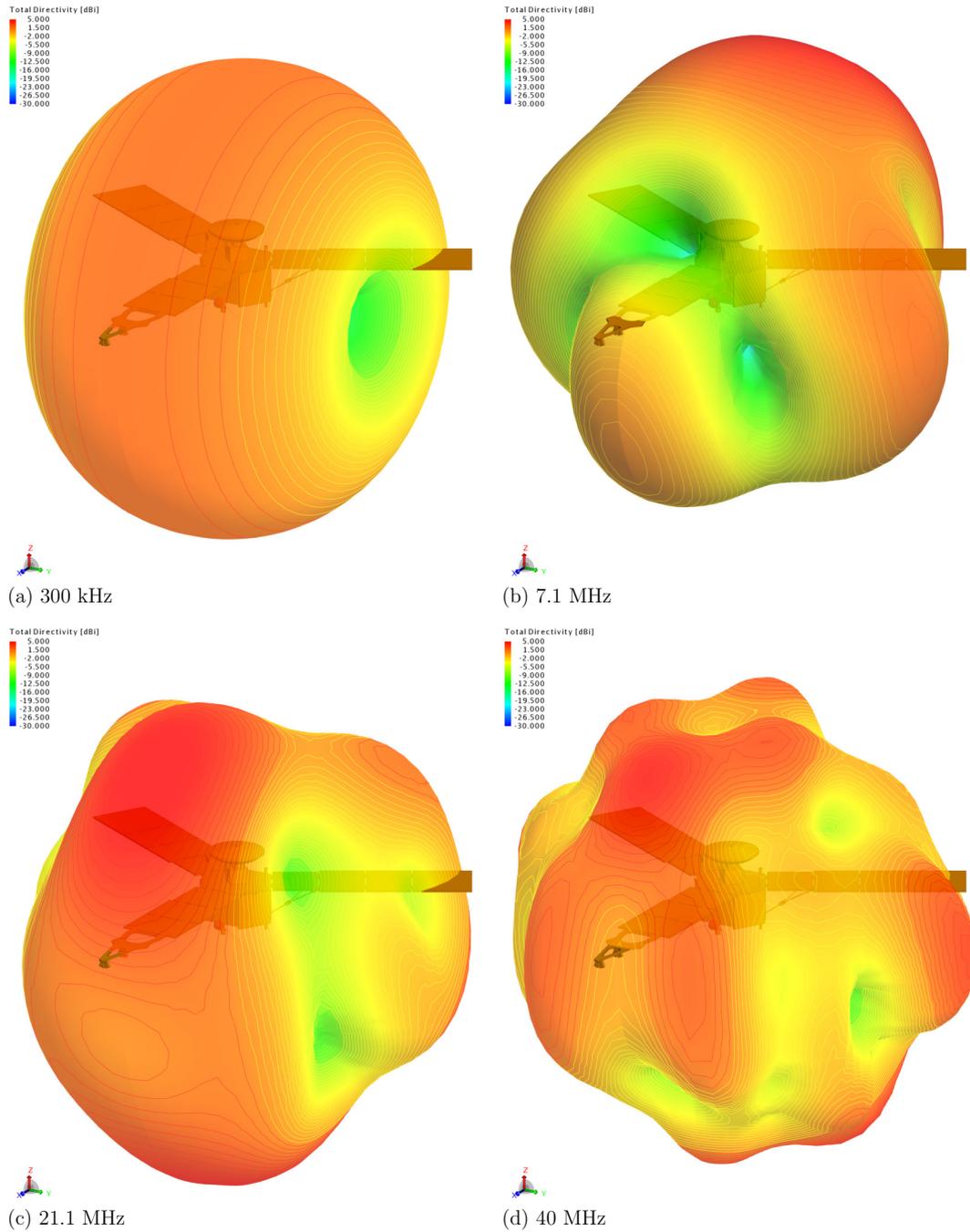


Figure 4. Figures (a)–(b) outline the directional dependence of the Waves dipole radiation characteristic. Below the quasi-static frequency range, in Figure (a), the pattern shows a typical toroidal shape, while above 4–5 MHz the radiation pattern develop multiple elaborate lobes and null directions. For example at 7 MHz a reception maximum in -X direction can be observed, while around 21 MHz the maximum is oriented in the opposite direction +X

