Wide band Polarization Detector for Sub–6 GHz 5G Applications

Eman M. Eldesouki¹, Shimaa A. M. Soliman¹, Sherine. I. Abd El-Rahman¹, and Ahmed M. Attiya¹, (Member, IEEE)

¹ Microwave Engineering Department, Electronics Research Institute (ERI), Joseph Tito St., Huckstep, El Nozha, Cairo 11843, Egypt.

Corresponding authors: Eman M. Eldesouki (eman@eri.sci.eg) and Shimaa A. M. Soliman (shimaa_megahed@eri.sci.eg).

ABSTRACT This paper presents the analysis and the design of a polarization detector for sub–6 GHz 5G applications. The proposed design operates in the frequency range from 2 to 4 GHz to cover the 5G NR frequency bands n40/n41/n78. The proposed polarization detector consists of a six-port network with two input and four output ports. The two input ports of this six-port network are connected to two identical orthogonal linearly polarized wideband antennas; vertical and horizontal. Two orthogonal tapered slot antennas are designed for this purpose. On the other hand, the other four outputs of the six-port network are connected to power amplitude detectors. These four power magnitudes are used to determine the ratio of the vertical to horizontal components of the field incident on the polarization detector. They are also used to determine the phase difference between the vertical and horizontal components. Based on the relative magnitude and the phase difference between the horizontal and vertical components of the incident field, the polarization parameters of the incident field are obtained; like the tilt angle, the axial ratio and the direction of rotation. To obtain the phase difference between the two inputs, a reference wideband differential phase shifter is included in the six-port network. This wideband differential phase shifter is implemented by using a modified Schiffman phase shifter. From the experimental results, the proposed polarization detector proved its reliability in detecting the polarization and discriminating the direction of the wave rotation; left or right hand rotation.

INDEX TERMS Polarization Detector, Six-port Network, Schiffman Phased shifter.

I. INTRODUCTION
Polarization is an important parameter of electromagnetic waves. Since, polarization is not correlated with intensity or spectral information of the electromagnetic wave, the knowledge of the polarization increases the energy efficiency and reduce power loss at the receiver [1]. Thus, polarization detection is quite important in many applications like radar [2], microwave imaging [3], [4], and wireless communication [2]–[6]. Measuring polarization in a direct way is difficult due to the complexity of straightforward methods. Several methods are introduced in literature to obtain the electromagnetic wave polarization such a linear component method (LCM) [7], [8], multiple amplitude component method (MACM) [9] and the three-antenna method (TAM) [10]. The common disadvantages of these methods are complexity, time-consuming and they require mechanical movement for multiple antennas and structures. A new technique for polarization detection by using passive circuit composed of cascaded Wilkinson power dividers (WPD) was introduced in [11]. This method uses amplitude measurements to determine the polarization features of antenna under test (AUT). The main disadvantage of this method is that it is applicable for only a single frequency.

The present paper introduces a simple polarization detector to determine the polarization of the incident field in a wide frequency range from 2 to 4 GHz. This frequency range covers the entire 5G NR frequency bands n40 (2.3–2.4) GHz, n41 (2.496–2.690) GHz and n78 (3.3–3.8) GHz [12]. The proposed polarization detector consists of a six-port network [13]–[15] with two input ports and four output ports. The two input ports are connected to two identically orthogonal linearly polarized wide band antennas to detect the horizontal and vertical components of the incident field. Two output ports are dedicated to introduce output signals proportional to the magnitudes of the received horizontal and vertical components of the incident field. The third output port introduces an output signal proportional to the magnitude of the summation of these two components. Finally, the fourth output introduces an output signal proportional to the magnitude of the summation of these two components after adding a reference phase shift 90° between these two components. These four outputs are treated together to obtain the relative amplitude and the phase...
difference between the horizontal and the vertical components of the incident field. Thus, the other polarization parameters can be obtained directly, like the tilt angle, the axial ratio and the direction of rotation of the polarization. The main advantages of this polarization detector are it is based on magnitude only measurements and there is no need to rotate or replace the measuring antenna or the detector circuit [16], [17]. On the other hand, the key points in this polarization detector are the wide band linearly polarized antennas and the wideband reference differential phase shifter. In the present design, the wideband linearly polarized antenna is implemented by using a tapered slot antenna [18] and the wideband differential phase shifter is implemented by using a modified Schiffman phase shifter [19]-[21].

