Periodic Leaky-Wave Antenna Array With Horizontally Polarized Omnidirectional Pattern

Kunpeng Wei, Zhijun Zhang, *Senior Member, IEEE*, Zhenghe Feng, *Fellow, IEEE*, and Magdy F. Iskander, *Fellow, IEEE*

***Abstract*—A novel periodic leaky-wave antenna for broadside radiation with horizontally polarized omnidirectional pattern is proposed, designed and demonstrated experimentally. The objective is to achieve a horizontally polarized omnidirectional antenna array with highly directive beam. The proposed structure is based on combining the rotating electric field method and a leaky-wave antenna design. By properly designing a unit cell with a pair of crossed dipoles spaced at a distance approximately equal to $\lambda_g/4$ at the center frequency, not only omnidirectional radiation pattern with horizontal polarization is obtained by crossed dipoles fed in phase quadrature, but also the open-stopband effect exhibited at broadside scan is significantly reduced. The analysis and design of this structure are performed on the basis of a simple equivalent network model. A finite structure formed by 16-unit cell (32-element) for broadside beam is developed at the 2.4 GHz band, which offers a horizontally polarized omnidirectional radiation pattern with enhanced gain of 9.9–10.6 dBi and measured radiation efficiency of 90% for the covering bands.**

***Index Terms*—Antenna array, broadside scanning, horizontal polarization, omnidirectional antenna, open stopband, periodic leaky-wave antenna, rotating electric field.**

I. Introduction

IN indoor and urban areas, although many current wireless systems are vertically polarized, the polarization of the propagating electromagnetic wave may change significantly after going through complicated multiple reflections and/or scatterings [1], [2]. Polarization diversity has been receiving much attention as it is considered an optimization process for multiple-input-multiple-output (MIMO) technique, especially in rich multi-path communication environments [3]. Omnidirectional antenna has led to a wide range of applications in wireless communications such as wireless local area network (WLAN) systems or distributed MIMO systems [4]. Hence, a horizontally polarized antenna with an omnidirectional pattern is required as a component of a polarization diversity antenna to obtain high polarization purity for maximizing system's capacity [5]. However, the design of a horizontally polarized omnidirectional antenna array with a high gain and input impedance acceptable for matching is still a challenge.

According to the theory of magnetic dipole, small loop antenna with a uniform current distribution may be a proper choice to achieve a horizontally polarized radiation pattern. However, a small loop antenna is difficult for impedance matching by the reason of very small radiation resistance and high reactance [6]. Several kinds of Alford-loop-structure antennas [7]–[11] have been studied and introduced to generate magnetic dipole radiation patterns. A fundamental problem of Alford loop antennas is its narrow bandwidth and poor impedance matching. Turnstile or superturnstile antennas [12]–[15] have also been widely used as transmitting antenna in TV broadcasting, when there is a need for a omnidirectional radiation pattern with horizontal polarization. They are based on the rotating field method, which is usually realized with conventional folded or crossed dipoles in horizontal plane or with so-called batwing radiating elements that are fed in phase quadrature. These designs, however, have limited gain as the power decays along the length of the antenna and do not reach the end [16]. Moreover, the arrangement of the series feed system with rotating phase is complicated.

Leaky-wave antennas (LWAs) thanks to their highly directive beams and simple feeding structure, have also received much attention in the last decades [17]. Periodic LWAs form an important class of traveling-wave antennas, in which the geometry of the guiding structure is periodically modulated along its length. These antennas are low-profile and relatively simple to fabricate, and can produce narrow beams that point either in the backward direction or the forward direction in space. The radiation is due to a leaky mode that leaks power into free space, with radiation occurring from the fast $m=-1$ space harmonic [18]. These designs also have a special problem near broadside where a narrow region around broadside, usually known as the *open-stopband* region occurs within this narrow angular region the amount of radiation drops substantially, and a large VSWR is encountered.

