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Measurements and Simulations of Overmoded Waveguide Components at 70-118 GHz, 220-330 GHz and 610-720 GHz

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Abstract

For low-loss transmission of the LO in the ALMA Band 6 cartridges, overmoded waveguide may be used. In this paper, we report on the theoretical and measured losses of various sizes of waveguide in the frequency ranges 70-118 GHz, 220-330 GHz and 610-720 GHz.

Introduction

The use of overmoded waveguide for low-loss transmission dates back to the earliest usage of waveguide [1], [2], [3], [4]. This use is now quite standard in circular waveguide [5], [6] and in rectangular waveguide [7], [8], [9], [10], [11]. In this paper, we deal with the transmission of the fundamental mode (TE_{10}) in overmoded rectangular waveguide. This mode is the easiest mode to generate and control in signal sources and components such as couplers, hybrids, etc. as no mode converters are required. This will be used for the transmission of the LO from the final multiplier in the Band 6 multiplier chain, which will be on the 90 K or 20 K stage of the cartridge for thermal loading reasons, and the mixer, which will be on the 4 K stage. Depending on the thermal load conditions, these two components will be up to 200 mm apart. Due to waveguide losses and available/required power levels, losses should be as low as possible.

Theory

Loss, α, of the fundamental (TE_{10}) mode is given by [12]

$$\alpha = \frac{\lambda}{b \cdot \lambda g} \cdot \left(\frac{\pi}{\lambda \cdot \eta \cdot \sigma}\right)^{1/2} \cdot \left[1 + \left(\frac{\lambda g}{\lambda c}\right)^2 \cdot \left(1 + 2 \cdot \frac{b}{a}\right)\right] \cdot 8.676 \, \frac{dB}{m}$$

where λ is the free space wavelength; λg is the waveguide wavelength in waveguide of width, a, and height, b; σ is the conductivity of the wall material and η is the impedance of free space (120*π Ω). Figure 1 shows the theoretical loss for a frequency of 243 GHz as a function of waveguide width, a, for a/b = 2 (standard waveguide). The waveguide for Band 6 (WR3.7) is 0.94 x 0.47 mm (37 x 18.5 mils), for which the loss would be approximately 18 dB/m for copper ($\sigma = 4.00 \cdot 10^7$ S/m), and nearly 117 dB/m for stainless steel ($\sigma = 0.1*10^7$ S/m), which is preferred for its thermal properties. These include the usual factor of 1.3[1] to account for surface roughness effects [13], and Tischer [13] also gives a further factor of 1.135[**] for an "anomalous skin effect" in copper. Using WR10 (2.54 x 1.27 mm, 100 x 50 mils) as the overmoded waveguide, the loss reduces to 3.8 dB/m and 26 dB/m, respectively.

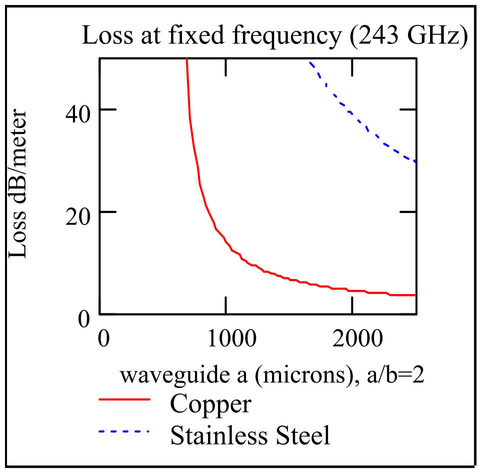

Figure 1

Losses will, of course, be reduced by a factor of approximately 3 to 4 upon cooling copper to about 4 K according to Wollack *et al.* [14], whereas for stainless steel the reduction is only about 25% [15].

