# 300GHz Linearly Tapered Slot Antenna Design and Measurements

**Seunghwan Kim1 and Alenka Zajić1**

**1**School of Electrical and Computer Engineering, Georgia Institute of Technology, Atlanta GA 30332, USA

Corresponding author:

Alenka Zajić,

Georgia Institute of Technology, School of Electrical and Computer Engineering,

Address: 85 5th Street NW, Atlanta GA 30308

Email: alenka.zajic@ece.gatech.edu

Telephone: +1 404-385-6604

Fax: +1 404-894-4701

**Abstract:** *The first 300 GHz linearly-tapered slot antenna with the average gain of 13dBi has been designed, fabricated, and tested. To reduce the cost of testing, this paper also proposes to pair relatively low-cost 300-320 GHz communication system with 10 MHz - 30 GHz vector network analyzer and use the signal processing techniques.*

**Key words:** *Terahertz antenna design, terahertz antenna measurements, planar antenna, linearly tapered slot antenna.*

## 1. Introduction

Ultra-broadband terahertz (THz) communication systems are expected to help satisfy the ever-growing need for smaller devices that can offer higher speed wireless communication anywhere and anytime. In the past years it has become obvious that wireless data rates exceeding 10 Gbit/s will be required in several years from now [1]. The opening up of carrier frequencies in the terahertz-range is the most promising approach to provide sufficient bandwidth required for ultra-fast and ultra-broadband data transmissions [2]. A suitable frequency window can be found around 300 GHz, offering an unregulated bandwidth of 47 GHz [2]. This large bandwidth paired with higher speed wireless links opens up the door to a large number of novel applications such as ultra-high-speed cellular links, wireless short-range communications, secure wireless communications for military and defense applications, and on-body sensors for health monitoring systems. To design THz communications systems, we need wideband, high-gain antennas to cover large frequency range and compensate for high propagation losses. In addition, THz antennas need to be planar and suited to be realized in integrated or printed circuit board (PCB) technology.

A horn antenna has been widely used at high mm-wave and THz frequencies because it provides a high gain and wide bandwidth [3]. However, the horn antenna has a three-dimensional (3-D) structure, which prevents the integration with other electronics. An alternative method for obtaining high antenna gain is to combine elliptical dielectric lenses and slot or dipole feeds [4]-[8]. However, this approach also has 3-D structure and a very limited bandwidth. To address the problem of bandwidth, leaky lenses have been proposed [9]-[13]. However the lenses still have 3D structure and thus are not easily integrated with other electronics. The third method for obtaining high-gain is to use a slotted waveguide antenna. Several different non-mechanical-machining-based techniques have been proposed in [14]-[16]. Their success was limited by the compromise between the fabrication accuracy, material losses, and process complexity. In attempt to overcome these limitations, Wang et *al*. have proposed 300 GHz SU-8-based slotted waveguide antenna [17]. Although the antenna presents a simple and cost-effective SU-8 technique, the multilayered structure requires high precision micromachining, where even a slight gap between the layers leads to significant power loss. In addition, the measured gain of the slotted waveguide antenna was around 5 dBi, due to the resistive losses from the imperfect joints between SU-8 layers. This is a problem because antennas with high gain are essential at THz frequencies to compensate for the high atmospheric path loss. Finally, the fourth type of antenna that has high-gain, wide bandwidth, and a two-dimensional (2-D) structure is a tapered slot antenna [3]. However, the 300 GHz multilayer linearly tapered slot antenna (LTSA) simulated in [18] has either maximum gain of only 6 dBi or a very wide 3 dB beamwidth of up to 170° with a maximum gain around 3 dBi.

In this paper, we propose a LTSA with the average gain of 13 dBi and return loss below -10 dB across 280-320 GHz frequency range. These results are verified experimentally for the frequency range of 305–320 GHz because our low-cost measurements set-up can test only the upper-band of our antenna design. The LTSA has been fabricated using a standard PCB milling machine and the Rogers RT/Duroid 5880 material.

One of the main obstacles in designing THz communication systems is the cost of testing equipment. To overcome this problem, this paper also shows how pairing relatively low-cost 300-320 GHz communication system with 10 MHz- 30 GHz vector network analyzer (VNA) and using additional signal processing can be used to calculate the gain and return loss of a tested antenna.

The remainder of this paper is organized as follows: Section 2 describes the LTSA design, simulation results and fabrication process. Section 3 describes the measurement setup and the measurement environment in which the antenna was tested. Section 4 describes the post-processing algorithm for calculating the gain and return loss of a measured antenna. Section 5 compares the simulated and measured results. Finally, Section 6 presents concluding remarks.