In this paper we propose, design and experimentally verify a periodic leaky-wave antenna for broadside radiation with horizontally polarized omnidirectional pattern. This paper presents a full-space scanning leaky-wave antenna using periodic phase-reversal unit cells interconnected by balanced parallel stripline sections. As the phase-reversal unit cells are in continuity of

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K. Wei, Z. Zhang, and Z. Feng are with State Key Lab of Microwave and Communications, Tsinghua National Laboratory for Information Science and Technology, Tsinghua University, Beijing 100084, China (e-mail: zjzh@tsinghua.edu.cn)

M. F. Iskander is with HCAC, University of Hawaii at Manoa, Honolulu, HI 96822 USA (e-mail: iskander@spectra.eng.hawaii.edu).

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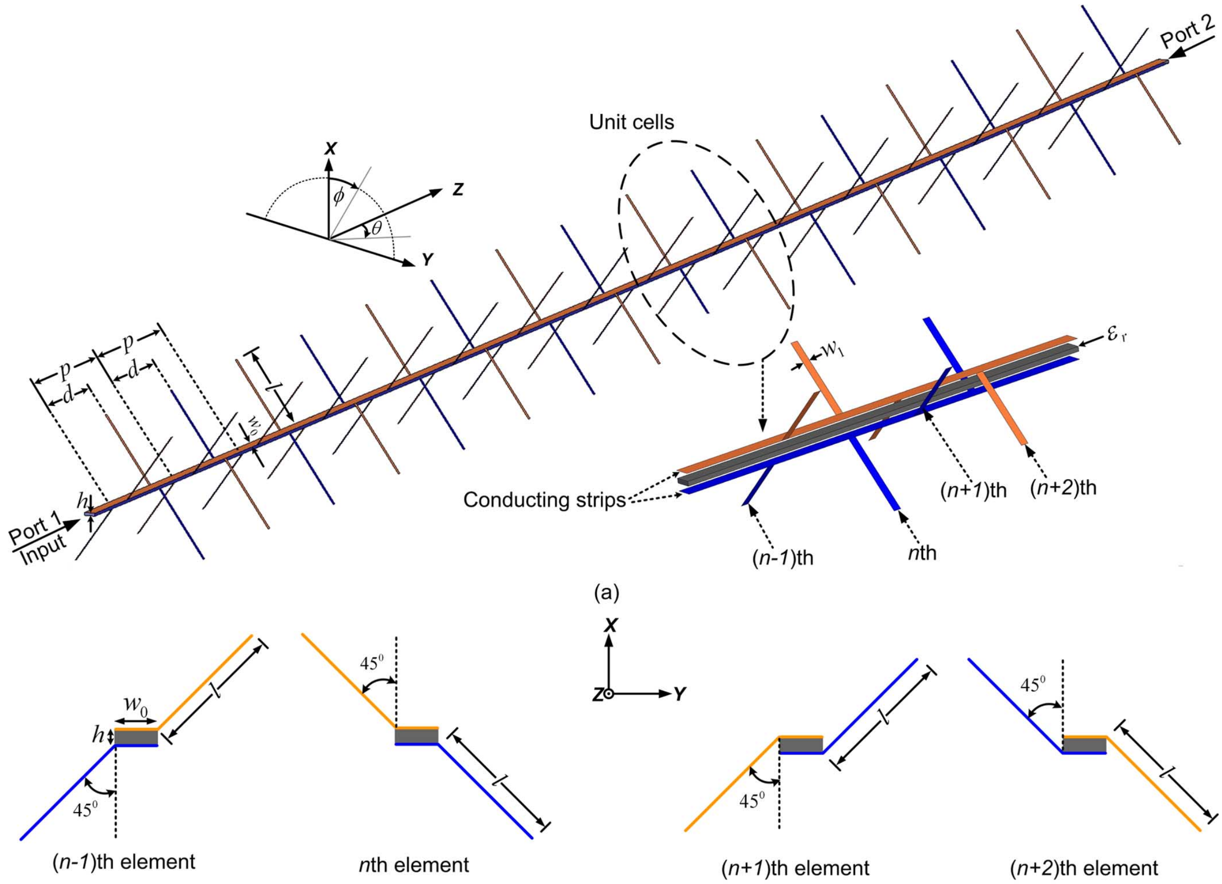

Fig. 1. Configuration of the proposed periodic leaky-wave antenna for broadside radiation with horizontally polarized omnidirectional pattern. The two colors (yellow and blue) indicate different conducting metal strip, which are placed at each side of a thin substrate. (a) 3-D view with the coordinate system and a representation of the unit cell. (b) Side view of the crossed dipole elements.

the balanced transmission line, they induce minima reflections and also lead to a small leakage constant and a large directivity. Moreover, these minor discontinuities can be easily matched within a unit cell with a pair of crossed dipoles spaced at a distance approximately equal to $\lambda_g/4$ at the broadside frequency, so as to mitigate this open-stopband effect, which is similar to that adopted in earlier publications [19]–[25].

