Uniaxial Dielectric Waveguide Filter Design Accounting for Losses Using Mode Matching Technique

Luke Murphy [1], Joseph Mautz [1], Mohsen Yazdani [1], Ercument Arvas [1], and Samir Tozin [2]

[1] Department of Electrical Engineering and Computer Science
Syracuse University, Syracuse, NY 13244, USA
lemurphy@syr.edu, jmautz@syr.edu, myazdani@syr.edu, earvas@syr.edu

[2] Deptartment of Electrical Engineering
Ajman University of Science and Technology, Ajman, UAE
s.tozin@ajman.ac.ae

Abstract **─** Dielectric filters can provide compact solutions for filter design problems. However, most dielectrics exhibit uniaxial properties, as well as, losses that will undoubtedly affect performance if not accounted for. This paper derives dispersion relations for lossy uniaxial media in dielectric waveguides and also accounts for lossy conducting walls. The waveguide discontinuity problem in the presence of lossy uniaxial media and finite conductivity waveguide walls, is calculated by mode matching technique and the results are applied to a Ka band filter. The design specifications for the proposed filter are a 32.5 GHz center frequency with 6%. Good agreement between simulated and measured results are shown.

Index Terms **─** Band-Pass Filter (BPF), Computer-Aided Design (CAD), dielectric waveguide filters, microwave filters, Mode-Matching Technique (MMT) and uniaxial media.

I. INTRODUCTION

Waveguide filters offer far superior performance to microstrip filters and dielectric filled waveguide filters significantly reduce waveguide filter size without sacrificing filter performance [1]. In addition to high performance, low manufacturing cost and small size, dielectric waveguide filters allow for flip-chip bonding; which makes for easy integration in millimeter-wave systems [2-3]. At millimeter-wave frequencies, dielectric waveguide filters also circumvent radiation loss; which is common in planar filters [4]. Recently, [5-6] use mode matching/hybrid method in the analysis of waveguide class filter problems. Wexler [7] first mentioned the advantages of choosing the Mode Matching Technique (MMT) versus all other methods when solving a class of waveguide problems. In [8-10], the MMT is applied to analyze dielectric waveguide and SIW filters as well as couplers. However, the MMT has never been applied to a lossy uniaxial dielectric waveguide filter with non-PEC walls. This paper will use the MMT to design and simulate a dielectric waveguide filter that is manufactured on a lossy uniaxial media with finite conductivity in the waveguide walls. The filter is designed in the Ka band, due to interest expressed by the manufacturing company that graciously manufactured and measured the filter free of charge.

II. DISPERSION RELATION IN LOSSY UNIAXIAL MEDIA

In order to account for a lossy dielectric in the MMT dispersion relations, Maxwell's equations must be modified. From [11], one finds Maxwell's equations for a conducting media. In this paper, it is assumed the relative permittivity tensor $\overline{\overline{\varepsilon_r}}$ takes the form of:

$$\overline{\overline{\varepsilon_r}} = \begin{bmatrix} \varepsilon & 0 & 0 \\ 0 & \varepsilon_y & 0 \\ 0 & 0 & \varepsilon \end{bmatrix}. \quad (1)$$

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Replacing the relative permittivity tensor in (1) with that of the relative permittivity tensor for a lossy uniaxial media:

$$\overline{\overline{\varepsilon_{cd}}} = \begin{bmatrix} \varepsilon - j\tau\varepsilon & 0 & 0 \\ 0 & \varepsilon_y - j\tau\varepsilon_y & 0 \\ 0 & 0 & \varepsilon - j\tau\varepsilon \end{bmatrix}. \quad (2)$$

Putting (2) into Maxwell's equation, multiplying by the dot product of $\left(\bar{\bar{\varepsilon}}_{cd}^{-1}\right)^{-1}$ and then taking the curl, we obtain:

$$\nabla \times \left(\bar{\bar{\varepsilon}}_{cd}^{-1} \cdot (\nabla \times H)\right) = j\omega\varepsilon_0(\nabla \times E). \quad (3)$$

