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RESEARCH ARTICLE

A Complete Circuit Model for Terahertz Spoof Surface Plasmon Polariton Waveguides for Ultrafast and Accurate Synthesis of Terahertz Integrated Circuits

MUHAMMED ABDULLAH UNUTMAZ¹, KORKUT KAN TOKGOZ², (Member, IEEE),
AND MEHMET UNLU³, (Senior Member, IEEE)

¹Department of Electrical and Electronics Engineering, Ankara Yıldırım Beyazıt University, 06760 Ankara, Turkey

²Faculty of Engineering and Natural Sciences, Sabanci University, 34956 Istanbul, Türkiye

³Department of Electrical and Electronics Engineering, TOBB University of Economics and Technology, 06510 Ankara, Türkiye

Corresponding author: Mehmet Unlu (munlu@etu.edu.tr)

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ABSTRACT The increasing demand for ultra-wideband wireless communications has pushed requirements of integrated circuits (ICs) to terahertz band. Operation in the terahertz band requires ultra-high efficiency for all the building blocks in ICs; however, the conventional passive components prevent achieving desired output power levels. Moreover, these components also strongly impede the synthesis of ICs as the design flow requires an enormous number of full-wave analyses. In this paper, we present the first-ever, complete, closed form, theoretical model for Terahertz Spoof Surface Plasmon Polariton Waveguides (THz-SSPP WGs), which hold the record for the lowest loss performance among all the planar waveguides at 0.3 THz. The electromagnetic field distribution around the THz-SSPP WG is non-uniform that even makes the solution of the problem of a nonlinear differential equation of multiple variables even more challenging. Because the guided-wavenumber is non-linear because of the dispersive behavior of the waveguide and the field distribution, hence the boundary conditions depend on all the corrugation dimensions, substrate thickness, and material properties. The proposed model is the first complete closed-form model not only among all THz-SSPP WGs, but also any single conductor planar waveguide. The proposed model illustrates the behavior of the confined and propagated electromagnetic wave around the THz-SSPP WGs, where the field distribution pattern changes as any of the waveguide parameter changes in contrast with any conventional waveguide with a ground.

INDEX TERMS Distributed circuit model, equivalent circuit model, single-conductor waveguides, terahertz spoof surface plasmon polaritons, terahertz integrated circuits.

I. INTRODUCTION

The escalating motivation for high data rate wireless communications has brought the desperate need to increase the bandwidth of modern wireless communication systems. Several groups have been working on develop-

ing wireless communication systems with the target of increasing the bandwidth using both photonics-based [1] and electronics-based [2] approaches. These systems have reached up to wireless data rates higher than 100 Gbit/s with these approaches, also pushing the limits for the operating frequencies significantly, reaching up to the terahertz band (0.1-10 THz) [3], [4]. Besides, these systems that target the terahertz band are capable of transmitting data thru at

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least 110 m long wireless links [3]. On the other hand, the implementation of a high-performance ultra-wideband wireless communications (UWBWC) system confronts the excruciating atmospheric attenuation [1]. The first terahertz communication window, that is 0.2-0.32 THz band [5], has a great potential for the demonstration of novel wireless communication networks, thanks to the fact that it has the widest bandwidth that does not suffer from the atmospheric attenuation within a bandwidth of 120 GHz [5]. Thus, there is great competition to demonstrate high-performance terahertz systems, where an initial industrial standardization has already been introduced with IEEE802.15.3d.

The most critical set of components for the high-performance terahertz wireless communication systems are either photonic- or electronic-based high-performance integrated circuits (ICs), which include several building blocks such as sources [6], oscillators [7], [8], mixers [8], low-noise amplifiers [9], harmonic generators [10], phase shifters [11], modulators [12], [13], generators [14], detectors [14], [15] and antennas [16]. However, the performance of the state-of-the-art terahertz wireless communication systems has not reached the desired performance, most importantly output power levels. This is basically not only caused by the transistor performance, but also, more commonly, caused by the extreme losses of the passive components in each building block. One of the main reasons behind this situation is the scarce performance of the conventional planar waveguides [2], [9], which unavoidably exist in all the above-mentioned ICs as irreplaceable fundamental components or interconnects between each other. Thus, the performance of the conventional planar waveguides, such as coplanar waveguides [17], microstrip lines [18], coplanar strip lines [19] and substrate-integrated-waveguides [20] dramatically affect the performance of each and every IC component due to their excessive losses at terahertz frequencies, and hence the overall performance of the terahertz wireless communication system. As a result of that, the signal-to-noise-distortion-ratio (SNDR) of the entire system degrades dramatically [21]. Meanwhile, novel, high-performance, easy-to-fabricate, planar terahertz waveguides, namely Terahertz Spoof Surface Plasmon Polariton Waveguides (THz-SSPP WGs) have recently been demonstrated with a record-low insertion loss per unit length performance at 0.3 THz [22]. Additionally, they exhibit advantageous properties, which are tuneability of the effective dielectric constant in a very wide range [23], high lateral confinement [24], potential for chip area miniaturization [25] and ability to suppress the signal integrity issues significantly [26]. All these properties of THz-SSPP WGs make them desired building blocks, enabling the possibility of integration of multiple terahertz waveguides and delay lines in small areas.

The SSPP WGs were first investigated at lower frequencies, such as the microwave band, before the real potential for the terahertz band was discovered. Several research groups

investigated the properties of the surface wave propagation at the interface between the metal and dielectric media [27], [28], [29], [30], [31], [32], [33], [34], [35], [36], [37], [38], [39], [40], [41], [42], [43], [44], [45], as the spoof surface plasmon waves traveling in the SSPP WGs in these frequencies exhibit similar behaviors that are also observed at the terahertz band. However, all these studies of THz-SSPP WGs [22], [23], [24], [25], [26], [27], [28], [29], [30], [31], [32], [33], [34], [35], [36] and SSPP WGs [37], [38], [39], [40], [41], [42], [43], [44], [46] had to employ time-consuming full-wave simulations because of their complex structures that result in complicated electric field distribution patterns. Consequently, the mathematical solution of the electric field distributions requires solving complex mathematical problems with multiple and three-dimensional boundary conditions, where the THz-SSPP WGs' electrical lengths are also frequency dependent. Such a complex modeling problem requires a complicated circuit model that significantly differs from the classical models that are used for conventional waveguides in the IC simulation environments.