Trapped Modes

One problem with using overmoded waveguide systems is the possibility of exciting unwanted modes that can cause resonant losses [14], [16], [17], especially in systems where two tapers from/to fundamental mode waveguide are used. The unwanted modes are reflected at some point in the tapers and are "trapped" causing deep, narrow resonances. The frequency spacing, δf, can be very fine, depending upon the total length between the reflection points, L [9]

$$\frac{\lambda f}{f} = \frac{\lambda}{2L} \quad .$$

The depth of the resonances is given by [10] as

$$\frac{Pmin}{Pmax} = \frac{1 - \frac{(2 \cdot Pc \cdot A)}{(1 + A)}}{1 + \frac{(2 \cdot Pc \cdot A)}{(1 - A)}}$$

where Pmax is the power transmitted through the system away from resonance, Pc is the power converted into the trapped mode (mode conversion), and A is the one-way power transmission of the trapped mode. (Note that equation (2) given in [9] is incorrect.) Calculations are shown in Figure 2.

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[1] Losses in dB are multiplied by these factors.

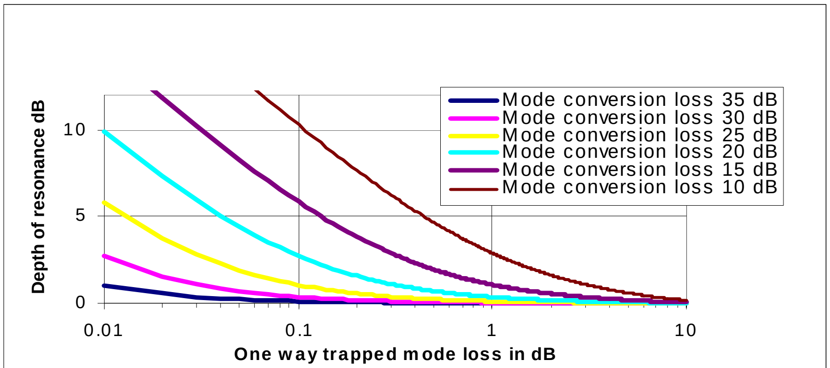

Figure 2

For typical tapers, the mode conversion is -20 dB (or better), with a one-way mode loss of 2 dB, and the resonances are approximately 0.2 dB deep which is negligible. Cooling to 20 K, however, will reduce the one-way loss to about 0.67 dB which will increase the resonance depth to 0.5 dB. Also of concern for ALMA is any change in phase of the LO as it is tuned in frequency across a resonance. Unfortunately, it has not been possible to measure the phase with any of the present measuring systems.

Trapped modes will also be excited by any bends or twists included in the waveguide between the tapers. (H-plane bends are worse than E-plane bends in this regard for oversize waveguide with the usual E-field orientation [9][10]. The opposite is true for "Tall" guide (reciprocal a to b ratio), but "Tall" guide has more resonances.) Mode filters are difficult to use at these frequencies due to the small sizes and the large number of modes (WR10 has 14 modes at 300 GHz, 62 at 700 GHz), all of which can be excited above their cut-off frequencies by any discontinuities.

The width of the transmission resonances also gives an indication of the mode conversion level. [14] gives the conversion as

$$Pc = \frac{\pi}{2} \cdot \frac{\Delta f}{f} \cdot (1 - Rmin)$$

where Rmin is the depth of the resonance, f is the resonant frequency, and Δf is the half-power width of the resonance.

QuickWave and CST Calculations

QuickWave [18] or CST [19] Finite Difference Time Domain EM simulators can be used to analyze such systems, but run into difficulty with such large structures (in wavelengths). As meshes with cell sizes of less than approximately 0.1 wavelengths should be used [20], complete structures cannot be analyzed due to memory limitations and calculation times. Partial structures can be analyzed (tapers, bends, *etc*.) which give some idea of the problems. In runs with bends, trapped modes are clearly seen and the change in their depth can be shown to be a function of material conductivity. QuickWave was used for the following. Figure 3 shows the calculated transmission of a 12.7 mm linear taper from WR3.7 to WR10 with 2.54 mm lengths of waveguide at each end, a 25.4 mm radius (center-line) 90-degree E-plane bend in WR10 and then a similar taper with waveguide sections back to WR3.7 for a perfect conductor. Figure 4 is the same geometry using a conductivity of $0.5*10^7$ S/m (no surface roughness). Figures 5 and 6 show the phase for these two cases, respectively, and show that there are phase effects for deep resonances. For small resonances (< 1 dB), there are no phase effects. Similar effects can be achieved by inserting lossy material in the waveguide to obtain the same one-way path loss, but this can add extra reflections at the front and back surfaces of the material. Figure 7 shows the transmission and reflection of two lossless linear tapers of 7.62 mm length, from WR3.7 to WR10, back-to-back with a section of overmoded waveguide, WR10, of length 2.54 mm between them. Figure 7 also shows that tapers of sufficient length and perfect flange alignment have very little mode conversion. Figure 8 shows the results for two 1.27 mm long lossless tapers.