Substituting the $\nabla \times E$ term of Maxwell's equation into (3) gives:

$$\nabla \times \left(\bar{\bar{\alpha}}_{cd} \cdot (\nabla \times H)\right) = j\omega\varepsilon_0(-j\omega\mu_0 H), \quad (4)$$

where $\bar{\bar{\alpha}}_{cd} = \bar{\bar{\varepsilon}}_{cd}^{-1}$. Assuming propagation is in the $\hat{z}$ direction and solving for the k_z of the H_z components provide the dispersion relation. After some tedious but straight forward calculations, the $\hat{z}$-components of (4) are given by:

$$\left(\alpha_{22}\frac{\partial^2}{\partial x\partial z} - \alpha_{23}\frac{\partial^2}{\partial x\partial y} - \alpha_{12}\frac{\partial^2}{\partial y\partial z} + \alpha_{13}\frac{\partial^2}{\partial y^2}\right)H_x + \left(-\alpha_{12}\frac{\partial^2}{\partial x\partial z} + \alpha_{23}\frac{\partial^2}{\partial x^2} + \alpha_{11}\frac{\partial^2}{\partial y\partial z} - \alpha_{13}\frac{\partial^2}{\partial x\partial y}\right)H_y + \left(\alpha_{12}\frac{\partial^2}{\partial x\partial y} - \alpha_{22}\frac{\partial^2}{\partial x^2} - \alpha_{11}\frac{\partial^2}{\partial y^2} + \alpha_{12}\frac{\partial^2}{\partial x\partial y}\right)H_z = k_o^2 H_z. \quad (5)$$

It can be seen that the coupling effect of H_x and H_y makes the calculation of k_z difficult for uniaxial media. Later in this paper, MMT is carried out to analyze the TE_{10} mode. This assumption enforces the conditions on (5) such that $H_y = E_x = k_y = 0$. Under these conditions, using Gauss's law and knowing from calculations that $\alpha_{23} = \alpha_{12} = \alpha_{12} = 0$, simplifies (5) to:

$$-\alpha_{22}\frac{\partial^2}{\partial z^2}H_z - \alpha_{22}\frac{\partial^2}{\partial x^2}H_z - \alpha_{11}\frac{\partial^2}{\partial y^2}H_z = k_o^2 H_z. \quad (6)$$

Under the assumption in a waveguide, H_z is of the form $\cos(k_x x)\cos(k_y y)\,e^{-jk_z z}$, solving for k_z in (6) provides the dispersion relation for lossy media:

$$k_z = j\sqrt{k_x^2 - k_0^2/\alpha_{22}} = j\sqrt{k_x^2 - \left(\varepsilon_y - j\tau\varepsilon_y\right)k_0^2}, \quad (7)$$

where $k_x = m\pi/a$ and a is the x dimension of the dielectric waveguide in Fig. 1.

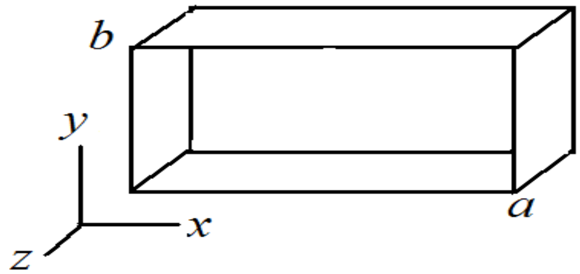

Fig. 1. Rectangular waveguide dimension reference.