Two examples of theoretical modeling studies for the SSPP WGs have been presented for the microwave band [43], [44]. These models entail laborious eigenmode simulations to extract the effective dielectric constant of each complex-structured, single-conductor waveguide. Then, the extracted effective dielectric constant is used to calculate capacitive circuit elements of the SSPP WGs. Such an approach that uses time-consuming full-wave simulations may even require an iterative optimization process to meet the exact performance criteria in the case of some complicated circuits, such as phase shifters [25], [38], where several phase states with very high accuracy are required. In consequence, the design and optimization of the targeted THz-SSPP WG-based ICs may consume an intolerably long time eliminating the practicality of the existing models. Additionally, an alternative model for the THz-SSPP WGs operating at the terahertz band is suggested [45], where the model provides closed-form formula for only the reactive circuit elements. Nonetheless, it does not provide any closed-form formulation for the resistive circuit elements that represent the conductor and radiation losses, which are crucial not only for estimating the insertion loss, but also for predicting the insertion phase with a very low error. Instead, it entails a post-fabrication optimization process for determining the resistive circuit elements. Although it provides the equations of reactive circuit elements of the THz-SSPP WGs, it unavoidably requires measurements of all the THz-SSPP WG circuits to extract the resistive circuit elements. On the other hand, high-performance and compact ICs for the terahertz applications require estimating both insertion phase and insertion loss performances to demonstrate different building blocks of a terahertz wireless communication system inside a single chip. Nevertheless, as the circuit complexity increases, the number of the THz-SSPP WGs and their corresponding states increase significantly, and hence the necessary memory for the full-wave simulations may reach

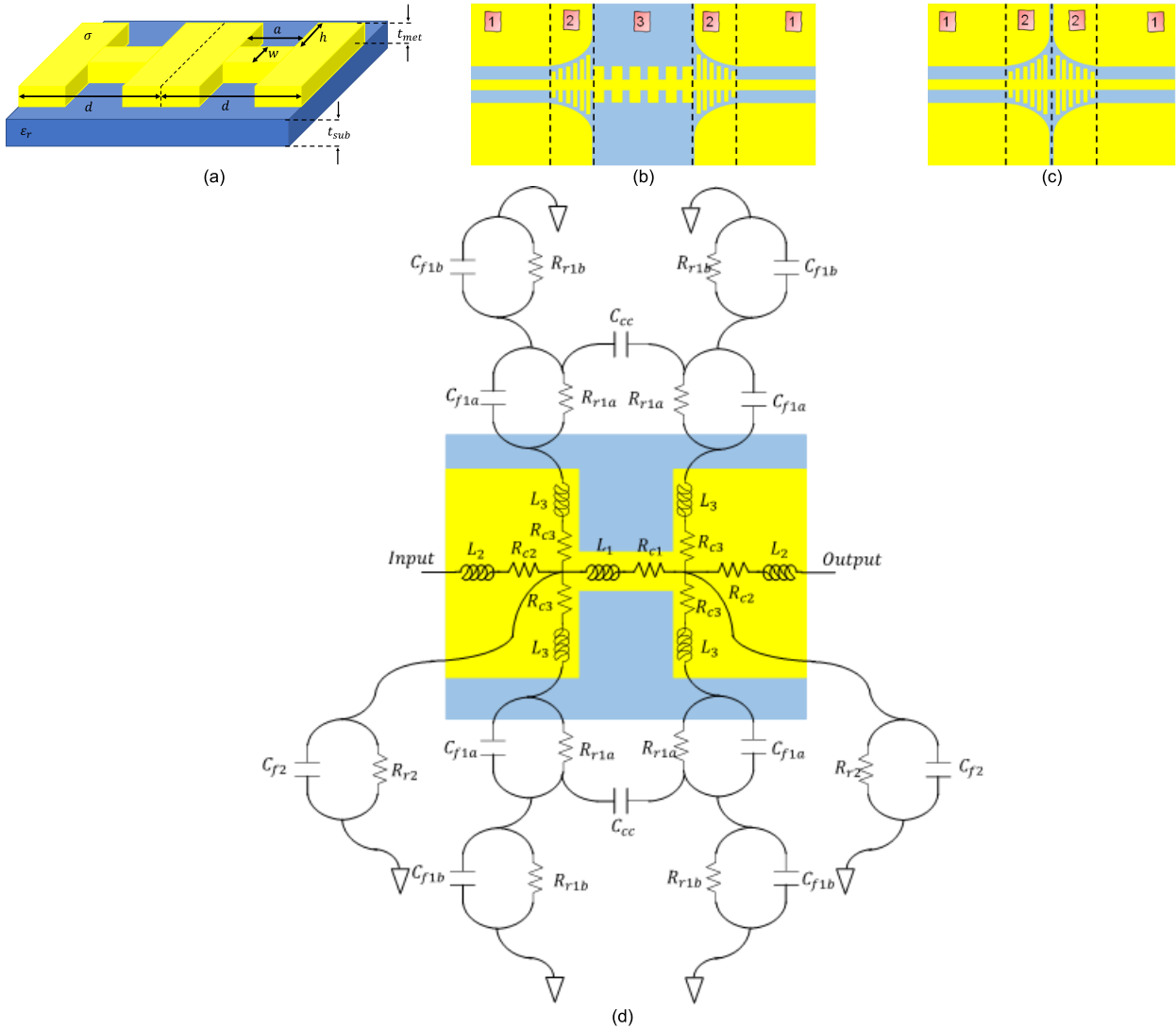


FIGURE 1. The electric field and current field distribution at the top surfaces of the substrate and conductor of Circuit-1 (a and b).

extreme levels that only high-performance computers (HPCs) may undertake successfully at the expense of intolerably long full-wave simulation times.

Moreover, the iterative IC synthesis processes of the complex terahertz ICs ineluctably require numerous employments of these lengthy full-wave simulations, which makes the overall design flow challenging and time consuming. Consequently, there is a vital need for a theoretical model of THz-SSPP WGs that enables the direct and rapid synthesis of complex ICs, eliminating the employment of time-consuming full-wave simulations.

In this paper, to the best of our knowledge, we present, for the first time in the literature, a complete theoretical model for the THz-SSPP WGs and any single conductor planar waveguide attached on top of a substrate, while the first example of the single conductor waveguides was introduced by Sommerfeld in 1899 [54]. Compared to the earlier solutions in the literature [43], [44], [45], the proposed model brings

the advantage of presenting all the equations that can model all the circuit elements and eliminates the need for utilization of eigenmode analysis and post-fabrication optimization that entails measurements of the THz-SSPP WG under investigation. The proposed model is the first complete theoretical model of the THz-SSPP WGs that can predict the field confinement around the single-conductor waveguide and the radiation loss concurrently. Furthermore, it successfully predicts the complicated, frequency-dependent conductor loss behavior of the THz-SSPP WGs at the terahertz band using entirely closed-form formulae. To verify the proposed model, we have designed, fabricated, and measured several THz-SSPP WGs. We selected 0.25-0.3 THz band owing to the large available bandwidth for ultra-high data-rate wireless communications and SSPP WG dimensions that would provide examples from the dimensions which allow low loss [22] avoiding the dispersion effects. Because the dispersion would require additional engineering in any communication link to

maintain the high performance of the communication system. The experiments show that the circuit model solutions and measurement results show very good agreement with each other. The performance of the proposed model is also compared with the full-wave simulation methods. The details of the proposed model are given in the following sections of the paper. The idea behind the circuit model is given in Section II. The experimental verification and performance analyses are given in Section III, which includes error and estimation speed compared to numerical analysis simulation.