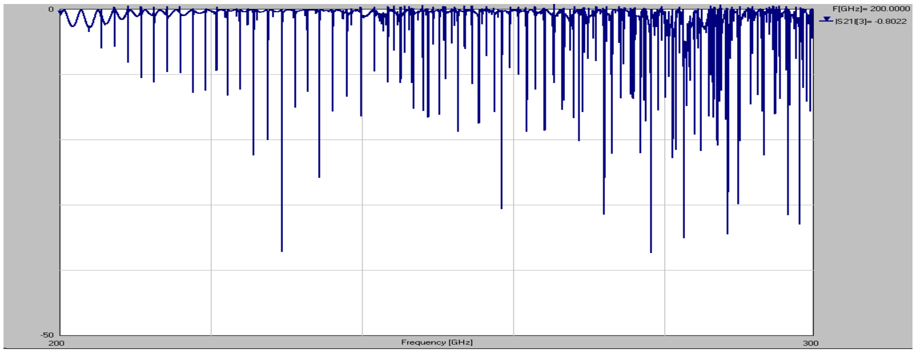

Figure 3. Calculated transmission (0-50 dB, 10 dB per division) as a function of frequency (200-300 GHz, 20 GHz per division) of a 12.7 mm linear taper from WR3.7 to WR10 with 2.54 mm lengths of waveguide at each end, a 25.4 mm radius 90 degree E-plane bend in WR10, and then a similar taper with waveguide sections back down to WR3.7 for a perfect conductor.

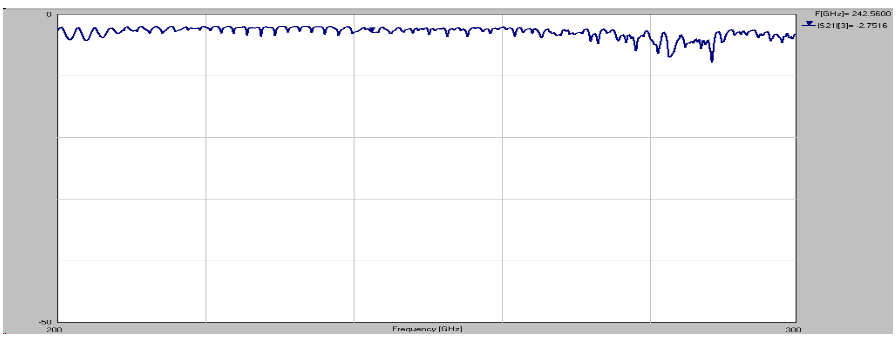

Figure 4. Calculated transmission (0-50 dB, 10 dB per division) as a function of frequency (200-300 GHz, 20 GHz per division) of a 12.7 mm linear taper from WR3.7 to WR10 with 2.54 mm lengths of waveguide at each end, a 25.4 mm radius 90-degree E-plane bend in WR10 and then a similar taper with waveguide sections back down to WR3.7 for a conductivity of $0.5*10^7$ S/m.