148 The directivity patterns shown in Figure 4 provide a first insight into the complex
 149 shapes the directivity of the sensor-dipole assumes at higher frequencies over the spa-
 150 tial domain. All patterns presented are calculated by FEKO with a realistic dipole and
 151 S/C layout (CAD configuration 2, section 6 in Sampl et al. (2016)). The reception prop-
 152 erties of the dipole, while rather stable at frequencies below some 4–5 MHz, are extremely
 153 dependent on the wave incident direction. For frequencies below the quasi-static border,
 154 the angular dependence is typically toroidal shaped, but for frequencies above said bor-
 155 der the directivity pattern takes arbitrary shapes. Obviously this has a significant im-
 156 pact on the instrument’s recorded data. While there are multiple null directions, there
 157 are also unexpectedly high directivity values, for a dipole, to be found.

158 A complete picture of the dipole’s reception pattern is provided in the accompa-
 159 nying data package (Sampl & Macher, 2021), as 2D color coded images (co-latitude over
 160 azimuth in the S/C reference spherical coordinate system), as well as simple to process
 161 raw data tables.

162 **5 Waves antenna input impedance**

163 For further understanding of Waves’ reception properties it is necessary to under-
 164 stand the antenna systems impedance performance. The antennas foot point impedance
 165 allows together with the pre-amplifiers terminal impedance the calculation of the over-
 166 all mismatch loss of the antenna–pre-amplifier interface. Besides the antennas effective
 167 length and directivity, the impedance mismatch at the dipoles foot point has a signif-
 168 icant impact on the recorded power flux.

169 The antenna systems impedance matrices are readily available from the 3D full wave
 170 solver runs as already mentioned in the previous section. The solvers voltage excitation
 171 together with the antennas terminal current conveniently yields the complex antenna foot
 172 point impedance. These results are usable directly as input for further calculations in
 173 section 6.

174 These results also point out a few interesting features of Waves antenna system.
 175 It can be observed that several pseudo-resonances are induced by the geometric prop-
 176 erties of the spacecraft. It can be clearly seen from Figure 5 that the resonant behav-
 177 ior of certain geometric features of the spacecraft influence the monopoles and to some
 178 certain extent the dipole. The solar panels, which are approximately 9 m long are account-

179 able for a barely visible suppression at approximately 7.5 MHz, mimicking a resonant be-
180 havior at 6 MHz. This suppression stems from the fact that, in transmission mode, cur-
181 rents on the solar panels are oriented opposite to the currents on the dipole at about 7.5 MHz.
182 In contrast to that, at 21 MHz the induced currents on the solar panels are oriented in
183 the same way as on the dipole, being responsible for a local resonant effect at around
184 21 MHz. The lowest proper dipole resonance is usually slightly above the frequency where
185 the imaginary part of the impedance crosses zero, which is found at 26 MHz; similarly,
186 the monopoles show their lowest resonance at 28 MHz.

187 The immediate effects of these resonances can be seen not only in the impedance
188 plots but also in the directivity patterns in Figure 4.