II. THE CIRCUIT MODEL

The structure of the THz-SSPP WGs that is studied for the proposed model is given in Fig. 1-(a). The waveguide is composed of a patterned single conductor with a thickness of t_{met} that is placed on top of a substrate with a substrate thickness and relative permittivity of t_{sub} and ϵ_r . The structure is composed of cascaded cells of the single-conductor waveguide that are corrugated on both sides and have different periods (d), aperture widths (a) and corrugation depths (h) for each different THz-SSPP WG. A group of THz-SSPP WG circuits is designed for the investigation of the proposed model, where the circuits also include coplanar waveguides (CPW) and CPW to THz-SSPP WG transitions on both input and output ports of the THz-SSPP WGs. The CPWs are necessary for feeding the overall circuits with the conventional transverse electromagnetic (TEM) wave excitation techniques for both the simulations and measurements. In contrast, the CPW to THz-SSPP WG transitions are required for surface wave excitation at the single-conductor and dielectric interface of the THz-SSPP WGs. In addition to the THz-SSPP WG circuits, a back-to-back (B-to-B) connected reference circuit is designed using only CPW and transition regions to examine the effects of these regions and extract them from the THz-SSPP WG circuit measurements. The detailed layouts of the THz-SSPP WG circuits and B-to-B reference circuit are given in Fig. 1-(b) and (c), respectively. The CPW and transition regions are kept identical in all designed, manufactured, and measured THz-SSPP WGs for a fair comparison between them and to maintain the reliability of the THz-SSPP WG characterization. The equivalent circuit model that is used for the CPW and transition regions is based on the model proposed by Cinar et al. [43], which is verified to have very good agreement thanks to an optimization process. The transitions bring an additional total loss of -3.7 , -3.9 , and -4.4 dB at 0.25, 0.275, and 0.3 THz, respectively [34]. Since each different THz-SSPP WG has different corrugation dimensions, it has been shown to exhibit different electrical characteristics that result in different effective dielectric constant [23], radiation loss [22], [25] and conductor loss [22], [25] parameters, several circuit elements are incorporated in the model to be able to discriminate the effect of each and every parameter in the electromagnetic wave propagation behavior. The proposed model is given in Fig. 1-(d).

Several THz-SSPP WGs are utilized to examine the proposed model considering a wide dimension range. The

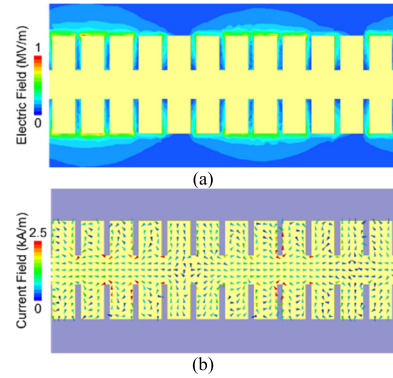


FIGURE 2. The electric field and current field distribution at the top surfaces of the substrate and conductor of Circuit-1 (a and b).

TABLE 1. THz-SSPP WG dimensions.

Circuit #	1	2	3	4	5
a (μm)	5	5	5	10	10
d (μm)	25	25	25	40	25
h (μm)	30	40	50	30	30
Period number	16	16	16	10	16

dimensions of the THz-SSPP WGs under investigation are given in Table 1 and selected to provide a variety of circuits that exhibit different levels of insertion loss and phase characteristics. The reactive circuit elements are used for modeling the field distributions around the single conductor waveguide, which are the self-capacitances, coupling capacitance between the corrugation walls, and self and mutual inductances of the metals. C_{f1a} and C_{f1b} model the fringing fields at the end of the corrugations, C_{f2} represents the field confinement around the connection points of the corrugation and central waveguide conductor, and C_{cc} represents the coupling fields in between the corrugation walls. C_{f1a} , C_{f1b} and C_{f2} are calculated by using the approach proposed by Kianinejad et al. [42] by replacing the extracted effective dielectric constant from fullwave simulations by the effective dielectric constant formula proposed by Unutmaz et al. [23]. C_{cc} is calculated using an approach that is employed for calculating the capacitance between two planar metals placed on top of a dielectric substrate [47]. Nevertheless, this formula utilizes multiplication of the relative permittivity of the dielectric substrate with the vacuum permittivity. This approach necessitates replacement of this dielectric permittivity parameter with the effective dielectric constant of the THz-SSPP WGs [23], which is denoted as ϵ_{eff} , because of the mode of the propagating wave. The inductive circuit elements L_1 and L_2 are used for modeling the self-inductance of the central waveguide conductor, whereas L_3 is used for modeling the self and mutual inductance of the corrugation metals. L_1 and L_2 are calculated with respect to the self-inductance part of formulation proposed by Meade et al. [48] and L_3 is the combination of self- and mutual-inductances due to the current distribution that emerge on and in between the

TABLE 2. The Capacitive circuit elements.

Circuit #	C_{cc} (fF)	C_{f1a} (fF)	C_{f1b} (fF)	C_{f2} (fF)
1	5.44	0.75	0.25	7.62
2	8.47	0.90	0.30	8.89
3	12.94	1.05	0.35	10.86
4	5.02	0.90	0.30	9.64
5	4.04	0.68	0.22	7.48

TABLE 3. The inductive circuit elements.

Circuit #	L_1 (pH)	L_2 (pH)	L_3 (pH)
1	3.11	0.68	23.81
2	3.11	0.68	35.86
3	3.11	0.68	48.92
4	8.98	1.08	18.49
5	4.92	0.69	22.86

TABLE 4. Electric field penetration towards dielectric substrate floor.

Circuit #	1	2	3	4	5
δ_s —Simulation (μm)	64.9	55	46.1	65.6	68.9
δ_s —Calculation (μm)	50.8	46.7	41.9	52.8	51.4

corrugations, respectively, which are also calculated by using the formulation presented by Meade et al. [48]. The electric field and current distributions of the THz-SSPP WGs also verify the validity of the elements in the proposed model, which is examined by using full-wave simulations in ANSYS HFSS environment. The corresponding simulation results are given in Fig. 2. The reactive circuit elements are found by using the formulae from (1) to (12) [45]. All the reactive circuit element values are given in Table 2 and 3.