II. Antenna Configuration and Analysis

Fig. 1(a) shows the overall configuration of the proposed periodic leaky-wave antenna for broadside radiation with horizontally polarized omnidirectional pattern. The structure consists of a plurality of balanced parallel stripline sections periodically interconnected by phase-reversal unit cells with a period of p. The parallel striplines in this structure consists of two parallel conduction strips of width w_0 separated by a dielectric material of permittivity ε_r and height h. All the parallel strip transmission line sections have the same characteristic impedance Z_0 and phase constant β_{TL}. The proposed antenna is fed at one end and is terminated at the other end by a matched (Z_0) load.

Each unit cell is a pair of two crossed dipole elements spaced by distance d. As shown in Fig. 1(b), one unit cell is composed of the $(n-1)$th and nth dipole elements, the next unit cell is composed of the $(n+1)$th and $(n+2)$th dipole elements. The adjacent unit cells exchange the relative positions of the two arms of dipole element and thereby reverse their polarity. Although the unit cell composed of the $(n-1)$th and nth dipole elements and the unit cell composed of the $(n+1)$th and $(n+2)$th dipole elements are not exactly equal, the small difference can be neglected in practices, so that the period of the structure can be considered to be equal to p. The phase-reversal unit cells have a threefold function: providing a periodic perturbation to generate space harmonics, 180° phase shift for single-beam operation and rotating electric field for horizontally polarized omnidirectional pattern.

Without the phase-reversal unit cells, the structure is merely a uniform TEM transmission line. It is purely guiding (nonradiative) structure. The periodic phase-reversal unit cell transforms the structure into a periodic structure. Using Bloch-Floquet's theorem [18], the wave in a periodic structure consists of the superposition of an infinite number of space harmonics waves, and the phase constants β_m of the mth space harmonic is given by

$$\beta_m(\omega) = \pm\left[\beta_{TL}(\omega) + \frac{(2m+1)\pi}{p}\right], \quad m = 0, \pm1, \pm2, \ldots. \quad (1)$$

As the phase reversal induced in each unit cell generates an extra frequency-independent π phase shift per cell, a horizontal shifting of the entire dispersion diagram by an amount of π

is appeared in this relation. Furthermore, it is assumed that $\beta_{TL}(\omega) > 0$ so that the positive and negative signs correspond to waves propagating along the positive (forward) and negative (backward) z-directions, respectively. The angle of radiation of the beam associate with mth space harmonic of the leaky-wave antenna is then given by the classical formula $\theta_m = \cos^{-1}(\beta_m/k_0)$. For a practical antenna, only a single radiated beam is wanted, so the forward $m = -1$ space harmonic is chosen. Therefore, with (1), we find that the frequency corresponding to broadside radiation is given by

$$f(\theta_{-1} = 90^{circ}) = \frac{c}{2p\sqrt{\varepsilon_e}}. \tag{2}$$

Here, c is the velocity of light and ε_e is the effective permittivity of the original parallel strip transmission line. Therefore, the period p is equal to $\lambda_g/2$ at the broadside frequency. The (1) and (2) come into existence in the limit of infinitesimally small perturbations on the stripline. When the perturbations by the radiating elements become large and should be taken into account, the dispersion curves on the Brillouin diagram will depart from the unperturbed dispersion curve of the quasi-TEM mode supported by the stripline. However, the proposed periodic leaky-wave antenna for broadside radiation is designed with a small leakage constant and a large directivity, so that the perturbations by the radiating elements are very small and the (1) and (2) are almost correct.