## 2. Antenna Design and Fabrication

Figure 1 shows the layout of the proposed 300 GHz LTSA. The layout consists of 2 copper layers, each having a slot tapered from the top of the WR-3 waveguide to the top of the antenna. The antenna has been simulated with CST [19] and fabricated on a 15 mil Rogers RT/Duroid 5880 R3 board using a standard PCB milling machine.

The dielectric constant of the material is 2.2 and the copper thickness is 36 μm. The design parameters (i.e., initial slot width, a, tapered end slot width, b, slot length, h, substrate thickness, t, and the flare angle of the tapered slot, θ) and their numerical values are summarized in Table 1.

To connect our LTSA to the measurement system described in Section 3, we had to include a tail section that is inserted into the WR-3 waveguide as shown in Fig. 5. We have designed the tail in a way that it not only serves as the connection between the antenna and the waveguide, but it also helps reducing the reflections inside the waveguide. By introducing the tail section, the propagating wave does not experience an abrupt change of medium (from air to Duroid), which leads to smaller reflections at the waveguide-antenna interface. The CST simulation results are shown in Fig. 2. We can observe that the average antenna gain is around 13 dBi and the return loss is below -10 dB across all frequencies. The ripples observed in the LTSA gain are the result of multiple reflections off the face of the horns, as it will be confirmed in the measurements in Section 5.

## 3. Measurement Setup

The measurement setup consists of the N5224A PNA vector network analyzer (VNA), the VDI transmitter (Tx210) and the VDI receiver (Rx148). The input signal provided by the VNA at the Intermediate Frequency (IF) port of the Schottky diode mixer is mixed with the Local Oscillator (LO) signal, generated by subsequent doubling and tripling of the 25 GHz signal from a phase-locked dielectric resonator oscillator (DPRO). The sub-harmonic mixer plays a dual role of doubling the carrier frequency and mixing it with the baseband signal (10 MHz–20 GHz, delivered by the VNA). The resultant terahertz-range signal is then transmitted by the horn antenna that has a gain of 23 dBi in the range 280-320 GHz. At the receiver side, the same components are used to down-convert the signal, except that the DPRO is tuned to 24.2 GHz, resulting in a down-conversion of the received RF signal to an IF signal of 9.6 GHz. The upper sideband of the down-converted signal is then recorded by the VNA in the frequency range of 9.6-29.6 GHz. The corresponding block diagram is shown in Fig. 3. By recording the frequency dependent scattering parameter S21 for the test signal frequencies ftest = 10 MHz−20 GHz at the VNA, the channel transfer function at f = 300GHz+ ftest is measured.

It has been found that the inherent loss in the transceiver is very high (40-50 dB) and that this loss has to be de-embedded from any S21 measurements to obtain true S21 transfer functions. The bandwidth of 15 GHz is used in all measurements to avoid the Tx amplifier distortions present in the range of 300–305 GHz. This provides the temporal resolution of 0.067 ns.

The start frequency is bound to a minimum of 10 MHz by the VNA and the stop frequency could not exceed the

system limitations of 20 GHz. Due to input power restrictions of the mixers, a test signal with a power of −5 dBm is used, providing a dynamic range of approximately 90 dB for the chosen intermediate frequency filter bandwidth of ΔIF = 10 kHz. The number of sweep points is set to 801, and the maximum excess delay is 53 ns.

To obtain the gain and the return loss of the fabricated LTSA, two measurement scenarios have been employed. First, the channel transfer function, S21, is measured between the two identical horn antennas, one on the Tx and the other on the Rx module as shown in Fig. 4. This measurement setup is used to find the frequency dependent gain and return loss of the horn antenna.

Second, S21 is measured between the horn antenna on the Tx and the LTSA on the Rx side, as shown in Fig. 5, to find the frequency dependent gain and return loss of the LTSA. The post-processing of the measured data is described in Section 4.

## 4. POST-PROCESSING OF MEASURED DATA

This section describes the signal processing used to obtain the gain and the return loss of the LTSA. The steps of signal post-processing are as follows:

1) The transceiver loss is de-embedded from the measured S21. This step is necessary because the system calibration can only be performed at the input and the output of the Tx and Rx modules, while the transceiver introduces significant frequency-dependent loss into the system.

2) The measured channel transfer function between two horn antennas is used to calculate the frequency dependent gain and S11 of the horn.

3) The obtained gain and S11 of the horn from Step 2 along with the measured S21 between the horn and LTSA are used to calculate the frequency dependent gain and S11 of the LTSA.

The following two subsections describe the detailed signal processing techniques used in steps 2 and 3.