$$C_{f1} = 8\epsilon_{eff}\epsilon_0\sqrt{\left(\frac{d-a}{2}t_{met}\right)} \quad (1)$$

$$C_{f1a} = (1 - K_f) C_{f1} \quad (2)$$

$$C_{f1b} = K_f C_{f1} \quad (3)$$

$$C_{f2} = 8\epsilon_{eff}\epsilon_0\sqrt{\left(\frac{d}{2}\frac{w}{\pi}\right)} \quad (4)$$

$$C_{cc} = \begin{cases} \frac{\epsilon_{eff}h \ln\left(2\left(\frac{1+\sqrt{k'}}{1-\sqrt{k'}}\right)\right)}{377\pi v_0} & 0 \leq k \leq \frac{1}{\sqrt{2}} \\ \frac{\epsilon_{eff}}{120v_0 \ln\left(2\left(\frac{1+\sqrt{k}}{1-\sqrt{k}}\right)\right)} & \frac{1}{\sqrt{2}} < k \leq 1 \end{cases} \quad (5)$$

$$k = a/d \quad (6)$$

$$k' = \sqrt{1 - k^2} \quad (7)$$

$$L_1 = \frac{\mu_0}{2\pi} \left(\frac{a+d}{2}\right) \left(\ln\left(\frac{2\left(\frac{a+d}{2}\right)}{w+t_{met}}\right) + .5 + .22\left(\frac{w+t_{met}}{\left(\frac{a+d}{2}\right)}\right)\right) \quad (8)$$

$$L_2 = \frac{\mu_0}{2\pi} \left(\frac{d-a}{4}\right) \left(\ln\left(\frac{2\left(\frac{d-a}{4}\right)}{w+t_{met}}\right) + .5 + .22\left(\frac{w+t_{met}}{\left(\frac{d-a}{4}\right)}\right)\right) \quad (9)$$

$$L_3 = L_s + L_m \quad (10)$$

$$L_s = \frac{\mu_0}{2\pi} h \left(\ln\left(\frac{2h}{\frac{d-a}{2} + t_{met}}\right) + 0.5 + 0.22\left(\frac{\frac{d-a}{2} + t_{met}}{h}\right)\right) \quad (11)$$

$$L_m = \frac{\mu_0}{2\pi} h \left(\ln\left(\frac{2h}{a}\right) - 1 + \frac{a}{h}\right) \quad (12)$$

In addition to the reactive circuit elements, the resistive circuit elements are also needed for modeling the insertion losses, which also have a direct effect on the insertion phase characteristics. The insertion loss of the THz-SSPP WGs are fundamentally dependent on the conductor and radiation losses, where the dielectric loss has already been shown to be negligible for high-resistivity dielectric substrates [22], [25]. R_{cn} ($n = 1, 2$ or 3) are used for modeling the conductor losses, and R_{r1a} , R_{r1b} and R_r are used for modeling the radiation losses. It should be mentioned here that there is a strict relationship in between R_{r1a} , R_{r1b} and C_{cc} , because not all the field is radiated, but some of the field is coupled to the adjacent corrugation wall. In consequence, a distributed model is proposed to represent coupled field in between the adjacent corrugations and the radiated field from the waveguide. The ratio between R_{r1a} and R_{r1b} is found empirically due to its complexity and their ratio is denoted by a constant K_f , which is the same parameter used for C_{f1a} and C_{f1b} . Initially, K_f is assigned a value of 1 but is later adjusted to 0.75. This is because the initial value cannot accurately predict radiation loss, which affects the estimation of insertion phase, return and insertion losses. The dispersive nature of the waveguides causes an alteration in the electric field distribution based on frequency. This distribution depends on various factors such as relative dielectric constant, substrate thickness, metal thickness, and corrugation dimensions. The radiation loss calculation at the corrugation walls is calculated using (13) and (14) below.

$$R_{r1a} = (1 - K_f) R_r \quad (13)$$

$$R_{r1b} = K_f R_r \quad (14)$$

R_r is a parameter that is related with the radiation loss of the THz-SSPP WGs, which is unavoidably defined having a unit of (1/m), since the series resistance and shunt conductance that are used to model the conductor and dielectric losses of microstriplines, for example, are normalized by twice the characteristic impedance and characteristic admittance, respectively [55]. This is because of the fact that the motivation behind the proposed manuscript is to model the single conductor under investigation whose characteristic impedance was unknown to the literature. R_r approximated using an inverse proportion of the penetration depth of the electromagnetic wave inside the substrate towards the bottom, i.e., δ_s . Because the areas with higher dielectric media in a surface waveguide are the locations where the confinement maximizes, and the radiated field is the field that cannot be confined around the waveguide. This is adopted from the surface plasmon waveguides, since the spoof surface plasmon polariton waves mimic the surface plasmons [50]. It has also been shown experimentally in our previous studies that

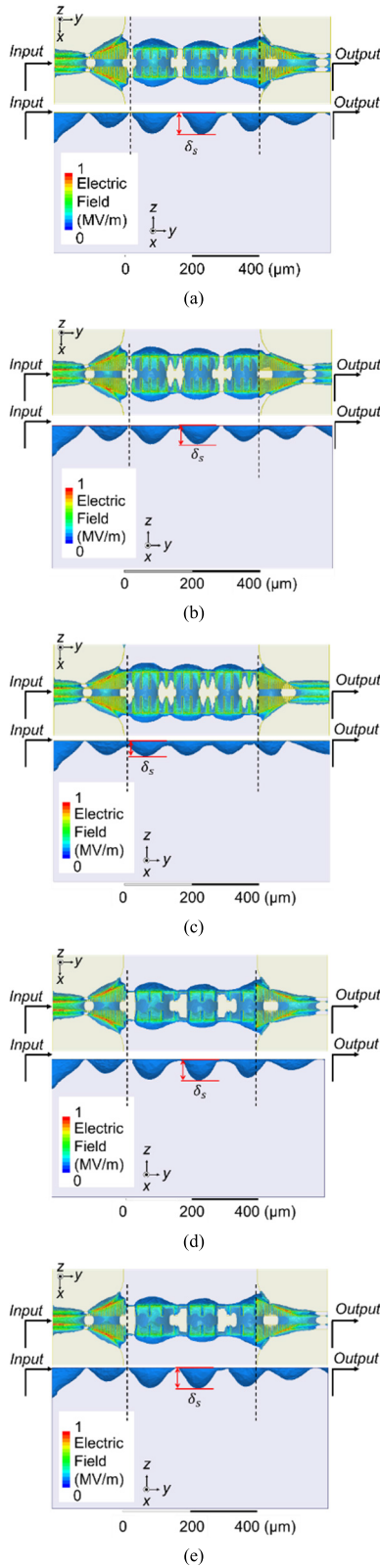


FIGURE 3. The electric field penetrations thru substrates of Circuit-1 to -5 (a to e).