III. LOSS IN CONDUCTOR WALLS

Most MMT calculations of the dielectric waveguides assume perfect conducting walls. In practice however, metallic walls exhibit a finite conductivity, σ_c and therefore cause attenuation of the signal. Under the assumption of TE_{10} mode of propagation, Kong's [12] perturbation method is used to calculate the attenuation for the waveguide wall dimensions that are defined in Fig. 1:

$$\alpha_c = \frac{P_w}{2P_f}, \quad (8)$$

where P_w represents the time-average power loss in the walls and P_f is the time-average power flowing through a cross section of the waveguide. P_f and P_w are defined as:

$$P_f = \frac{1}{2}Re\{\iint \boldsymbol{E} \times \boldsymbol{H}^*\,dx\,dy\}, \quad (9)$$

$$P_w = 2P_{w,x=0} + 2P_{w,y=0}, \quad (10)$$

where $P_{w,x=0}$ and $P_{w,y=0}$ is the power loss on the conducting walls at x=0 and y=0, respectively. The factor of two arises because all four walls must be considered. $P_{w,x=0}$ and $P_{w,y=0}$ are calculated by integrating the square of the current density on the waveguide wall multiplied by the surface resistance over the length of the wall, namely:

$$P_{w,x=0} = \int_0^b J_s(x=0)\cdot J_s^*(x=0)R_s\,dy, \quad (11)$$

$$P_{w,y=0} = \int_0^a J_s(y=0)\cdot J_s^*(y=0)R_s\,dx, \quad (12)$$

where $J_s = \hat{n} \times H$ and

$$R_s = \sqrt{\frac{\omega\mu_0}{2\sigma_c}}. \quad (13)$$

The assumption of only TE_{10} of mode propagation states only the existence of:

$$E_y = E_0 \sin\frac{\pi x}{a} e^{-jk_z z}, \quad (14)$$

$$H_x = Y_0 E_0 \sin\frac{\pi x}{a} e^{-jk_z z}, \quad (15)$$

$$H_z = Y_0 E_0 \cos\frac{\pi x}{a} e^{-jk_z z}, \quad (16)$$

in the waveguide, where E_0 is a wave amplitude coefficient and Y_0 is the admittance that relates the H-field amplitude to the E-field amplitude. Using (14)-(16) in (9) and (10) provides the attenuation constant, due to finite conductivity in the waveguide walls as:

$$\alpha_c = \frac{4Y_0 R_s(a+b)}{ab}. \quad (17)$$

It is important to note again that this attenuation constant is only valid for dominant TE_{10} mode analysis.

IV. MODE MATCHING FORMULATION

Figure 2 shows a waveguide step discontinuity. Assuming only TE_{m0} modes propagating in the waveguide, one can express the electric and magnetic fields in region A as a sum of the incident and reflected waves from the junction at z=0:

$$E_y^A = \sum_{m=1}^{M} A_m^+ \sin\frac{m\pi x}{a} e^{-\gamma_{am} z} + A_m^- \sin\frac{m\pi x}{a} e^{\gamma_{am} z}, \quad (18)$$

$$H_x^A = \sum_{m=1}^{M} Y_{am} A_m^- \sin\frac{m\pi x}{a} e^{\gamma_{am} z} - Y_{am} A_m^+ \sin\frac{m\pi x}{a} e^{-\gamma_{am} z}. \quad (19)$$

In region B, the fields are expressed as:

$$E_y^B = \sum_{n=1}^{K} B_n^+ \sin\frac{n\pi x}{c} e^{-\gamma_{bn} z} + B_n^- \sin\frac{n\pi x}{c} e^{\gamma_{bn} z}, \quad (20)$$

$$H_x^B = \sum_{n=1}^{K} Y_{bn} B_n^- \sin\frac{n\pi x}{c} e^{\gamma_{bn} z} - Y_{bn} B_n^+ \sin\frac{n\pi x}{c} e^{-\gamma_{bn} z}, \quad (21)$$

where in Fig. 2, A_m^+, B_m^- are unknown incident wave amplitude coefficients and A_m^-, B_m^+ are the reflected wave amplitudes in their respective region. In (14)-(16), the propagation constant for a respective region is:

$$\gamma_z = \alpha_c + jk_z, \quad (22)$$

where α_c and k_z are defined by (17) and (7), respectively, and

$$Y_i = \frac{k_z}{j\omega\mu_0}. \quad (23)$$

The propagation constant (k_z) and admittance (Y_i) are the two calculated values used in determining the unknown wave amplitude coefficients of the TE_{10} mode, because the MMT equations are only valid at z=0 and $0 \le x \le c$, they account for dielectric and conductor loss of wave amplitudes at this finite location. The propagation constant of (22) is also used later on to account for losses in the connecting sections of waveguides for the TE_{10} mode as well. Setting the tangential fields at z=0 equal to each other, assuming TE10 excitation and making use of mode orthonogality, one can calculate the transmission coefficients:

$$S_{21} = \left(\sum_{v=1}^{K}\left(\sum_{m=1}^{M}\left(\frac{2}{a} Y_{am} H_{mv} H_{mp}\right) + Y_{bv}\frac{c}{2}\delta_{vp}\right)\right)^{-1} (2Y_{a1}H_{1v})\ for\ p = 1,2,\ldots K. \quad (24)$$

The reflection coefficients are given by:

$$S_{11} = \frac{2}{a}\sum_{p=1}^{K}\left(H_{mp} S_{21}\right) - \delta_{m1} \quad \text{for m} = 1,2,\ldots M, \quad (25)$$

where

$$H_{ij} = \int_0^c \sin\frac{i\pi x}{a} \sin\frac{j\pi x}{c} dx, \quad (26)$$

where S_{21} are the transmission coefficients of the discontinuity, S_{11} are the reflection coefficients of the discontinuity, M are the number of modes in region A and K are the number of modes in region B. The reader is invited to explore [13] on MMT, in order to gain a full understanding on MMT formulation.

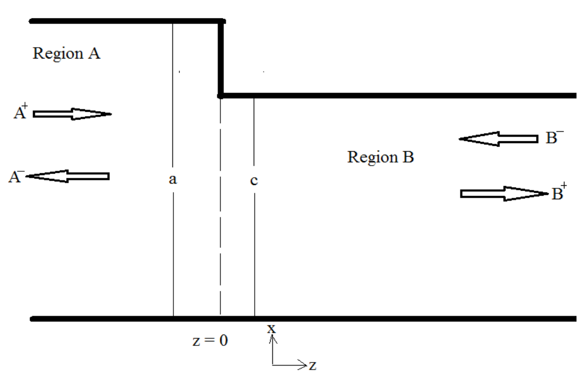

Fig. 2. Waveguide step discontinuity between region A and Region B.

V. FILTER DESIGN

A three pole filter with 6% bandwidth in the Ka band is designed using a dielectric with ε=67 and ε_y=61, due to expressed interest from the filter manufacturer. The filter is designed using Hong and Lancaster's [14] method to calculate filter coupling coefficients (k_{ij}) from g values.

This method provides coupling coefficients $k_{12}=k_{23}=0.107$. Using the resonant peak method of [14], coupling coefficients are then calculated and compared with ideal coupling coefficients to provide initial filter dimensions. Typically, the inductive coupling irises between resonators are kept thin with respect to waveguide dimensions, as shown by [15]. Bearing this in mind, Fig. 3 shows calculated coupling coefficients assuming the thickness (*d*) of the coupling irises is set to the minimum that manufacturing tolerances allow. One can see from Fig. 3, that if the dielectric is not treated as uniaxial, the coupling coefficients vary drastically. Using the results of Fig. 3 as a starting point and S-parameter theory to cascade the discontinuities with respective connecting waveguide sections, the filter dimensions are given in Fig. 4 after the first and third resonator lengths were optimized for return loss.