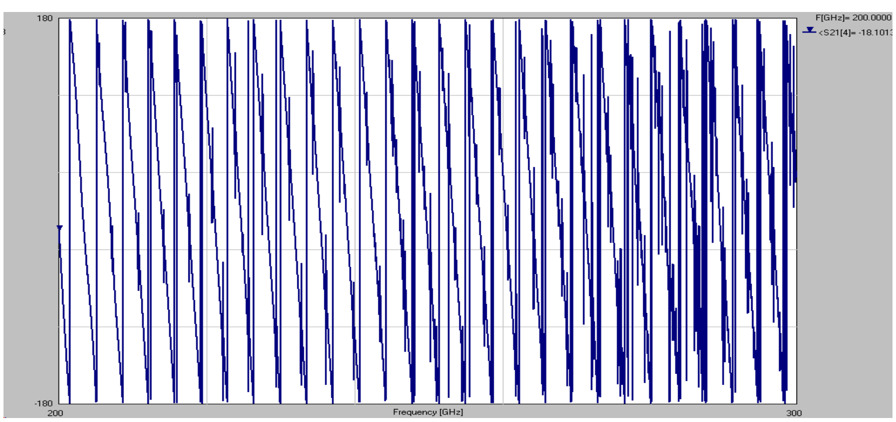

Figure 5. Transmission phase (+180 to –180 degrees) as a function of frequency (200-300 GHz, 20 GHz per division) of a 12.7 mm linear taper from WR3.7 to WR10 with 2.54 mm lengths of waveguide at each end, a 25.4 mm radius 90-degree E-plane bend in WR10 and then a similar taper with waveguide sections back down to WR3.7 for a perfect conductor.

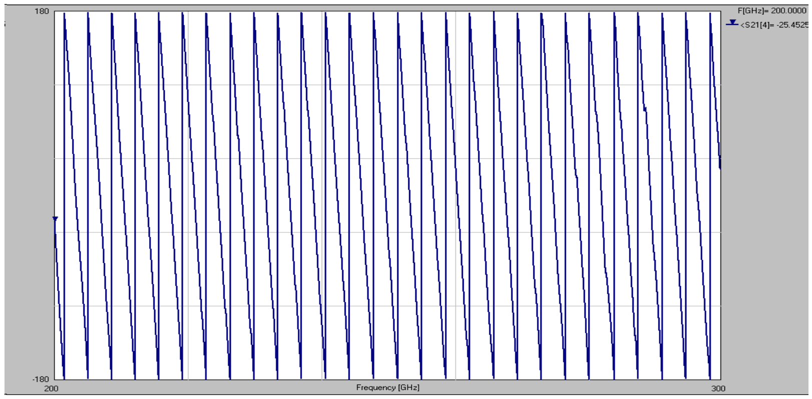

Figure 6. Calculated transmission phase (+180 to –180 degrees) as a function of frequency (200-300 GHz, 20 GHz per division) of a 12.7 mm linear taper from WR3.7 to WR10 with 2.54 mm lengths of waveguide at each end, a 25.4 mm radius 90-degree E-plane bend in WR10 and then a similar taper with waveguide sections back down to WR3.7 for a conductivity of $0.5*10^7$ S/m.

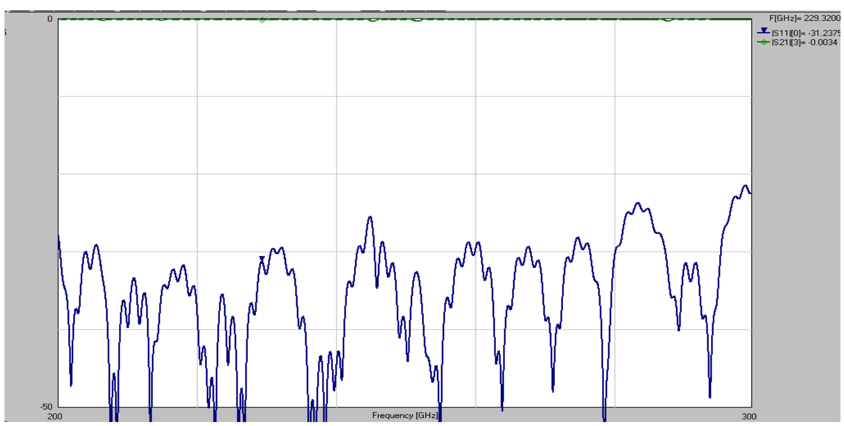

Figure 7. S11 and S21 for two lossless WR3.5 to WR10 tapers (7.62 mm long) back-to-back.