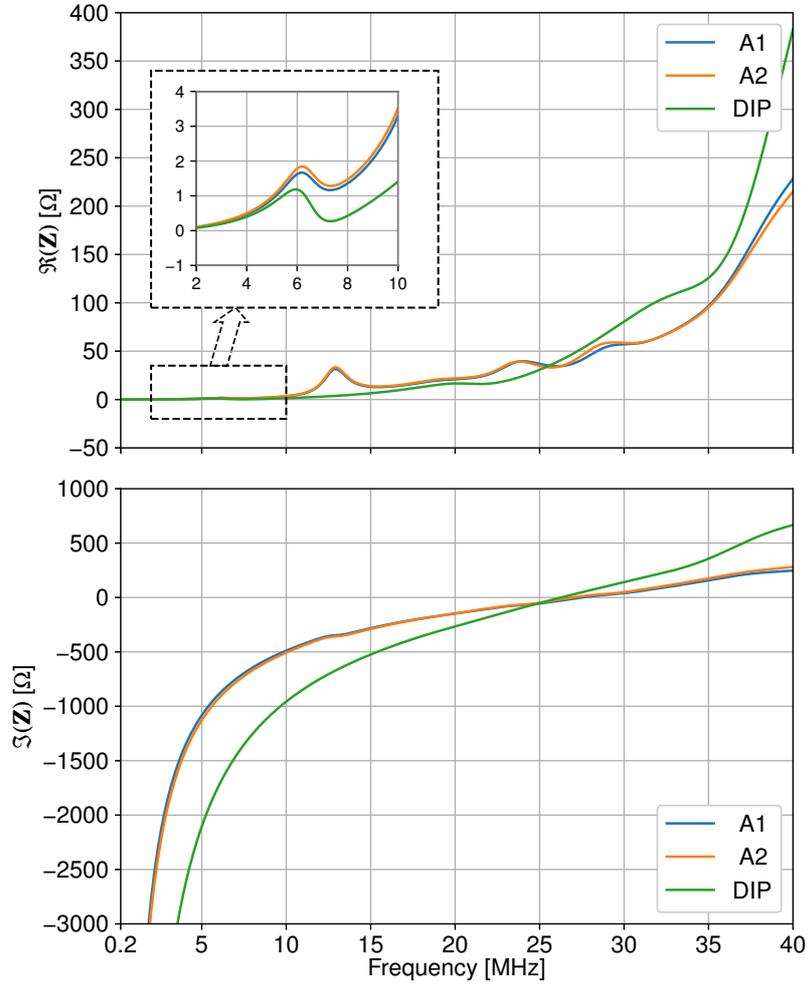


Figure 5. Self impedances of the monopoles A1, A2 and in dipole (DIP) configuration. The shown quantities are the diagonal elements of the impedance matrix as a function of frequency. The upper panel contains the real parts, lower panel the imaginary parts. The curves for the monopoles are nearly identical, which is due to the high symmetry of their deployment on the satellite. The inset in the upper panel outlines pseudo-resonances of the antenna system slightly above the quasi-static area

189 6 Impedance mismatch loss at the antenna foot point

190 Since the observation range of Waves extends from near DC to 40 MHz, it is nec-
 191 cessary to examine the impedance matching of the antenna and the input amplifier cir-
 192 cuit over the sensors frequency range. Such a broadband case poses a substantial chal-
 193 lenge for the instrument designers to optimize the antenna–receiver impedance match-
 194 ing and the resulting power or voltage transfer.

195 In case of the quasi-static range, the discontinuity loss at the antenna foot point
 196 is calculated by applying a voltage divider term as in (1). Above the quasi-static bor-
 197 der, the dipole foot point impedance is generally not purely capacitive anymore, so it
 198 cannot be represented in the form $Y = j\omega C$ with a real capacitance C .

199 Therefore we will outline two different approaches to calculate the losses at the an-
 200 tenna foot point. The first approach, derived in section 6.1, is applying a transfer ma-
 201 trix description (Macher, 2014) with the dipole arms treated as individual monopoles
 202 A1 and A2, and the measured single-ended input impedances of the pre-amplifier used
 203 as load impedances. The second approach, described in section 6.2, is by calculating the
 204 differential-mode input impedance of the pre-amplifier from single-ended measurements,
 205 as in Carrasco et al. (2012); Bockelman and Eisenstadt (1995) and applying it to the Waves
 206 dipole as load impedances.

207 The input impedance of the Waves instrument pre-amplifier was measured on a flight-
 208 like qualification unit using a state of the art network analyzer. Wave’s pre-amplifier in-
 209 cludes an automatic attenuation capability to increase the dynamic range of the instru-
 210 ment and reduce saturation effects in the receiver. The corresponding data package con-
 211 sists of data sets for each attenuator state “ON” and “OFF”. Touchstone format S-parameter
 212 files covering the relevant frequency range were provided as input for this investigation.

213 Both approaches yield similar impedance matching losses, within an acceptable ac-
 214 curacy, taking the applied assumptions and simplifications into account, e.g. disregard-
 215 ing the common mode part in section 6.2. The results, presented in Figure 6 show nicely
 216 the frequency dependence of the dipole—pre-amplifier interface. The first observation
 217 which can be made is that the lower panel in Figure 6 confirms the calculation as in sec-
 218 tion 3. The occurring loss is at quasi-static frequencies at 5 dB and then gradually drop-
 219 ping to a zero around 27 MHz. In this area of the spectrum two effect can be observed.