THz-SSPP WGs with larger corrugation depth exhibit lower radiation loss [22], [23] which also agrees well with that δ_s

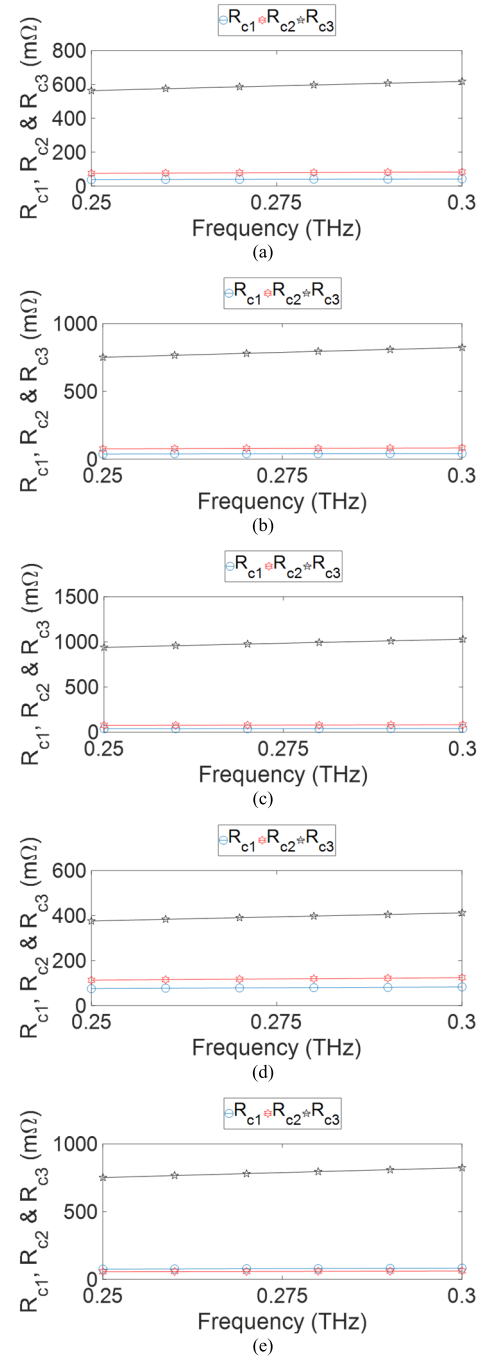


FIGURE 4. The resistive circuit elements that model the conductor losses: Circuit 1 to Circuit 5 (a to e).

decreases with increasing effective dielectric constant [49]. Hence, R_{r2} is selected as equal to R_r , which is calculated using (15).

$$R_{r2} = R_r = \frac{1}{\delta_s} = k_0 \sqrt{\epsilon_{eff} - 1} \quad (15)$$

The comparison of the calculated and simulated penetration depths inside the substrate of THz-SSPP WGs is given in Table 4 and Fig. 3, for verification of the reliability of the

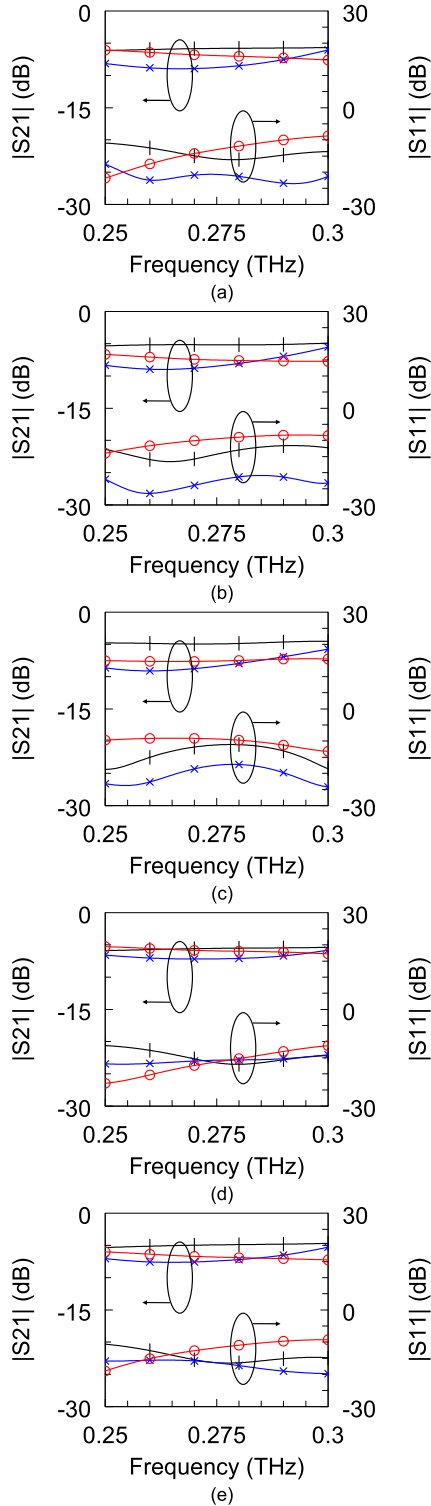


FIGURE 5. The S-parameter comparisons of model, FEM simulations and measurements: Circuit-1 to -5 (a to e). (Model: ○, full-wave Simulation: | and Measurement: ×).

idea behind the radiation loss estimation. The values of the resistive circuit elements that are used for modeling the radiation loss are given in Table 5. The resistive circuit elements that represent the conductor loss of the THz-SSPP WGs are

modelled using the equations given below.

$$R_{c1} = \frac{a}{\sigma_c t_{sd} w} \quad (16)$$

$$R_{c2} = \frac{\frac{d-a}{2}}{\sigma_c t_{sd} w} \quad (17)$$

$$R_{c3} = \frac{h}{\sigma_c t_{sd} \frac{(d-a)}{2}} \quad (18)$$

The frequency dependent metal conductivity behavior at terahertz band [53] is employed for calculating the resistive circuit elements that are used to model the conductor losses and the corresponding circuit element values are given in Fig. 4. Here, t_{sd} is the skin depth of the electromagnetic wave and changes as a function of the frequency.

III. THE EXAMINATION OF THE MODEL

The circuit model is examined by using experimental analyses and comparing its performance with full-wave analysis simulations. ANSYS HFSS is used for full-wave analysis, where the material properties are set in accordance with the fabrication process. The circuit model simulations are performed using AWR Microwave Office, which is employed in the simulations of the proposed model, where the circuits are designed to compose CPW, transition, and THz-SSPP WG regions to guarantee the equivalency of both simulation scenarios for fair comparison. The fabrication of the THz-SSPP WGs is realized using a single lithography fabrication process. The substrate is selected to be high resistivity Si with a resistivity of 10 kΩ-cm and a thickness of 425 μm. The lift-off process is employed, where an adhesion layer of Cr with 20 nm thickness and Au with 520 nm are used. The equivalent conductivity of the metal layer, which is composed of both Cr and Au, is 28 MS/m. The non-contact probing technology by TeraProbes, Inc. is used for the S-parameter measurements. Quick Offset Shorts (Q-SSS) calibration method [52] is used for measuring the circuits with CPW, transition, and THz-SSPP WG regions. The measurement setup that is employed during experimental verification is given in Fig. 3 in [52]. The comparisons of the S-parameters and insertion phase of the THz-SSPP WG circuits are given in Fig. 5 and 6, where the circuit model simulations, full-wave simulations, and measurement results are compared. The model of the CPW and transition region presented by Cinar et al. [45] is used because it provides very good agreement with the measurements according to Fig. 6 in [45]. The average and worst-case error rates for the insertion phases between the model and measurement results at 0.25, 0.275, and 0.3 THz are found as 4.8, 2.7 and 1.8%, and 10.7, 4.2, and 3.5%, respectively, which clearly shows the accuracy of the proposed model. The derivations between the proposed first-ever model of the THz-SSPP WGs and measurements are fundamentally caused by the empirical modeling approach behind especially the effective dielectric constant model [23].