In this study, two experimental setups are established. The first setup is carried out to verify that the designed polarization detector circuit (PDC) can extract the phase difference between two input signals. The second setup uses two identical linearly–polarized tapered slot antennas connected in the input ports of the PDC. The two antennas are arranged to receive the horizontal and vertical components of the field radiated by the AUT. Three antennas are used independently as AUT; dual circularly–polarized horn antenna model LB-OSJ-20180-P03 [22], dual circularly–polarized U–shaped slot antenna [23] and a linearly–polarized antenna. Based on the amplitude measurements from the four outputs of the PDC the polarization features of the AUT are obtained.

This paper is organized as follows. Section II presents the analysis of the proposed six-port network and how it can be used to determine the polarization parameters of the incident field. Section III presents the details for the analysis of Schiffman phase shifter which represents one of the main critical parts in the proposed six-port network. On the other hand, Section IV presents the details for the analysis of the wideband linearly polarized antenna which is used as inputs for the six-port network. Section V presents the implementation and experimental verification of the proposed six-port network separately. Then Section VI presents the experimental results for polarization detection. These experimental include three different antennas; a commercial tapered horn dual circularly polarized antennas, a dual circularly polarized slot antenna and a linearly polarized antennas. Finally, Section VII presents the concluding remarks.

II. ANALYSIS OF A SIX-PORT NETWORK FOR POLARIZATION DETECTION
The schematic diagram of the proposed six-port network for polarization detection is shown in Fig. 1. It consists of four power dividers (PD), two power combiners (PC) and a differential phase shifter. Ports 1 and 2 represent the two input ports which correspond to the horizontal and vertical components of the incident field; \( E_h \angle \phi_h \) and \( E_v \angle \phi_v \). The magnitude of the output at Port 3 is directly proportional to the magnitude of the input signal at Port 1. Similarly, the magnitude of the output at Port 4 is directly proportional to the magnitude of the input signal at Port 2 as follows:

\[
E_3 = E_h S_{31} + O(E_v S_{32}), \quad (1.a)
\]

\[
E_4 = E_v S_{42} + O(E_h S_{41}), \quad (1.b)
\]

where \( S_{mn} \) is the complex transmission coefficient between Ports \( m \) and \( n \). The term \( O(\cdot) \) corresponds to the error due to the finite isolation in \( S_{32} \) and \( S_{41} \). For an isolation level more than 30 dB, these error terms can be ignored compared to the direct path signals \( S_{31} \) and \( S_{42} \).

For equal division ratios for all power dividers, the proportionality constant of Port 3 to Port 1 would be the same as in Port 4 to Port 2 which would be nearly 1/2 in the magnitude of the electric field. This corresponds to \(-6\) dB in terms of the power; assuming that the losses in power division is negligible and the power dividers have equal power division. Thus, \( |S_{31}| = |S_{42}| \cong 1/2 \) and the ratio of the magnitude of horizontal to the vertical components of the incident wave would equal to the ratio of the output of Port 3 to the output of Port 4 as follows:

\[
\frac{|E_h|}{|E_v|} = \frac{|E_3|/|S_{31}|}{|E_4|/|S_{42}|}. \quad (2)
\]

On the other hand, the output at Port 5 of the first power combiner is given by:

\[
E_5 = E_h S_{51} + E_v S_{52}. \quad (3)
\]