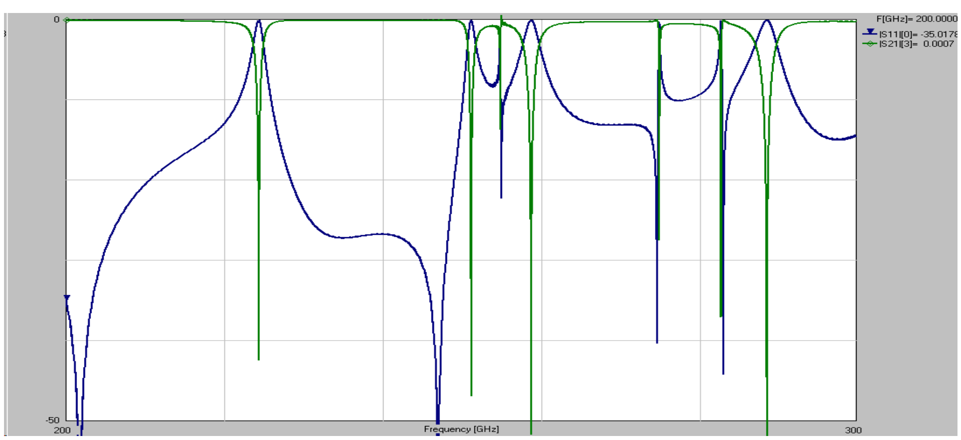

Figure 8. S11 and S21 for two lossless tapers WR3.5 to WR10 (1.27 mm long) back-to-back.

Scale Model Measurements at 69-118 GHz (Scale factor 7.14 from Band 9, 600-700 GHz)

In vector network analyzer (VNA) measurements on a scale model, the distance between the VNA (HP8510) heads is limited by the cable lengths to about 500 mm (equivalent to 70 mm at Band 9), and mounting bends and twists to simulate the final waveguide run may not be possible. Figure 9 shows the measured transmission of two tapers from WR10 to WR71.4, each 100 mm long placed back- to-back, which clearly shows resonances (theoretical loss 0.18 dB for aluminum, roughness factor included). In the scale model, aluminum was used for ease of manufacture, but the surface losses do not scale as required to simulate the Band 6 waveguide (increasing as the square root of the frequency), so that the resonances are much deeper than they would be in the final structure. The frequency scans were made with 801 points. Figure 10 shows the transmission for two tapers 300 mm long (theoretical loss 0.52 dB for aluminum, including roughness factor). Figure 11 shows the loss for a straight section of WR71.4 of length 190 mm placed in between two tapers (divided by the transmission of the tapers alone). This figure has both positive and negative resonances due to the shift in frequency of the resonances as the length is changed and the subtraction of the taper loss. The loss of the straight section matches well with the theoretical, including roughness factor (also shown in Figure 11).