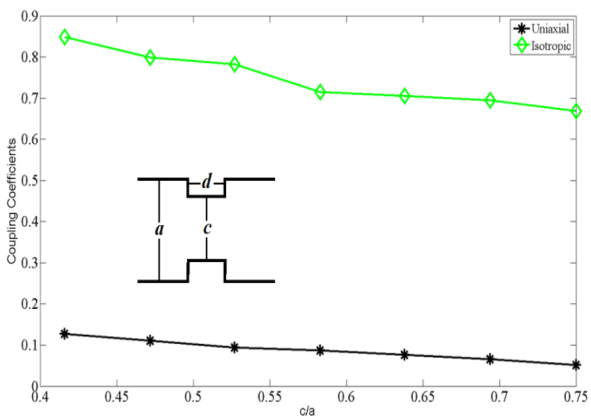

Fig. 3. Coupling coefficients for isotropic and uniaxial media for varying ratios of c/a and d=0.127 mm. Insert displays the junction and coordinate system under analysis.

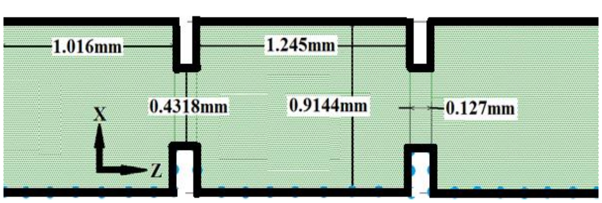

Fig. 4. Dimensions of initial filter.

VI. RESULTS

A narrowband waveguide filter is designed and simulated at 32.5 GHz. The dielectric is assumed to have a thickness of 0.254 mm, τ = 3e-4 and σ_c=3.5e7. The waveguide is excited with a TE_{10} mode using an optimized microstrip-to-waveguide transition, as was done in [16]. The exact dimensions of the transition are omitted at the request of the manufacture. The post-optimization results provided by the MMT are in Fig. 5, with measured filter performance in solid circles and results from a commercial FEM solver are also given for comparison (solid blue). The measured results of Fig. 5 show a VSWR of 2:1 or better from 31.6-33.8 GHz, with a center frequency of 32.68 GHz. The insertion loss is 6 dB or better in this band, with the passband ripple better than 1 dB peak-to-peak. It should be noted that according to the manufacturer 6 dB of insertion loss is acceptable. Figure 6 shows the manufactured filter above the word liberty on a US penny for size comparison. One can see good agreement between MMT and measured results. After analysis, it was discovered that the dimensions of the filter varied slightly from those in Fig. 4. These dimensional changes are noted in Fig. 6. It was also discovered that the donated substrate had higher dielectric loss and lower conductivity than initially estimated. However, when the adjusted losses and dimensional differences are accounted for in the MMT analysis, one can notice that there is now excellent agreement in Fig. 7 between the MMT and measured results.

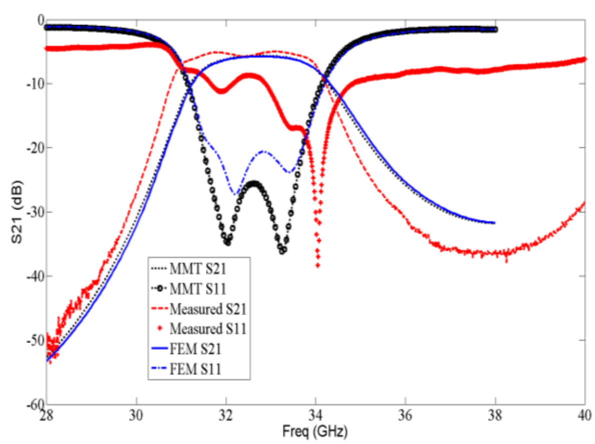

Fig. 5. Measured filter performance vs. simulated performance in black triangles.