For symmetric equal power combination, \( S_{51} = S_{52} = A \angle \phi_{51} \). Thus, the magnitude of this summation is proportional to the magnitude of the summation of the horizontal and vertical components of the incident field with their corresponding phase differences as follows:

\[
|E_5| = A|E_h \angle \phi_h + E_v \angle \phi_v|, \quad (4)
\]
where $A$ is the proportionality constant which equals the magnitude of $|S_{51}| = |S_{52}|$. For a Wilkinson power combiner with equal power ratios, the two input ports are connected to a common resistance $R$ equals twice the characteristic impedance of the ports’ transmission lines to introduce the required isolation between the two input ports. This isolation resistance introduces a power loss of 3 dB. Thus, the received power from the two input ports is $-6$ dB of the input powers. Hence, the proportionality factor $A$ would be $1/2$; in terms of the magnitude of the electric field.

The magnitude of this summation can be simplified in terms of the phase difference between the horizontal and vertical components of the incident field as follows:

$$|E_5| = A|E_h| + |E_v|\Delta\phi,$$  

where $\Delta\phi = \phi_h - \phi_v$. By using simple trigonometric relations, it can be shown that:

$$|E_5| = A\sqrt{|E_h|^2 + |E_v|^2 + 2|E_h||E_v|\cos\Delta\phi}.$$  

By using (1) and (2), the magnitudes of $|E_h|$ and $|E_v|$ can be represented by $|E_h| = |E_3|/|S_{31}|$ and $|E_v| = |E_4|/|S_{42}|$ respectively. Thus, the phase difference $\Delta\phi$ can be obtained in terms of the measured three magnitude signals $|E_3|$, $|E_4|$ and $|E_5|$ and the magnitude of the S-parameters as follows:

$$\Delta\phi = \cos^{-1}\left(\frac{|E_5|^2 - (\alpha_1^2|E_3|^2 + \alpha_2^2|E_4|^2)}{2\alpha_1\alpha_2|E_3||E_4|}\right),$$  

where $\alpha_1 = |S_{51}|/|S_{31}|$, $\alpha_2 = |S_{52}|/|S_{42}|$.

It should be noted that the $\cos^{-1}(\cdot)$ is a multivalued function. Thus, the obtained phase difference in (7) can be either positive or negative value. To specify the exact value of the phase difference, an additional summation with pre-specified additional phase shift is required. This additional phase shift is introduced by using a differential phase shifter as shown in Fig. 1. This additional phase shift is assumed to be $90^\circ$. The summation of the two signals after the differential phase shifter is obtained at the output of the second power combiner in Fig. 1. By following the same steps in (4)-(6), the phase difference after the differential phase shifter is given by:

$$\Delta\phi - 90^\circ = \cos^{-1}\left(\frac{|E_6|^2 - (\alpha_3^2|E_3|^2 + \alpha_4^2|E_4|^2)}{2\alpha_3\alpha_4|E_3||E_4|}\right),$$  

where $\alpha_3 = |S_{51}|^2/|S_{31}|^2$, $\alpha_4 = |S_{52}|^2/|S_{42}|^2$.

Based on the two values obtained from (7) and the other two values obtained from (8), it would be possible to determine the exact sign of $\Delta\phi$ in (7), which is the common value for both solutions.

After determining the ratio of the magnitude of the horizontal component to the vertical component in (2) and the phase difference between these two components in (7) and (8), it would be possible to obtain the details of the polarization of the incident wave. In general, the incident wave can be represented as an elliptical wave as shown in Fig. 2. The main parameters of the elliptical polarization are the axial ratio, the tilting angle and the direction of rotation; right or left hand rotation. For a special case of circular polarization the axial ratio would be unity in magnitude and for linear polarization the axial ratio would be infinity in magnitude.