220 First the real part of the monopole input impedance and the pre-amplifier input impedance
221 is equal at around 50Ω . Secondly the monopoles resonance frequency is at 27.5 MHz.
222 This circumstances seem to result in a favorable transmission behavior for higher fre-
223 quencies. Above 27.5 MHz the loss figure turns into negative suggesting a gain situation,
224 with a minimum of -4 dB (or maximum, depending on the perspective) at 35 MHz as per
225 calculation in section 6.1 and -6 dB in section 6.2. This is also the point where the two
226 applied calculation methods show the largest difference of 2 dB. Above 35 MHz the curve
227 swings back to cross zero at 38 MHz ending in 4 dB loss at the upper end of the instru-
228 ments spectral capabilities. One possible explanation for the occurring negative losses
229 in this configuration is the Ferranti effect. Above the antenna resonance at 27.5 MHz the
230 monopoles impedance turns inductive and together with the capacitive pre-amplifier in-
231 put, reactive gain could lead to an increased voltage at the receiving pre-amplifier.

232 The two attenuator states “ON” and “OFF” show the same performance, within
233 the given digits. The results therefore suggest that data post-processing does not need
234 to take the attenuator state into account.

235 The data package (Sampl & Macher, 2021), which is part of this contribution, con-
236 tains tables with the insertion loss in 100 kHz steps, from 200 kHz up to 40 MHz, for both
237 cases: attenuator 'ON' and 'OFF'.

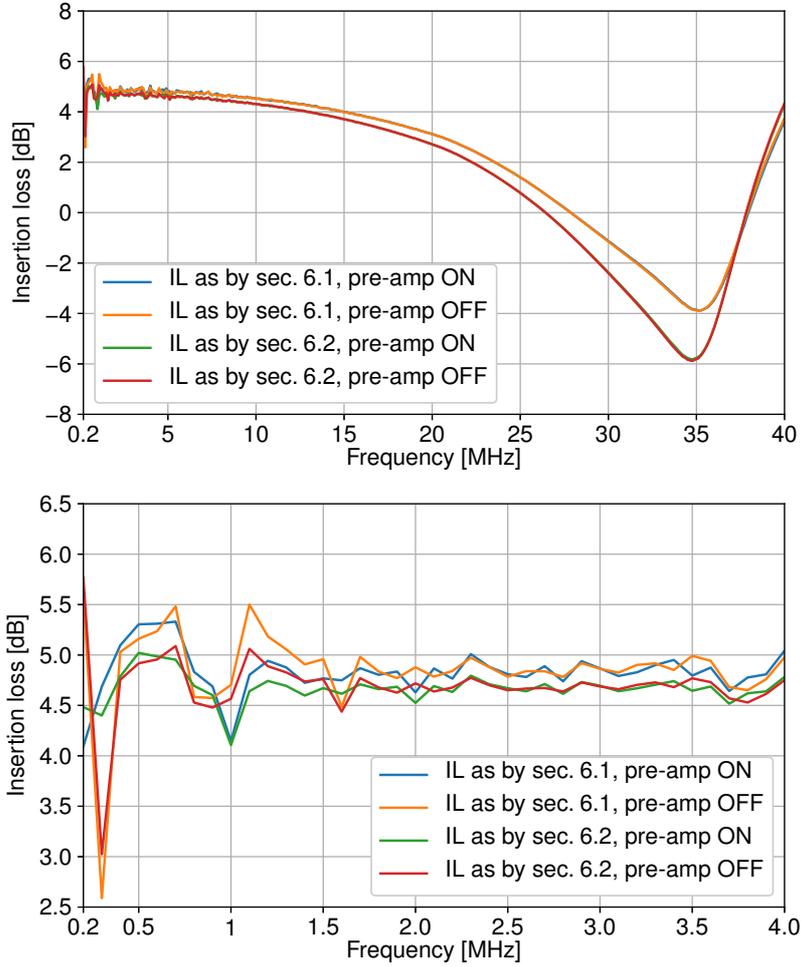


Figure 6. Insertion loss (IL) at the antenna–pre-amplifier interface. The upper panel shows the insertion loss over the full frequency range from near DC to 40 MHz, while the lower panel presents the quasi-static range up to 4 MHz. The lower panel proves that the quasi-static range has a stable performance over frequency; the spikes in the lowest frequency segment can be attributed to the challenging task of performing RF measurements near DC. The upper panel visualizes the frequency dependence of the applicable loss at the investigated interface. For both calculation methods the pre-amplifier state 'ON' and 'OFF' is depicted.