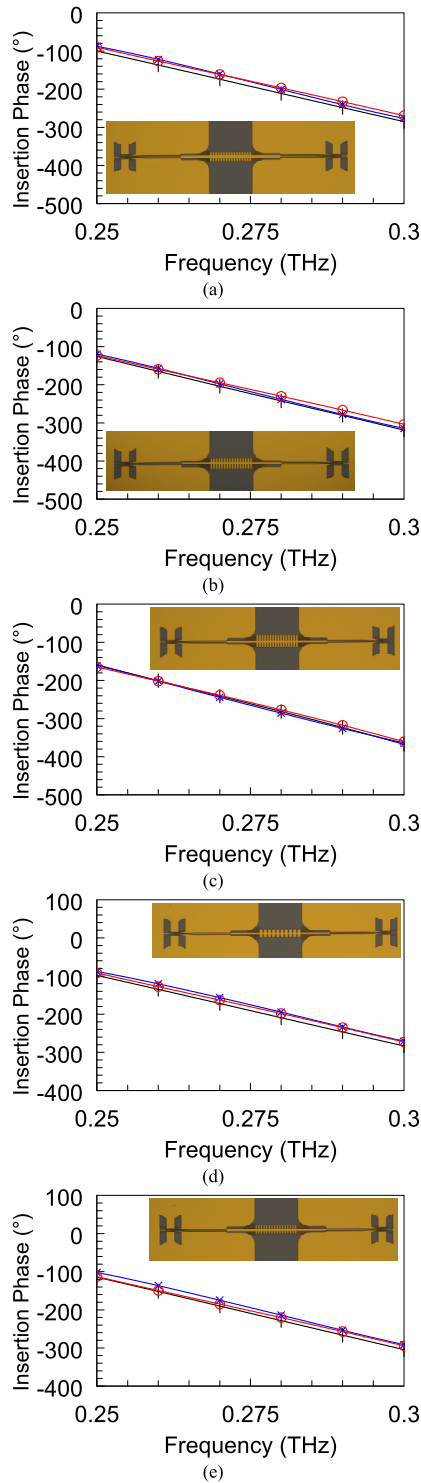


FIGURE 6. The insertion phase comparisons of model, FEM simulations and measurements: Circuit-1 to -5 (a to e). (Model: $\circ$, full-wave Simulation: $\times$ and Measurement: $\times$).

Because the effective dielectric constant model targets to solve the frequency-dependent effective dielectric constant, i.e., frequency-dependent wavenumber of the THz-SSPP WGs, that is a function of all the material properties and the dimensions. On the other hand, the full-wave simulations

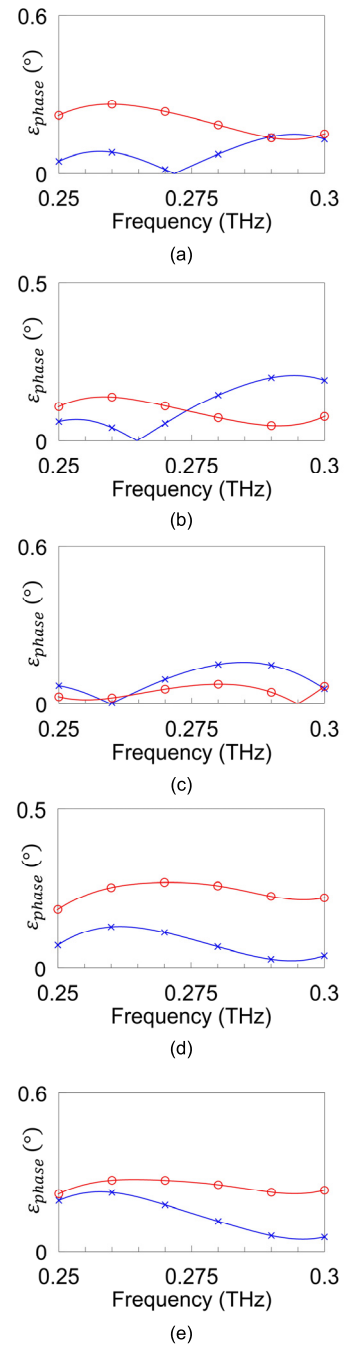


FIGURE 7. Insertion phase error ($\epsilon_{phase} = |IP_{Mea} - IP_{Model}|$): Circuit 1 to 5 (IP: insertion phase, Mea: measurement) (a to e). (FEM: $\circ$ and Model: $\times$).

TABLE 5. The calculated resistive circuit elements for radiation loss modeling.

Circuit #	R_r (k Ω)	R_{ra} (k Ω)	R_{rb} (k Ω)
1	19.67	14.75	4.92
2	21.40	16.05	5.35
3	23.84	17.88	5.96
4	19.67	14.75	4.92
5	19.47	14.60	4.87

do not agree perfectly with the measurements. The details of the insertion phase simulation and measurement data and

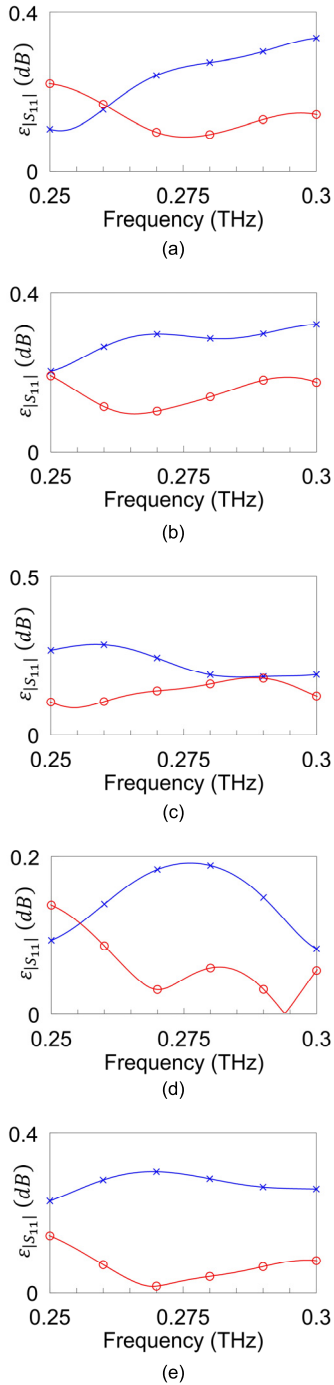


FIGURE 8. Return loss error ($\epsilon_{|S_{11}|} = |S_{11}|_{Mea} - |S_{11}|_{Model}$): Circuit 1 to 5 (Mea: measurement) (a to e). (FEM: $\circ$ and Model: $\times$).

TABLE 6. Simulation times.