![FIGURE 2. Parameters of general elliptical polarization.](image)

The AR can be determined from the amplitude measurements as the ratio between major and minor axes of the polarization ellipse as follows:

$$AR = 10\log_{10}\frac{OA}{OB}.$$

where $OA$ and $OB$ are the major and the minor axes, respectively, of an ellipse centered in the origin as shown in Fig. 2. The amplitudes of the major and the minor axes can be obtained from the magnitude of the S-parameters and the obtained phase difference as follows:

$$OA = 0.5 \sqrt{|E_3|^2 + |E_4|^2 + (|E_3|^4 + |E_4|^4) \cos 2\Delta\phi}^{0.5},$$  

$$OB = 0.5 \sqrt{|E_3|^2 + |E_4|^2 - (|E_3|^4 + |E_4|^4) \cos 2\Delta\phi}^{0.5}.$$

On the other hand, the tilt angle $\tau$ of the major axis w.r.t the x-axis, as shown in Fig. 2, can be obtained as follows:

$$\tau = 0.5 \tan^{-1}\left(\frac{2|E_3||E_4|}{|E_3|^2 - |E_4|^2}\right)\cos \Delta\phi.$$

Finally, the direction of the rotation of the polarization depends on the sign of $\Delta\phi$. Positive phase difference corresponds to Left Hand rotation while negative phase difference corresponds to Right Hand rotation.

### III. WIDEBAND SCHIFFMAN PHASE SHIFTER

The key point in the above design of the proposed six-port network is the differential phase shifter. Usually the power dividers and power combiners have wider bandwidth than a simple differential phase shifter based on direct delay lines with different lengths. Thus, it is required to design a wideband differential phase shifter for this purpose. A good choice for this differential phase shifter is the Schiffman phase shifter.

The schematic diagram of Schiffman phase shifter is shown in Fig. 3. It consists of two C-section coupled branches.
operating in the first and second Schiffman periods [21], respectively. Then, two cascaded transmission line section of characteristic impedance $Z_0 = 50\Omega$ are added at input and output ports of the coupled branch to attain impedance matching. The differential phase shift between the two Schiffman branches $\Delta \phi_{sh} = \phi_r - \phi_m$ is given by:

$$
\Delta \phi_{sh}(\theta) = \left( \begin{array}{c}
2(K_m - K_r) \\
- \cos^{-1} \left( \frac{\rho_r - \tan^2 \theta}{\rho_r + \tan^2 \theta} \right) \\
+ \cos^{-1} \left( \frac{\rho_m - \tan^2(2\theta)}{\rho_m + \tan^2(2\theta)} \right)
\end{array} \right),
$$

where $\theta$ is the electrical length and $K_r$ and $K_m$ are the length ratios for the reference and main branches. For the operation in first and second period with $\Delta \phi_0 = 90^\circ$, $K_r = 1 - \Delta \phi_0/\pi = 0.5$ and $K_m = 0$. $\rho_r$ and $\rho_m$ are the impedance ratios of the two coupled lines in the reference and main branches, respectively. The electrical length $\theta = 0.57f/f_0$. $f$ is an arbitrary frequency and $f_0$ is the center frequency at which $\theta = \theta_0 = 90^\circ$. The phase slope of $\Delta \phi_{sh}(\theta)$ is the derivative of equation (12) and it can be deduced as:

$$
d\Delta \phi_{sh}(\theta) = \left( \begin{array}{c}
\frac{2(K_m - K_r)}{2\sqrt{\rho_r (1 + \tan^2 \theta)}} \\
- \frac{\rho_r + \tan^2 \theta}{8\sqrt{\rho_m}} \\
+ \frac{\rho_m + (\rho_m - 1)\cos(4\theta)}{8\sqrt{\rho_m}}
\end{array} \right),
$$

where $\rho_r$ and $\rho_m$ can be expressed in terms of even/odd-mode characteristic impedances of the reference branch and the second branch as follows:

$$
Z_{0er} = Z_0\sqrt{\rho_r} \quad \text{and} \quad Z_{0or} = \frac{Z_0}{\sqrt{\rho_r}},
$$

$$
Z_{0em} = Z_0\sqrt{\rho_m} \quad \text{and} \quad Z_{0om} = \frac{Z_0}{\sqrt{\rho_m}}.
$$