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6.1 Insertion loss by impedance and scattering matrix calculation

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As we already stated in section 6, for the case of the quasi-static range, the discontinuity loss at the antenna foot point is calculated by applying a voltage divider term as in (1). Above the quasi-static border, the dipole foot point impedance is generally not purely capacitive anymore, so it cannot be represented in the form $Y = j\omega C$ with a real capacitance C . It is necessary to resort to the general description by means of a complex admittance or impedance matrix which depends on frequency. In this way we take a non-capacitive load as well as possible cross-talk between the ports of the two antenna arms into account. So we describe the load that the antenna system sees when looking from the feed gaps to the receiver (pre-amplifier) by means of a 2×2 load impedance matrix \mathbf{Z}_L . The effective length vectors $\mathbf{h}^{(1)}$ and $\mathbf{h}^{(2)}$ of the dipole arms A1 and A2, respectively, are subsumed into a transfer matrix $\mathbf{T} = (\mathbf{h}^{(1)}, \mathbf{h}^{(2)})^t$ (Macher, 2014). Here $(\cdot)^t$ means the transpose of (\cdot) , and the quantities refer to the loaded system (situation in actual deployment, i.e. receiver connected to the antenna arms). Similarly, we define $\mathbf{T}^\circ = (\mathbf{h}^{(1)\circ}, \mathbf{h}^{(2)\circ})^t$ for the open ports system (receiver disconnected). The relation between \mathbf{T} and \mathbf{T}° is given by the matrix equation

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$$\mathbf{T} = \mathbf{Q} \cdot \mathbf{T}^\circ \quad (3)$$

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with the 2×2 matrix

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$$\mathbf{Q} = \mathbf{Z}_L \cdot (\mathbf{Z}_L + \mathbf{Z}_A)^{-1} \quad (4)$$

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Finally, the dipole effective length vector can be determined as the difference between the respective vectors of the monopoles (dipole arms): $\mathbf{h}^d = \mathbf{h}^{(2)} - \mathbf{h}^{(1)}$ for the loaded ports, and analogously for the open ports, $\mathbf{h}^{d\circ} = \mathbf{h}^{(2)\circ} - \mathbf{h}^{(1)\circ}$. We note that, in general (if $Z_{11} \neq Z_{22}$ and/or $Z_{12} \neq 0$), \mathbf{h}^d cannot be described as a multiple of $\mathbf{h}^{d\circ}$, since the coordinates of the dipole effective length depends on the open port effective lengths of both monopoles:

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$$\begin{aligned} h_n^d &= T_{2n} - T_{1n} \\ &= \sum_{k=1}^2 (Q_{2k} - Q_{1k}) T_{kn}^\circ, \quad n = 1 \dots 3 \end{aligned} \quad (5)$$

265 However, in the present application we can make the following assumptions: The
 266 two monopole terminals are connected to the pre-amplifier in a well balanced way, that
 267 is, we can assume $Z_{L11} \approx Z_{L22}$, as was confirmed by the results of the network ana-
 268 lyzer measurements. Further, there is a mirror symmetry of the spacecraft–antenna sys-
 269 tem, with the mirror plane running through the middle between the monopole feeds and
 270 the plane normal being parallel to the tip-to-tip vector of the dipole, in consequence $Z_{A11} \approx$
 271 Z_{A22} . This was verified for the quasi-static regime by means of rheometry measurements
 272 and simulations (Sampl et al., 2016). Finally, we can apply reciprocity, $Z_{L12} = Z_{L21}$
 273 and $Z_{A12} = Z_{A21}$. With these simplifications, we find $Q_{21} - Q_{11} = -(Q_{22} - Q_{12})$ and
 274 so \mathbf{h}^d becomes, in fact, a multiple of \mathbf{h}^{do} :

$$\begin{aligned}
 \mathbf{h}^d &= \frac{(Z_{L11} - Z_{L12})(Z_{L11} + Z_{L12} + Z_{A11} + Z_{A12})}{Z_{L11}^2 + 2Z_{L11}Z_{A11} - Z_{L12}^2 - 2Z_{L12}Z_{A12} + Z_{A11}^2 - Z_{A12}^2} \mathbf{h}^{do} \\
 &= \frac{Z_{L11} - Z_{L12}}{Z_{L11} - Z_{L12} + Z_{A11} - Z_{A12}} \mathbf{h}^{do}
 \end{aligned} \tag{6}$$