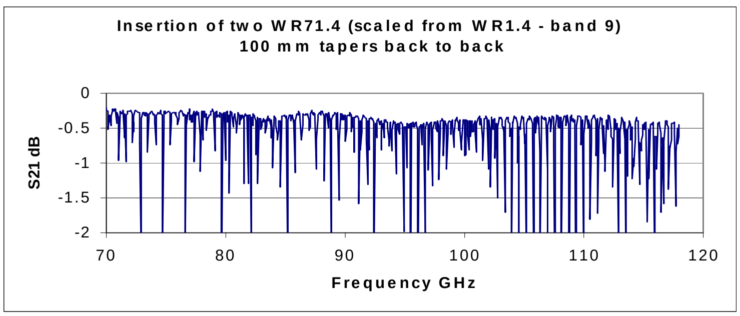

Figure 9

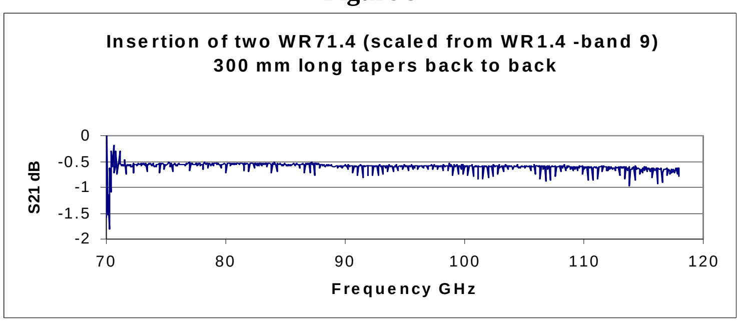

Figure 10

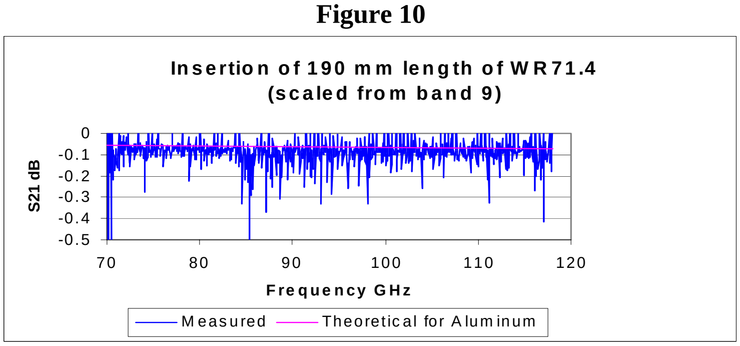

Figure 11

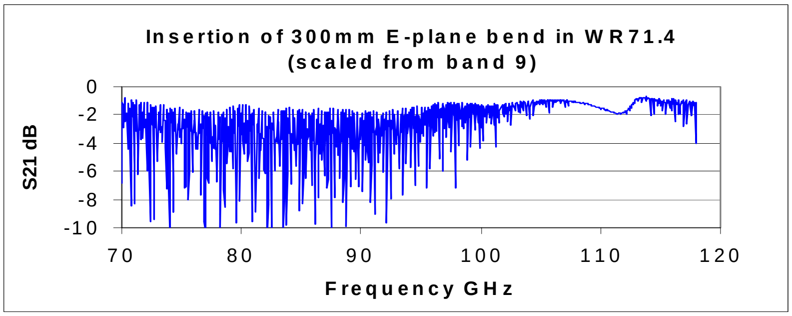

Figure 12

Figure 12 gives the insertion of a 300 mm long, (center-line) 90-degree, E-plane bend. Figure 13 gives the insertion of a 300 mm long (center-line), 90-degree, H-plane bend. (In both figures, the transmission of the tapers is subtracted.)

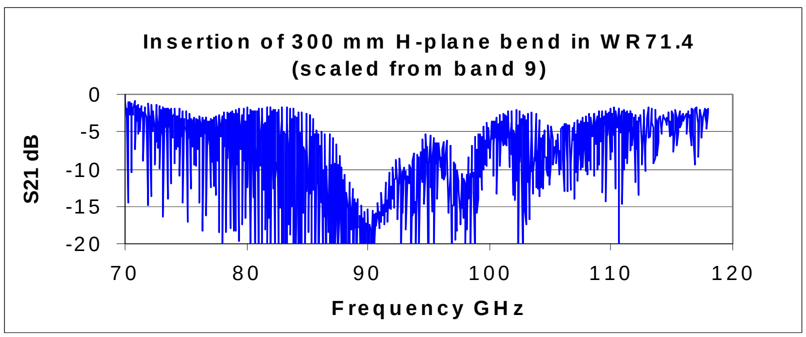

Figure 13

Measurements at 624-720 GHz with WR10 Components

Measurements at 624-720 GHz were made using a x6-multiplied Gunn (RPG [21]) and an overmoded power head (Anritsu 90-140 GHz [22]). The frequency was set by hand with no phase lock, and power levels were measured with and without the device under test between two WR3.4 to WR10 tapers (back-to-back). Figure 14 shows the measured and theoretical losses for a 152.4-mm length of WR10 stainless steel waveguide, and for the same waveguide after plating with 2.5 microns of copper. (12 mm at each end was copper-plated on the inside during attachment of the waveguide flanges; this is accounted for in the theoretical values.) The theoretical values include the 1.3 roughness factor for both, and the 1.135 skin-effect factor for copper but not for stainless steel [13]. There was insufficient power near 700 GHz to make any measurements.