Hence, the initial values of the design can be conveniently derived by (12)-(13). To design a $90^\circ$ phase shifter lines that cover frequency band $2 - 4$ GHz, the impedance ratios in the first and second phase periods are found to be $\rho_r = 1.2$ and $\rho_m = 2.2$. By solving equation (14); the even and odd mode characteristic impedances are calculated as $Z_{0er} = 77\Omega$, $Z_{0or} = 56.3\Omega$, $Z_{0em} = 80\Omega$ and $Z_{0om} = 36.5\Omega$. The physical dimensions of the corresponding coupled line sections with these impedances are obtained by using LINECAL tool in Advanced Design System (ADS). The used substrate is an FR4 substrate with relative permittivity equals 4.4 and dielectric thickness equals 1.5 mm. Fig. 4 shows the physical dimensions of the designed Schiffman differential phase shifter.

The response of the designed phase shifter is shown in Fig. 5. It can be noted that the reflection loss is below $-10$ dB on the entire operating band while the isolation between the two paths is greater than 25 dB and the insertion loss in each path is less than 0.3 dB. On the other hand, the differential phase shift between the two paths is $90^\circ \pm 6^\circ$ in the entire required frequency range.

IV. WIDEBAND LINEARLY POLARIZED ANTENNA
Other important elements in the proposed polarization detector are the two orthogonally linearly polarized antennas which are connected to its input ports. Tapered slot antennas represent good candidates for this purpose. Two identical tapered slot antennas are designed and fabricated. Fig. 6 shows the geometry of the designed antenna. This antenna is implemented on an FR-4 substrate with dielectric thickness 1.5 mm. The antenna consists of a tapered slot with periodic corrugations. These corrugations are added to reduce the fringing fields from the sides of the antenna. This tapered slot is excited by proximity coupled microstrip line on the other side of the slot. In order to understand the effect of these periodic corrugations on the antenna performance, the surface current distribution at the center frequency 3 GHz is shown in Fig. 7. It is observed that the surface current is concentrated near the inner edges of the tapered slots while less current is monitored in the corrugated region. The radiation pattern of this antenna in its two principal, E and H, planes at the center frequency 3 GHz is shown in Fig. 8. The total gain is 5.8 dB while the backward radiation is less than −10 dB. The frequency response of the total gain (at θ = 0°, φ = 0°) and radiation efficient of the linearly polarized antenna is shown in Fig. 9. The total gain value ranged from 3 to 6.5 dB while the computed radiation efficiency varied from 72 to 90% in the operating frequency band. On the other hand, Fig. 10 shows the input reflection coefficient $S_{11}$ and the coupling $S_{12}$ between the two orthogonal antenna arrangement. It can be noted that $S_{11}$ is less than −10 dB in the entire frequency band while the coupling coefficient is less than −30 dB.
V. IMPLEMENTATION AND VERIFICATION OF POLARIZATION DETECTOR CIRCUIT

In this section, the previous parts are connected together to implement the proposed polarization detector. The power dividers and the power combiners, discussed in Sec. 2, are implemented using Wilkinson power dividers/combiners. The simulation layout of the complete PDC is shown in Fig. 11. It should be noted that WPD is not the only choice for implementing this circuit. It can also be implemented by using hybrid coupler, rate race coupler or directional coupler. However, WPD would have the smallest size compared to these alternative choices. It should also be noted that the two input feeding lines to the Schiffman differential phase shifter from the left side are adjusted to have the same electrical length. Thus, the two input signals to the differential phase shifter would have the same amplitude and the same phase. Hence, the phase difference between the two outputs in the right side would be only due to the differential phase shifter. The designed PDC is fabricated as shown in Fig. 12.