277 where the denominator of the fraction in the first line is the determinant $\det(\mathbf{Z}_L + \mathbf{Z}_A)$.

278 As already stated, the measurements were conducted as simple single-ended two-
 279 port measurements, S_{11} giving the reflection coefficient of the pre-amplifier input for A1,
 280 S_{22} for A2 and $S_{21} = S_{12}$ being the power transfer between the ports. These param-
 281 eters define the 2×2 scattering matrix \mathbf{S} from which the load impedance matrix can
 282 be determined by Medley (1993)

$$\mathbf{Z}_L = Z_c(\mathbf{1} - \mathbf{S})^{-1} \cdot (\mathbf{1} + \mathbf{S}) \tag{7}$$

284 where Z_c is the characteristic impedance of the connected wave guides and $\mathbf{1}$ denotes the
 285 2×2 identity matrix.

286 The following procedure can be used to determine the gain (2) as function of di-
 287 rection: First, (7) is applied to obtain the load impedance matrix from the measured scat-
 288 tering parameters for each frequency of interest. Next, \mathbf{T}^o and from it \mathbf{h}^{do} are determined
 289 for all directions (and all frequencies) needed to perform the desired $G(\Omega)$ plot. Finally,
 290 formulae (3) and (5) are applied to determine \mathbf{T} and \mathbf{h}^d . Provided the mentioned sym-
 291 metry and reciprocity properties apply, we can omit the determination of the monopole
 292 properties (whole transfer matrix \mathbf{T}) and directly determine \mathbf{h}^{do} , from which \mathbf{h}^d can be

293 determined via (6). Since we measured S -parameters, it is convenient to write (6) in scat-
 294 tering parameter notation. For that purpose we also introduce the scattering matrix \mathbf{S}_A
 295 of the antenna system, regarded as a two-port consisting of the two monopoles and the
 296 spacecraft as a parasitic body. Thus, the relation between \mathbf{Z}_A and \mathbf{S}_A is totally analo-
 297 gous to (7). Substitution into (6) and a cumbersome rearrangement (for which we ap-
 298 plied *Mathematica*) yield

$$299 \quad \mathbf{h}^d = \frac{(S_{L11} - S_{L12} + 1)(S_{A11} - S_{A12} - 1)}{2(S_{L11} - S_{L12})(S_{A11} - S_{A12}) - 2} \mathbf{h}^{do} \quad (8)$$

300 which are used below for the study of the insertion loss.

301 Figure 6 shows the frequency dependence of the insertion loss, i.e. the loss due to
 302 the inclusion of the base impedance (load including pre-amplifier input impedance, ca-
 303 ble and stray capacitance at feed gap which is not taken into account in the antenna model),
 304 defined by

$$305 \quad \text{IL} = -20 \log_{10} \left(\frac{|\mathbf{h}^d|}{|\mathbf{h}^{do}|} \right) \quad (9)$$

306 which can be easily calculated from (6) or (8). In the present context, IL represents the
 307 loss in dB of the voltage autocorrelation $|V|^2$ due to the presence of the base impedance.

308 6.2 Insertion loss from mixed-mode S-parameter calculation

309 Waves' receiver is built to store data in NASA's Planetary Data System from a dipole
 310 antenna. Therefore it seems obvious to not only calculate the impedance mismatch losses
 311 via the individual monopoles, but also via the dipole. As the pre-amplifier measurements
 312 were conducted as simple single-ended two-port measurements, (S_{L11} giving the reflec-
 313 tion coefficient of pre-amplifier input for A1, S_{L22} for A2 and S_{L21}, S_{L12} being the power
 314 transfer between both ports) it was necessary to calculate the differential-mode input
 315 impedance from the given S-parameter file. Normally it is necessary for a full calcula-
 316 tion of the differential input impedance from single-ended measurements to know the mixed-
 317 mode S-parameter of the device, containing differential- and common-mode parameters.
 318 In the case of a balanced and symmetrical device under test, where the common-mode
 319 part is small compared to the differential part, the differential-mode parameters calcula-
 320 tion (Carrasco et al., 2012; Bockelman & Eisenstadt, 1995) simplifies to