Circuit #	B-to-B	1	2	3	4	5
FEM solver (minutes)	65	91	65	103	81	112
Circuit model (sec)	0.36	0.40	0.36	0.37	0.37	0.44

comparisons of the insertion phase and S-parameter errors of both the proposed model and full-wave simulations are given in Fig. 7 to 9. The circuit model is also used to

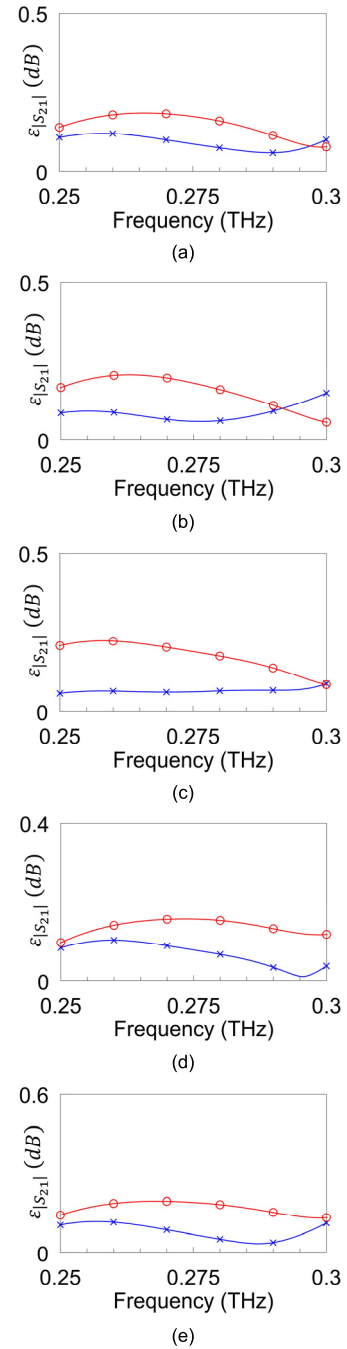


FIGURE 9. Insertion loss error ($\epsilon_{|S_{21}|} = |S_{21}|_{Mea} - |S_{21}|_{Model}$): Circuit 1 to 5 (Mea: measurement) (a to e). (FEM: $\circ$ and Model: $\times$).

estimate the upper cutoff frequency including the insertion phase, insertion loss, and return loss of the THz-SSPP WGs. In addition, eigenmode analysis simulations are done in HFSS environment, where the corresponding results are shared in Fig. 10-(d). The corresponding simulation results also show that the THz-SSPP WGs are band limited. On the other hand, using the available measurement setups, it is impossible to validate upper cutoff frequency estimation of both eigenmode analysis and circuit model simulations. The corresponding results are given on Fig. 10. After verification

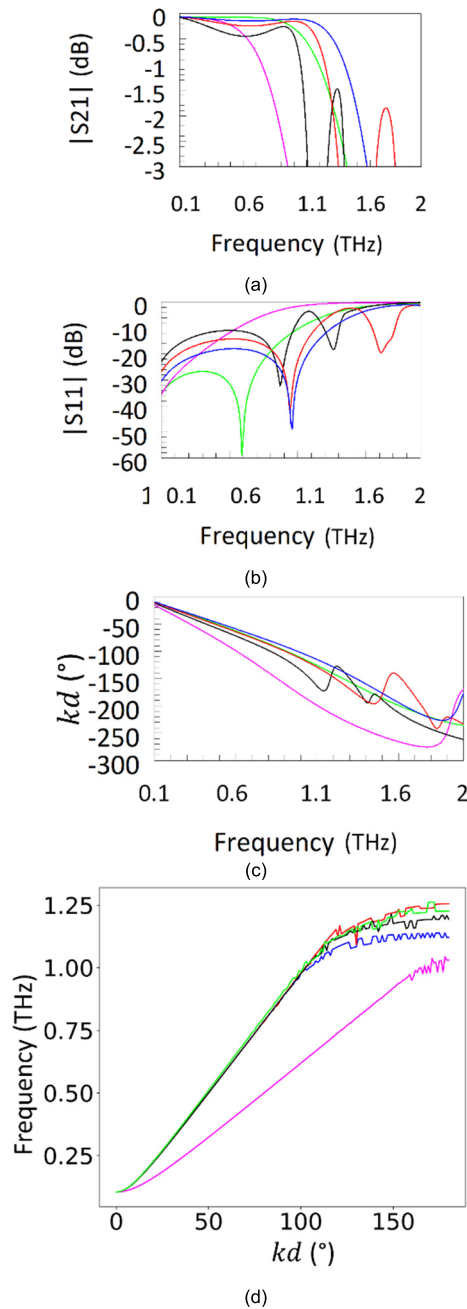


FIGURE 10. Single terahertz spoof surface plasmon polariton waveguide cell circuit model simulations (a to c) and eigenmode analysis solution (d). (Circuit-1: blue curve, Circuit-2: red curve, Circuit-3: black curve, Circuit-4: pink curve, Circuit-5: green curve).

of the accuracy and reliability of the proposed model, the simulation durations of the THz-SSPP WGs with the proposed model are investigated to demonstrate the vital impact that can cause on accelerating the design cycle. For this purpose, the simulation times of the THz-SSPP WG circuits using the proposed model are compared with that of the full-wave simulations performed in the ANSYS HFSS environment. The corresponding mesh sizes of the full-wave simulations are given in Table 7. The proposed circuit model decreases the simulation times as short as 0.36 seconds from 65 minutes in

TABLE 7. Fullwave simulation mesh size.

Circuit	FEM solver matrix size
B-to-B	334376
1	653712
2	543716
3	723810
1	589408
5	670032

the case of the circuit whose both circuit model and full-wave simulations are performed fastest in comparison with the other circuits. The details of the circuit model simulation times are given in Table 6. The average and longest simulation times of the circuits by using the circuit model are 0.39 and 0.44 seconds, whereas the average and longest simulation times of the circuits by using the full-wave simulations are 90.4 and 112 minutes, respectively; where the memory is 256 GB RAM and the system type is 64-bit Operating System, $\times 64$ -bit processor is used for both full-wave analysis and circuit model simulations. Both the simulation times and error rates with respect to the measurements show that the first examples of a THz-SSPP WG demonstration with entirely closed-form formulae pave the way to design the THz-SSPP WG for targeted circuit operation without needing any full-wave simulation and post-fabrication optimization with very low error rates and within supremely short simulation times.