The S-parameters of this circuit is shown in Fig. 13. Fig. 13-a shows the reflection coefficients of the different ports. It can be noted that the two input ports, Port 1 and 2 are well matched below $-10\,\text{dB}$ on the entire frequency band. The remaining ports show also good matching properties, except Port 6. This may be explained due to the effect of the Schiffman phase shifter. On the other hand, Fig. 13-b shows the required transmission coefficients from Ports 1 and 2 to the Ports 3, 4, 5 and 6. It can be noted that the transmission coefficients $S_{31}$, $S_{42}$, $S_{51}$ and $S_{52}$ are around $-6\,\text{dB}$ while $S_{61}$ and $S_{62}$ is around $-12\,\text{dB}$. These values agree with the expected values for the proposed circuit. The slight change from these average values can be explained due to the band width of the used WPD and the effect of the differential phase shifter on the magnitude of the transmission coefficients to the Port 6. Finally, Fig. 13-c shows the isolation between the ports to be isolated. It can be noted that the isolation between the two input ports 1 – 2 is more than 20 dB. Furthermore, the isolations between ports 1 – 4 and ports 2 – 3 are more than 35 dB up to 3.3 GHz. For higher frequency, they are still greater than 20 dB. These values for isolation are quite enough to ignore the cross error components as discussed in (2).

FIGURE 11. Geometry of the proposed PDC. $R$ is lumped resistance of $100\Omega$. Units: mm.

FIGURE 12. Fabricated PDC, (a) Front side, (b) Back side.

FIGURE 13. Measured S-parameters of the fabricated PDC, (a) Reflection coefficients, (b) Transmission coefficients and (c) Isolation coefficients between isolated ports.
To verify that the designed PDC has the ability to determine the phase difference between two field signals we introduced the experiment shown in Fig. 14. In this experiment, the input signal from a Rhode and Schwartz model ZVA67 VNA is splitted into two equal parts with the same phase by using a wideband resistive power splitter. The two outputs of the power splitter are carried out to the PDC through two identical coaxial cables. Then an additional coaxial adaptor is added to one of the two cables as shown in Fig. 14. This additional coaxial adaptor introduces a linear phase shift as function of the operating frequency assuming that the operating frequency band is much less than the cutoff frequency of the higher order modes in this coaxial adaptor. This phase shift is measured directly by using the VNA. The two signals are then applied on the two input ports of the designed PDC and the magnitude of the output signals at the four output ports are used to determine this phase shift by using (7) and (8).

VI. EXPERIMENTAL RESULTS OF POLARIZATION DETECTION

In this section, the PDC is combined with the two linearly polarized antennas to introduce the complete polarization detector. This polarization detector is used to measure the polarization of a commercial tapered horn dually circularly polarized antenna model LB-OSJ-20180-P03 [22], as shown in Fig. 16-b. This antenna can be switched between LHCP and RHCP by switching between its two input ports. The two input ports of the PDC are connected to the linearly polarized tapered slot antennas in orthogonal orientations as shown in Fig. 16-a. The four output signals are then used to determine the relative level of the horizontal to the vertical component and the phase difference between them to determine the polarization of the antenna under test. Fig. 17-a shows the retrieved phase difference for the two cases of the LHCP and RHCP. Fig. 17-b shows the calculated axial ratios in the two cases compared to the corresponding ones in the datasheet of the measured antenna. Good agreements with errors less than 1 dB are obtained by using the designed polarization detector circuit. Fig. 18 shows plot of the calculated tilt angle. This experiment proves the applicability of the designed polarization detector in the required frequency range from 2 to 4 GHz which covers the 5G NR frequency bands n40/n41/n78.

![FIGURE 14. Experimental setup to measure the relative phase shift by using the designed PDC.](image)

![FIGURE 15. Comparison between the measured phase shift and the calculated phase shift based on the magnitude-only measurements of the designed PDC.](image)

![FIGURE 16. (a) Setup of polarization determination. (b) Dual circular polarized horn antenna model LB-OSJ-20180-P03.](image)

Another experiment was developed to determine the polarization of a dual circularly polarized U-shaped slot antenna [23]. Fig. 19-a shows the measured $\Delta \phi$ between the vertical and horizontal components of the radiated fields of this antenna for the two different ports. Fig. 19-b shows the measured AR of this antenna compared to the simulated result. It can be noted the good agreement between the measured and the simulated results. Also, Fig. 20 shows the measured $\tau$ for the radiated fields in this case.
Finally, a third experiment was developed to determine the polarization of a linearly polarized antenna. In this case a Vivaldi antenna is used as AUT. Fig. 21 shows the measured AR in this case. It is quite clear in this case AR has a very large value more than +60 dB. This property corresponds to a polarization oriented mainly in one direction which the property of linear polarization. The measured tilt angle in this case was nearly 20° which corresponds to the physical tilting angle of the AUT with respect to the horizontal direction during the experiment.