$$\Gamma_d = \frac{S_{L11} - |S_L|}{1 - S_{L22}}. \quad (10)$$

With the given measurements, Γ_d , is the differential reflection coefficient of the sought input stage. (10) can be accepted as true, with $|S_L| = S_{L11}S_{L22}$, under the assumption that the input stage of the circuit is symmetrical and balanced ($S_{L11} = S_{L22}$, $S_{L21} = S_{L12}$), and the common-mode part is small compared to the differential part. Fortunately the measurements of the Waves pre-amplifier input stage support the above conditions. By keeping in mind that in the present case $\Gamma_d = S_d$ and by applying bilinear transformation we get the differential pre-amplifier input impedance

$$\mathbf{Z}_d = \frac{Z_c \mathbf{1} + \mathbf{S}_d}{2 \mathbf{1} - \mathbf{S}_d}. \quad (11)$$

From here we can use the same transfer matrix description as in (4), with \mathbf{Z}_{dip} as dipole foot point impedance and \mathbf{Z}_d as load impedance for the dipole

$$\mathbf{Q}_d = \mathbf{Z}_d \cdot (\mathbf{Z}_d + \mathbf{Z}_{\text{dip}})^{-1}. \quad (12)$$

Finally we can calculate the insertion loss with (9) in the same manner as in section 6.1. Due to the absence of multiple monopole arms the transfer matrix \mathbf{T} simplifies accordingly, and we can write:

$$\mathbf{IL} = -20 \log_{10} |\mathbf{Q}_d| \quad (13)$$

In contrast to \mathbf{Q} as applied in section 6.1, the present approach by applying (10), resulting in \mathbf{Q}_d as in (12), assumes that the mutual impedances $Z_{L12} = Z_{L21}$ are negligible in comparison to $Z_{L11} = Z_{L22}$.

7 Conclusions

In this contribution we present the directivity of Waves' receiving dipole and the impedance mismatch loss at the dipole's foot point. We show that above the quasi-static frequency range, the instrument's recorded data figures are drastically affected by the wave incidence direction and the wave length, as observed when compared to the true properties of an incident wave.

346 While the directivity patterns show a typical toroidal shape below the quasi-static
347 border, at higher frequency, arbitrary shapes emerge with several areas of null direction
348 and unexpectedly high values for a dipole (> 5 dBi). For lower frequencies, Waves' dipole
349 impedance mismatch loss calculations' lie at 5 dB, while for higher frequencies losses de-
350 crease up to a minimum of approximately -4 dB around 35 MHz.

351 For scientific data post processing we recommend to apply the results from the cal-
352 culations in section 6.1, motivated by the instruments architecture. Afterall the antenna
353 monopoles are independently connected to an impedance transformation stage and from
354 there to the differential interface of an op-amp. This differs quite from the common build-
355 ing practice of dipole antennas, often being connected to the receiver stage via a balun
356 or a similar transformation stage. It can also be stated from the results that the instru-
357 ments built-in attenuator does not have any noticeable impact on the instruments recorded
358 data and any scientific data post-processing does not need to take the attenuator states
359 into account.

360 Our work provides the necessary correctional data sets for fully calibrated post-
361 processing of both spectral and spatial analysis of Waves' recorded electrical wave data.

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367 The geometrical and computational model used to compute the results of this con-
368 tribution is available as supporting information in (Sampl et al., 2016).

369 The results yielded by this investigation are available as a data package in the Zen-
370 do repository (Sampl & Macher, 2021). It contains files of the Waves' calculated foot
371 point impedance mismatch loss (Data Set 1) and directivity (Data Set 2) over the in-
372 strument's whole frequency range in 100 kHz steps. The Plots to Data Set 2 provide im-
373 ages showing for every 100 kHz step a visual 2D-surface impression of the 3D directiv-
374 ity pattern.

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