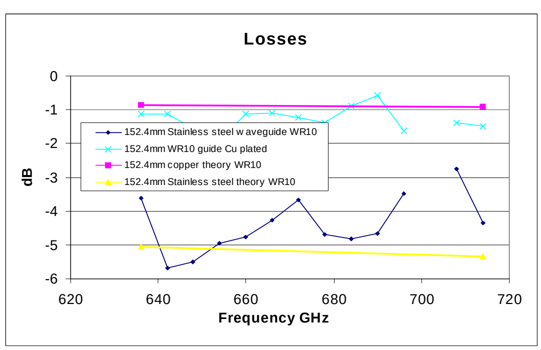

Figure 14

This clearly shows the improvement of plating copper or gold on the inside of the waveguide, but for very small waveguide (WR8 or smaller) or long lengths (100 mm or longer) this is extremely difficult. The sample was also measured at 75-110 GHz where it had 3.26-2.25 dB, unplated (compared to theoretical values, with roughness factor, and accounting for copper-plated sections at the ends, of 4.4 - 3.1 dB) and 0.63-0.47 dB, plated (theoretical values 0.76-0.54 dB with skin effect and roughness factors). The difference with the theoretical values can be explained if the waveguide is slightly larger than the nominal 2.54 by 1.27 mm. The losses of various components (H-plane bend {Aerowave [23]}, H-plane bend, E-plane bend and 90-twist {Baytron [24]}) are given in Figure 15. The reason for the difference between the two H-bends is unknown.

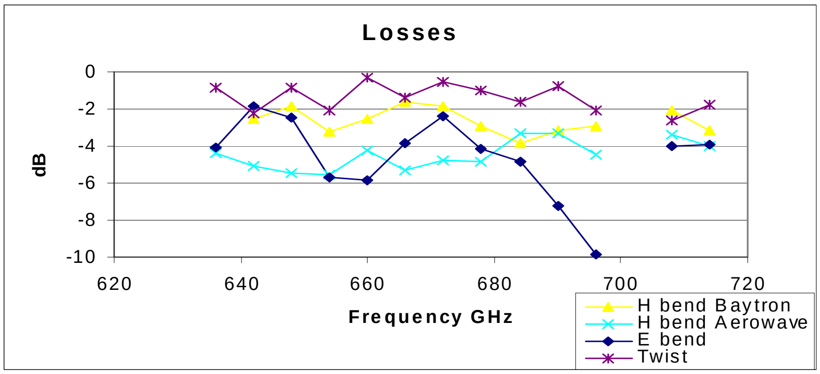

Figure 15

Two sections of WR1.4 (fundamental waveguide for this band) were made of brass in split-block technique and then gold-plated. One piece, 25.4 mm long, split in the narrow wall (incorrect wall) and one piece, 76.2 mm long, split in the broad wall. They had losses of 389 dB/m and 130 dB/m, respectively, at 684 GHz (theoretical value 130 dB/m with roughness factor).

Measurements at 220-330 GHz with WR10 Components

Measurements have been made in the range 220-330 GHz (0.5 GHz per point) using an HP8510 with WR3 extender heads (Oleson [25]). Again, spacing between the heads limits the size of structures which can be measured. Also, the system sensitivity and dynamic range are limited at these frequencies. Figure 16 shows the measured transmission of a taper from WR3.4 to WR10, 200 mm of waveguide followed by two H-plane bends and then one E-plane bend (all WR10 [24]) and then a taper from WR10 to WR3.4, divided by the system response with the heads connected directly together[2]. This shows that at room temperature there are no major resonances. Note that the fluctuations below 220 GHz and above 300 GHz are due to the low power levels in these frequency ranges. No resonances were seen when the frequency range was reduced to 240-245 GHz (0.025 GHz per point). Also shown in Figure 16 is the measured loss of two tapers, back-to-back, which have a theoretical loss of 0.38 dB for copper, including roughness and skin-effect factors.