IV. CONCLUSION

The terahertz communications have been a challenge for decades to enable wireless data transfer at a data rate that cannot be competed by any other commercial wireless communication technology. Nevertheless, the effort to demonstrate a commercial terahertz wireless communication technology is disabled by the integrated circuits that cannot provide high power radiation by the transmitter antenna. Despite the recent advancements in the development of terahertz integrated circuits, the lossy conventional waveguides have been one of the reasons behind this failure. On the other hand, the planar THz-SSPP WGs exhibit minimum per unit length loss among all the planar terahertz waveguides, which may become a solution to tackle the loss challenges that are experienced within the monolithic terahertz integrated circuits. Nonetheless, this requires a compact theoretical solution to enable the design of terahertz integrated circuits using these high-performance complex-structured single conductor waveguides. The proposed model is the first solution ever in the literature for these complex-structured single conductor waveguides despite the fact that the model is obviously limited to predict the waveguide response with minimum error. This is because of the fact that it is dependent on an effective dielectric constant model that was developed by using an empirical approach. And the proposed model targets $25^\circ < kd < 45^\circ$, that is the transmission regime of the SSPP WGs, where the dispersion curve is as linear as possible to avoid signal integrity challenges minimizing the radiation loss and conductor loss at the same

time [22], [25], especially in wideband wireless communications. Hence, unavoidably, as the frequency increases, the wavelength naturally decreases making it impossible for the surface wave to preserve its mode or at least distribution around the single conductor as in the case of linear region of the dispersion curve. In addition, neglecting such a distortion on the wave pattern, with the shorter wavelengths at optical regime, the material properties would not be identical to the material properties used for this circuit model. Consequently, the proposed circuit model cannot be valid for an infinitely wide band frequency of operation. On the other hand, the proposed model can be very useful for designing any THz integrated circuit that can take place in a 6G communication system. And considering analogy between the proposed model and the surface wave propagating through these waveguides, the proposed model can be extended to model more complex-structured single conductor waveguides. Therefore, the simulation of these complex-structured waveguides may require much longer simulation times proportional to the complexity of the structure, and hence the size of the resulting mesh size that is used in the full-wave simulations. Nevertheless, the complex pattern of a waveguide is not the only challenge. The estimation of the dispersion curve for a simpler technology that requires simpler measurement setups at a lower frequency band, verified that the eigenmode simulations may not show perfect agreement with measurements [56]. As the challenge of the waveguide geometry complexity increases, the proposed model can be an enabler to estimate the response of these waveguides within reasonable simulation times by decreasing the solution time by four orders of magnitude for the simplest-structured type of the complex-structured THz-SSPP WGs.

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MUHAMMED ABDULLAH UNUTMAZ received the B.Sc. degree from the Department of Electrical and Electronics Engineering, Hacettepe University, and the M.Sc. degree from the Department of Electrical Electronics Engineering, Ankara Yildirim Beyazit University. He was an Intern with the METU-Vision Research Laboratory, Department of Electrical and Electronics Engineering, METU. He founded and chaired the Student Society of Ham Radio at Hacettepe University.

He also worked in a start-up as an Research and Development Engineer, for nearly a year. He was a Research Assistant with the Department of Electrical Electronics Engineering, Ankara Yildirim Beyazit University. He was a Postdoctoral Researcher with the National Institute of Standard and Technologies, Boulder, CO, USA, as a NIST-PREP Program Postdoctoral Researcher with the Department of Physics, Colorado University, Boulder. He joined the SLAC National Accelerator Laboratory, in August 2024. During his Ph.D. studies, he was selected for Fulbright Visiting Student Researcher Scholarship. He has given four invited talks at international conferences and is the co-author of seven international journal articles, and several international conference papers. Prior to SLAC National Accelerator

Laboratory, he had the opportunity to achieve hands-on experience in video/image processing, telecommunications and radar signal processing, test and measurement of different circuits that can be used in radar, and telecommunications systems at various frequencies from RF to mmWave band and at various temperatures, design, fabrication, and measurements of micro-electro-mechanical-systems (MEMS) switches. He was selected for the 2022 Best Thesis Award by the IEEE Turkey Section.



KORKUT KAAAN TOKGOZ (Member, IEEE) received the B.Sc. and M.Sc. degrees from the Electrical and Electronics Engineering Department, Middle East Technical University, Ankara, Türkiye, in 2009 and 2012, respectively, and M.Eng. and Ph.D. degrees from the Department of Physical Electronics, Tokyo Institute of Technology, Tokyo, Japan, in 2014 and 2018, respectively. From 2018 to 2019, he was a Senior Researcher/an Assistant Manager with NEC Corporation, Kana-gawa, Japan, where he was involved in 5G systems and fixed point-to-point wireless links. From 2019 to 2022, he was an Assistant Professor with Tokyo Institute of Technology. He is currently a Faculty Member with the Faculty of Engineering and Natural Sciences, Sabanci University, Istanbul, Türkiye. He is the Co-Founder and the CTO of Evrim Company Ltd., Yokohama, Japan. His research interests include analog/RF/millimeter-wave/sub-terahertz transceivers for wireless communications, low-power Edge-AI for monitoring systems, the IoT, sensors and systems, de-embedding, device characterization, high-power, and high-efficiency PAs for wireless systems. He was a recipient of several awards, scholarships, and grants, including the TUBITAK 2232-B International Fellowship for Early-Stage Researchers 2022, the Marie Skłodowska-Curie Actions Post-Doctoral Fellowship 2022, the SSCS Predoctoral Achievement Award, in 2018, the IEEE MTT-S Graduate Student Fellowship, in 2017, the IEICE Student Encouragement Prize, in 2017, the Seiichi Tejima Overseas Student Research Award, and the IEEE/ACM ASP-DAC University LSI Design Contest 2017 Best Design Award.



MEHMET UNLU (Senior Member, IEEE) received the B.Sc., M.Sc., and Ph.D. degrees in electrical and electronics engineering from Middle East Technical University (METU), Ankara, Türkiye, in 2001, 2003, and 2009, respectively.

He was a Postdoctoral Researcher at METU-MEMS Center, in 2009. He was a Research Scientist at Cornell Nanoscale Science and Technology Facility (CNF), Ithaca, NY, USA, from 2009 to 2011, while he was working with Utah State University, Logan, UT, USA. He joined the University of Michigan, Ann Arbor, MI, USA, in 2011, as a Postdoctoral Research Fellow. From 2012 to 2018, he was a Faculty Member with Ankara Yıldırım Beyazıt University. He is currently a Faculty Member with the TOBB University of Economics and Technology, Ankara, Türkiye. His major research interests include terahertz components, plasmonics, optoelectronics, NEMS, and reconfigurable circuits and antennas for microwave, millimeter-wave, and terahertz applications. He received the Scientific and Technological Research Council of Turkey (TÜBİTAK) Ph.D. Scholarship, from 2003 to 2009, the Prof. Mustafa N. Parlar Education and Research Foundation-METU Thesis of the Year Award, in 2009, for his Ph.D. dissertation, the Best Student Paper Award (2nd Place) in the IEEE Antennas and Propagation Symposium (AP S), and the Best Student Paper Award (3rd place) in the IEEE International Microwave Symposium (IMS), in 2013, the Prof. Mustafa N. Parlar Education and Research Foundation Research Encouragement Award, in 2013, the International Union of Radio Science (URSI) Young Scientist Award, in 2014, and Turkish Academy of Sciences Outstanding Young Scientist Award, in 2015.

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