*A. Horn-to-Horn Measurement Scenario*

Figure 6 shows the diagram of the horn-to-horn measurement scenario. To obtain the frequency dependent gain of the horn antenna, we use the Friis equation [20]:

, (1)

where is the measured transfer function S21, Ghorn is the frequency-dependent gain of the horn antenna, and PL is the free-space path loss, which can be theoretically calculated as [21]

, (2)

where d denotes the distance between two horn antennas, and λ is the wavelength. In this measurement scenario, the distance between the horn antennas was d = 10 cm.

To find the S11 of the horn, we consider the diagram shown in Fig. 6. Here, the Tx and the Rx antennas have the same reflection and transmission coefficients, and , since two identical horn antennas are used. Assuming that the horn has 100% efficiency, the relationship between the Sh11 and Sh21 can be defined as [22]:

. (3)

The measured S21 includes path loss and twice the horn gain, which need to be compensated for to find the true channel transfer function. Therefore, in Fig. 6 is calculated as

, (4)

where S21deembed refers to the measured S21 after the de-embedding of the transceiver loss, and Gh is the average of the frequency dependent horn gain. From Fig. 6, we can relate and as follows:

, (5)

which leads to

. (6)

Finally, can be found by substituting (6) into (3) as

. (7)

The measured (i.e., post-processed) S11 and gain of the horn are shown in Fig. 7. They are also compared with simulated S11 and gain of the horn antenna to verify our approach. The comparison is further discussed in Section 5.

*B. Horn-to-LTSA Measurement Scenario*

The algorithm described in Section 4-A is also used in the second measurement scenario to find the gain and the return loss of the LTSA. The only modification from the first scenario is that the Rx horn antenna is now replaced with the proposed LTSA, and the and on the Rx side, and in Figure 6 are substituted with , , and , respectively. Note that the separation distance of d = 1.5 cm has been chosen in this measurement scenario to ensure direct line of sight between the horn and LTSA, as illustrated in Fig. 5.

Following the similar reasoning as in the horn-to-horn measurement scenario, we can observe that (5) can be rewritten as

, (8)

where is the de-embedded S21 between the horn and the LTSA that has been compensated for path loss and the gains of the two antennas, is the S21 of a single horn that was found in the previous section, and is the S21 of the LTSA that needs to be calculated. Rearranging (8) for , and employing the relationship between and as in (3), we obtain

. (9)

The measured S11 and gain of the LTSA are shown in Fig. 8. They are also compared with the simulated S11 and gain of the LTSA to verify our approach. The comparison is further discussed in Section 5.

## 5. COMPARISON BETWEEN SIMULATED AND MEASURED RESULTS

Figures 7 and 8 compare the simulated and measured gain and return loss of the horn and the proposed LTSA, respectively. For the simulated results, the measurement scenarios described in Section 4-A and 4-B are simulated in CST [19], and the obtained S21’s have been subject to the identical post-processing procedure described in Section 4.

In Figs. 7 and 8, it is observed that the measured gain and S11 are quite comparable with the simulated ones for both the horn and LTSA. The ripples seen in the measured gain are confirmed by simulation, and it has been found that they are the result of multiple reflections between Tx and Rx hardware. The fabricated LTSA achieves a high gain of around 13 dBi, and below −10 dB return loss for most of the measured bandwidth of 305–320 GHz, as observed in Fig. 8. Although the measured results have been plotted for only the upper sideband with additional loss of 5 GHz due to amplifier distortion present in 300–305 GHz, simulation has proven that, in principle, the LTSA has a high gain and low S11 below −10 dB across the entire bandwidth of 280–320 GHz as shown in Fig. 2, confirming the wideband (13.3 % fractional bandwidth) characteristic of the proposed antenna. The slight discrepancies observed between measurement and simulation can be attributed to imperfections in fabrication, transitions between the waveguide and antenna, rough surfaces, etc. that would introduce additional losses not accounted in our post-processing algorithm. It can be expected that at such high frequencies, the slightest dimensional irregularity of the structure that is in direct contact with the propagating wave can result in considerable reflections.

## 6. Conclusions

A broadband linearly-tapered slot antenna with the average gain of 13 dBi across 280–320 GHz range has been designed, fabricated, and tested. The LTSA has been fabricated using a standard PCB milling machine and the Rogers RT/Duroid 5880 material. To reduce testing cost, this paper proposed to pair relatively low-cost 300–320 GHz communication system with 10 MHz–30 GHz vector network analyzer (VNA) and use signal processing to extract the gain and return loss of the tested antenna. The results show that the measured average gain and return loss are in good agreement with the simulation results, suggesting that the fabricated LTSA has a high gain and wideband characteristics.