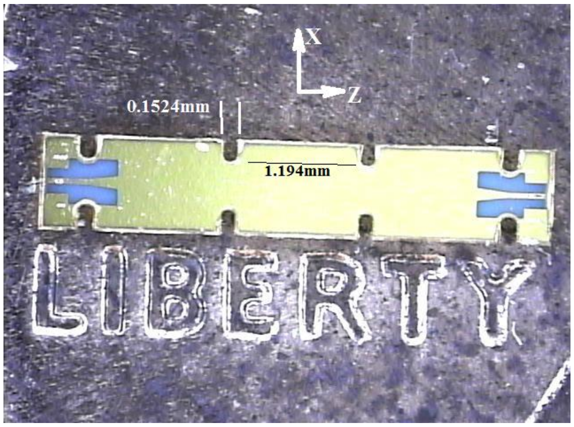

Fig. 6. Manufactured filter next to the word LIBERTY on a US penny for size comparison.

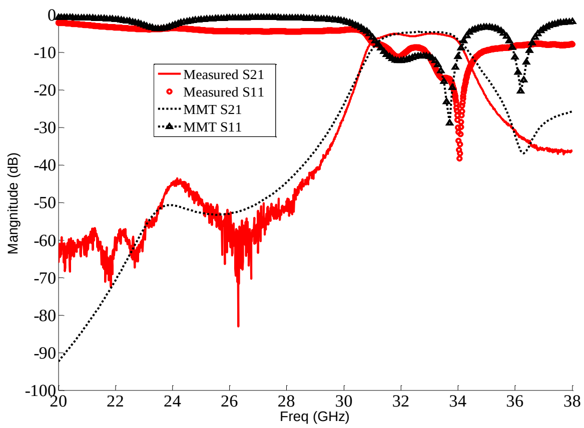

Fig. 7. Measured filter performance (solid circles) vs. simulated performance with adjusted dimension and loss numbers (black triangles).

VII. CONCLUSION

Derivations are presented for the propagation constant of a dielectric waveguide when the dielectric is assumed to exhibit uniaxial properties. Using this propagation constant with the MMT, a filter is designed, analyzed and manufactured. The MMT accounts for dielectric losses and conductor wall losses. It is shown that accounting for uniaxial media and losses in the analysis of dielectric filters using the MMT allows for accurate prediction of filter performance. The manufactured dielectric filter shows excellent agreement with simulated results.

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REFERENCES

[1] C. Y. Change and W. C. Hsu, "Novel planar, square-shaped, dielectric-waveguide, single-, and dual-mode filters," *IEEE Trans. on Microwave Theory and Tech.*, vol. 50, no. 11, pp. 2527-2536, 2002.

[2] A. C. Kundu, "Broadband TEM-mode planar-rectangular dielectric waveguide bandpass filter and its miniaturization," *IEEE MTT-S Int. Microwave Symp. Dig.*, Phoenix, AZ, pp. 381-384, 2002.

[3] M. Ito, et al., "A 60-GHz-band planar dielectric waveguide filter for flip-chip modules," *IEEE Trans. on Microwave Theory and Tech.*, vol. 49, no. 12, pp. 2431-2436, 2001.

[4] H. C. Chen, et al., "All-planar dual-mode asymmetric filters at ka-band," *IEEE Microwave and Wireless Comp. Letters*, vol. 12, no. 3, pp. 111-113, 2003.

[5] S. Liu, et al., "Rectangular conducting waveguide filled with uniaxial anisotropic media: a modal analysis and dyadic green's function," *PIER*, vol. 25, pp. 111-129, 2000.

[6] P. Grassi, et al., "Analysis of inhomogeneously filled waveguide devices by a hybrid mode matching-finite element technique," *IEEE AP-S Int. Symp.*, vol. 2, pp. 161-164, June 2003.

[7] A. Wexler, "Solution of waveguide discontinuities by modal analysis," *In IEEE Trans. on MTT*, vol. 15, no. 9, pp. 508-517, September 1967.

[8] R. Rezaiesarlak, et al, "Hybrid of moment method and mode matching technique for full-wave analysis of SIW circuits," *ACES Journal*, vol. 26, no. 8, pp. 688-695, 2011.