VII. CONCLUSION
A polarization detector in the frequency band from 2 to 4 GHz is introduced in the paper. This PDC consists of a six-port network with two orthogonal linearly polarized wideband antennas. The relative amplitude and the phase difference between the horizontal and vertical components of the incident plane wave are obtained by using amplitude-only measurements. The relative phase shift is obtained by using a wideband Schiffman phase shifter. The design of the wideband linearly polarized antennas and wideband Schiffman phase shifter are discussed in detail. The designed polarization detector is used to measure polarization of three different antennas: dual circularly polarized taper horn antenna, dual circularly polarized slot antenna and linearly polarized Vivaldi antenna. The proposed polarization detector showed good agreement with the expected polarizations of these antennas. The designed polarization detector would be a useful tool for antenna measurements in the frequency range from 2 to 4 GHz which covers the 5G NR frequency bands n40/n41/n78.
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Eman M. El-Desouki received B.Sc., M.Sc., and Ph.D., Electronics and Electrical Communications, Faculty of Engineering, Benha University at 2003 at Cairo University at 2009, and at Ain Shams University at 2018, respectively. She is currently a researcher in Microwave Engineering Dept. in Electronics Research Institute (ERI). Her research interests include Antennas, Electromagnetic wave scattering from rough surfaces, Wave Propagation, Synthetic Aperture Radars (SAR), Polometric Radar Imaging, Filter Design, Microwave measurement techniques, Numerical techniques in electromagnetics, periodic structures, Reflectarray Design, Electronic devices and Surface Plasmon Polaritons.

Shimaa A. M. Soliman B.Sc and M.Sc., in Electrical, Electronics and Communications Engineering, Faculty of Engineering, Zagazig University at 2004, 2014, respectively. Received Ph.D. from AL Azhar University, Cairo, Egypt at 2020, joined Electronics Research Institute as a Researcher Assistant in 2016. She is currently a researcher in Microwave Engineering Dept. in Electronics Research Institute (ERI). Interested in Antennas, Electromagnetic, Synthetic Aperture Radars (SAR), Microwave measurement techniques, Numerical techniques in electromagnetics, Circuits, Electronic devices and Reflectarray Design.

Sherine. I. Abd El-Rahman B.Sc from Shoubra Faculty of Engineering, in 2003. M.Sc. from Al Azhar faculty of engineering in 2012 and Ph.D from Shoubra Faculty of Engineering in 2019. She is currently a researcher in Microwave Engineering Dept. in Electronics Research Institute (ERI). Her current research interests include microwave and millimeter-wave circuits, and MIMO antennas.

Ahmed M. Attiya M.Sc. and Ph.D., Electronics and Electrical Communications, Faculty of Engineering, Cairo University at 1996 and 2001 respectively. From 2002 to 2004 he was a Postdoc in Bradley Department of Electrical and Computer Engineering at Virginia Tech. From 2004 to 2005 he was a Visiting Scholar in Electrical Engineering Dept. in University of Mississippi. From 2008 to 2012 he was a Visiting Teaching Member in King Saud University. He is currently Full Professor and the Head of Microwave Engineering Dept. in Electronics Research Institute. His research interests include Electromagnetic waves, antennas and wave propagation, microwave passive circuits and systems, microstrip and planar circuits and antennas, antenna measurement techniques, microwave measurement techniques, UWB and short pulse signals, numerical techniques in electromagnetics, analytical techniques in electromagnetics, periodic structures, artificial electromagnetic materials, nanotechnology, carbon nanotubes, graphene, and plasmonics.