## References

1. S. Cherry, “Edholm’s law of bandwidth”, *IEEE Spectrum*, vol. 41, pp. 58-60, July 2004.
2. P. H. Siegel, “Terahertz technology”, *IEEE Transactions on Microwave Theory and Techniques*, vol. 50, pp. 910–928, March 2002.
3. J. D. Kraus, *Antennas*, McGraw-Hill Book Company, 2nd edition, New York NY, 1988.
4. D. B. Rutledge and M. S. Muha, “Imaging antenna arrays,” *IEEE Transactions on Antennas and Propagation*, vol. 30, no. 4, pp. 535–540, July 1982.
5. D. F. Filipovic, S. S. Gearhart, and G. M. Rebeiz, “Double slot antennas on extended hemispherical and elliptical silicon dielectric lenses,” *IEEE Transactions on Microwave Theory and Techniques*, vol. 41, no. 10, pp. 1738–1749, October 1993.
6. X. Wu, G. Eleftheriades, and T. E. van Deventer-Perkins, “Design and characterization of single- and multiple-beam mm-wave circularly polarized substrate lens antennas for wireless communications,” *IEEE Transactions on Microwave Theory and Techniques*, vol. 49, no. 3, pp. 431–441, March 2001.
7. A. V. Boriskin, G. Godi, R. Sauleau, and A. I. Nosich, “Small hemielliptic dielectric lens antenna analysis in 2-D: boundary integral equations versus geometrical and physical optics,” *IEEE Transactions on Antennas and Propagation*, vol. 56, no. 2, pp. 485–492, February 2008.
8. P. Focardi, W. R. McGrath, and A. Neto, “Design guidelines for terahertz mixers and detectors,” *IEEE Transactions on Microwave Theory and Techniques*, vol. 53, no. 5, pp. 1653–1661, May 2005.
9. A. Neto, S. Bruni, G. Gerini, and M. Sabbadini, The leaky lens: A broad band, fixed beam leaky wave antenna, IEEE Trans. Antennas Propag., vol. 53, no. 10, pp. 32403246, Oct. 2005.
10. S. Bruni, A. Neto, and F. Marliani, “The UWB leaky lens antenna,” *IEEE Transactions on Antennas and Propagation*, vol. 55, no. 10, pp. 2642–2653, October 2007.
11. A. Neto, “UWB, non-dispersive radiation from the planarly fed leaky lens antenna-part I: theory and design,” *IEEE Transactions on Antennas and Propagation*, vol. 58, no. 7, pp. 2238–2247, July 2010.
12. A. Neto, S. Monni, and F. Nennie, “UWB, non-dispersive radiation from the planarly fed leaky lens antenna-part II: demonstrators and measurements,” *IEEE Transactions on Antennas and Propagation*, vol. 58, no. 7, pp. 2248–2258, July 2010.
13. N. Llombart, G. Chattopadhyay, A. Skalare, and I. Mehdi, “Novel terahertz antenna based on a silicon lens fed by a leaky wave enhanced waveguide,” *IEEE Transactions on Antennas and Propagation*, no. 6, pp. 2160–2168, June 2011.
14. Y. Fuh, A. Margomenos, Y. Jiang, and L. Lin, “Micromachined W-band plastic slot array antenna with self-aligned and integrated flange,” *15th Annual International Conference on Solid- State Sensors, Actuators, and Microsystems*, pp. 2122-2125, June 2009.
15. T. Tick, J. Jantti, M. Henry, C. Free, and H. Jantunen, “LTCC integrated air-filled waveguide for G-band applications,” *Microwave and Optical Technology Letters*, vol. 51, no. 1, pp. 176- 178, 2009.
16. M. Henry, C. E. Free, B. S. Izqueirdo, J. Batchelor, and P. Young, “Millimeter wave substrate integrated waveguide antennas: design and fabrication analysis,” *IEEE Transactions on Advanced Packaging*, vol. 32, pp. 93-100, February 2009.
17. Y. Wang, M. Ke, M. J. Lancaster, and J. Chen, “Micromachined 300-GHz SU-8-based slotted waveguide antenna,” *IEEE Antennas and Wireless Propagation Letters*, vol. 10, pp. 573–576, 2011.
18. M. Jenning, and D. Plettemeier, “Multilayer and multidirectional linearly-tapered slot antenna for 300 GHz applications, *The 4th European Conference on Antennas and Propagation (EuCAP10)*, pp. 1–5, Barcelona Spain, April 2010.
19. (2010) CST Website. [Online]. Available: <http://www.cst.com/>
20. T. Rappaport *Wireless communications: principles and practice*, Upper Saddle River, NJ, USA: Prentice Hall PTR, 2001.
21. A. Zajic, *Mobile-to-mobile wireless channels*, Artech House, Boston MA, January 2013.
22. D. M. Pozar: *Microwave Engineering*, 3rd Edition, New Jersey: John Wiley & Sons, 2005, pp. 337-352.