It was not possible to cool the waveguide run to determine if resonances become significant when cold.

Figure 17 shows the insertion loss of 150 mm of WR10, copper-plated, and stainless steel WR6 and WR5, measured between two tapers, over the frequency range 220-330 GHz, and compares it to the theoretical values (with roughness factors and skin-effect factors for copper-plated WR10) for those waveguides.

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[2]Full VNA calibration was not available on this instrument at the time it was used.

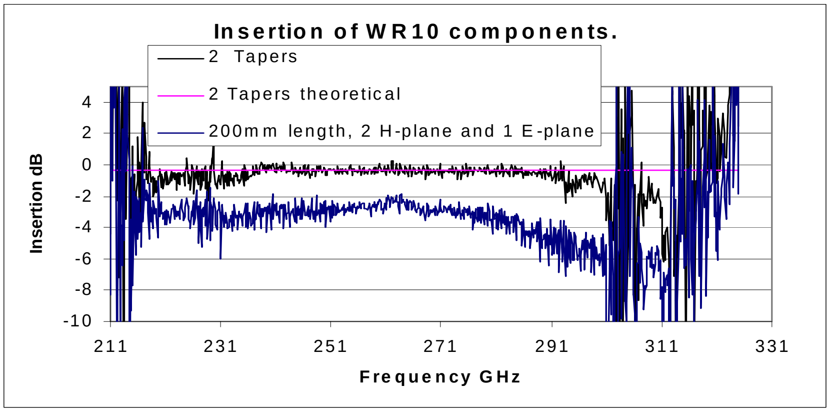

Figure 16.

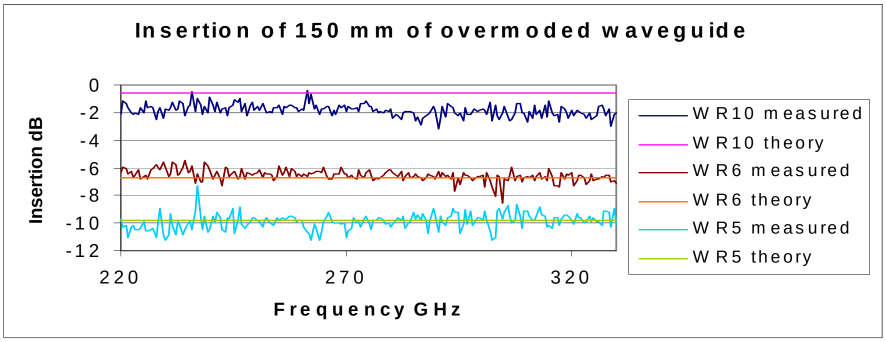

Figure 17. WR10 is copper-plated; WR6 and WR5 are stainless steel.

The excess loss of the WR10 is probably due to difficulties in plating, as measurements at 75-100 GHz show that this waveguide has twice the loss we usually measure for copper-plated stainless steel WR10.

Measurements of overmoded waveguide (WR10) made at IRAM for Band 7 at room temperature also show no resonances [26].

Conclusions

Overmoded waveguide is a possibility for low-loss LO transmission at Band 6, but trapped modes may be a problem when operated at cryogenic temperatures. Measured loss of straight waveguide compares well with theory when roughness and skin-effect loss factors are included. A simulation (at room temperature) of a 25.4-mm taper from WR3.4 to WR10, 150 mm of WR10 followed by a 25.4-mm taper back to WR3.4, had 1.0 dB loss with 0.3 dB resonances (20 dB mode coupling assumed), *i.e.*, maximum depth 1.3 dB, for copper (with roughness and skin-effect factors included), and 4.4 dB loss with 0.1 dB resonances if the WR10 section is unplated stainless steel (including roughness factor). To reduce phase problems, stainless steel may need to be used if the extra loss can be tolerated. Losses will reduce by approximately a factor of two when cooled to 90 K for copper, but the resonances will deepen (to approximately 0.6 dB loss with 2 dB resonances). Losses will reduce to 3.3 dB with 0.2 dB resonances for stainless steel.

Acknowledgment

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