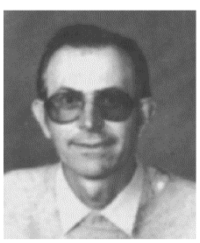
[9] Z. Kordiboroujeni, et al, "Mode-matching design of substrate-integrated waveguide couplers," *Symps. on Electromag. Compatibility, Asia-Pacific*, pp. 701-704, 2012.

[10] J. Bornemann and F. Taringou, "Substrate-integrated waveguide filter design using mode-matching techniques," *Proc. of 41st European Microwave Confer.*, pp. 1-4, October 2011.

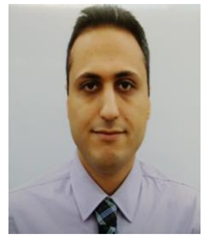
[11] R. F. Harrington, "Time-harmonic electromagnetic fields," *McGraw-Hill*, New York, 1961.

[12] J. A. Kong, "Electromagnetic wave theory," *EMW Publishing*, Massachusetts, 2005.
[13] T. Itoh, "Numerical techniques for microwave and millimeter-wave passive structures," 1st ed., *New York: Wiley-Interscience*, 1989.
[14] J. H. Hong and M. J. Lancaster, "Microstrip filters for RF/microwave applications," 1st ed., *New York: John Wiley & Sons, Inc.*, pp. 40-130, 2001.
[15] F. Arndt, et al, "Modal s-matrix method for the optimum design of inductively direct-coupled cavity filters," *IEE Proc. Micro., Ant. and Prop.*, vol. 133, no. 5, pp. 341-350, 1986.
[16] Z. Sotoodeh, et al, "A novel bandpass waveguide filter structure in SIW technology," *PIER*, vol. 2, pp. 141-148, 2008.

Luke Murphy received his B.Sc. degree from Rensselaer Polytechnic Institute, Troy, NY in 2007. In 2009 he earned his M.Sc. degree from Syracuse University, Syracuse, NY and is currently completing his Ph.D degree at Syracuse University. From 2010 he has interned and worked at Dielectric Laboratories, Inc. in Cazeonvia, NY designing RF filters and other passive RF components. His research interests are in electromagnetic theory, metamaterials and microwave devices.



Joseph R. Mautz was born in Syracuse, NY, in 1939. He received his B.Sc., M.Sc. and Ph.D. degrees in Electrical Engineering from Syracuse University in 1961, 1965 and 1969, respectively. Until July 1993, he was a Research Associate with the Electrical Engineering and Computer Science Department, Syracuse University, where he worked on radiation and scattering problems. Currently, he is affiliated with the Electrical Engineering and Computer Science Department at the same university. His primary fields of interest are electromagnetic theory and applied mathematics.



Mohsen Yazdani was born in Tehran, Iran. He received his Masters degree from Iran University of Science and Technology, Tehran, Iran. He earned his Ph.D. degree at Syracuse University, Syracuse, NY in 2014. He is currently working as a Post-Doc at UCLA, Los Angeles, CA. His research interests include electromagnetic scattering, antennas and microwave devices.

Ercument Arvas received his B.Sc. and M.Sc. degrees from the Middle East Technical University, Ankara, Turkey in 1976 and 1979, respectively, and the Ph.D. degree from Syracuse University, Syracuse, NY in 1983; all in Electrical Engineering. From 1984 to 1987, he was with the Electrical Engineering Department, Rochester Institute of Technology, Rochester, NY. In 1987, he joined the Electrical Engineering and Computer Science Department, Syracuse University, where he is currently a Professor in the Electrical Engineering and Computer Science Department. His research and teaching interests are in electromagnetic scattering and microwave devices.

Samir L. Tozin earned a B.S.E.E. degree from Seattle University in 1995 and his Ph.D. degree from Syracuse University in 2004. From 1997 to 2009, he was with Anaren Microwave in East Syracuse, New York. From 2002 to 2009, he taught Microwave Engineering courses at Syracuse University. He is currently working as an Assistant Professor at Ajman University of Science and Technology in the United Arab Emirates. His current interests are in the research and development of passive and active microwave devices as well as antennas.