III. Rotating Electric Field Method

With the aim of achieving a horizontally polarized omnidirectional pattern at the azimuth plane, the rotating electric field method is applied in this structure. It is realized by designing a unit cell with a pair of crossed dipole elements such as the $(n-1)$th and nth dipole elements shown in Fig. 1(b). The effective radiating current of a unit cell is shown in Fig. 2(a) and is obtained by the vectorial summation of the currents of the $(n-1)$th and nth elements. Because of the leakage, the periodic leaky-wave antenna has a complex propagation constant γ, with a phase constant β and a leakage constant α. Therefore, the current I_{n-1} and I_n correspondingly to $(n-1)$th and nth dipole element, has the relationship as

$$I_n = e^{-\gamma d} I_{n-1} = e^{-\alpha d} e^{-j\beta d} I_{n-1}. \tag{3}$$

The leakage constant α could be large or small depending on whether the leakage per unit length is large or small. In order to obtain a long effective aperture and a narrow beam, a low value of α is usually chosen and the leakage $e^{-\alpha d}$ should approximately be equal to 1 in a unit cell. When the distance $d = \lambda_g/4$, with (3), we find that the relationship translates into

$$I_n \approx -jI_{n-1}. \tag{4}$$

Hence, crossed dipole pairs in a unit cell are fed in phase quadrature by the traveling excitation along the transmission line. It's

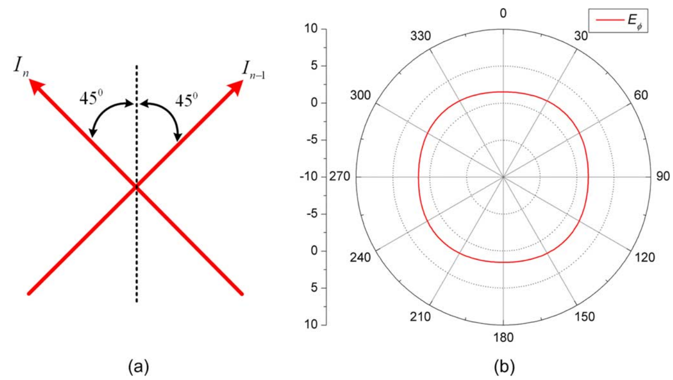

Fig. 2. (a) Effective radiating current of a unit cell. (b) Analytic radiation pattern of the unit cell at the azimuth plane (x-y plane), $l = 0.15\lambda_g, d = \lambda_g/4$.

assumed that the current along the dipole element is cosine distribution. Like the George Brown rotating field antenna [6], the electric field at the azimuth plane is given by

$$\overrightarrow{E}(r,\phi,0) = E_{n-1}(r)\left\{\hat{\phi}\left[\frac{\cos\left(k_0 l\sin(\phi-45^\circ)\right)-\cos k_0 l}{\cos(\phi-45^\circ)}\right.\right. \\ \left.\left. + j\frac{\cos\left(k_0 l\cos\left(\phi-45^\circ\right)\right)-\cos k_0 l}{\sin\left(\phi-45^\circ\right)}\right]\right\} \tag{5}$$

where l is the length of the dipole arm. When the length $l = 0.15\lambda_g$, according to (5), analytic radiation pattern of the unit cell at the azimuth plane is shown in Fig. 2(b). It is clear that good horizontally polarized omnidirectional radiation at the azimuth plane (x-y plane) with small gain variation less than 0.6 dBi is achieved. Therefore, an omnidirectional radiation pattern with horizontal polarization is obtained by designing a unit cell with a pair of crossed dipole elements spaced by a distance d equal to $\lambda_g/4$.

IV. Suppression of the Open Stopband

The phase-reversal array discussed in Section II can be regarded as series-fed antenna array, where the parallel strip transmission line sections build the series feeding structure and the unit cells constitute the radiating elements. In conventional phase-reversal array [23], it is well known that an open stopband at broadside occurs due to coupling of a pair of space harmonics in the radiating region of an open structure. To show the effects of the open stopband, the overall N -unit cell conventional phase-reversal leaky-wave structure is modeled using the equivalent circuit shown in Fig. 3. This model shows the self impedances of the dipole elements and assumes ideal transmission line section interconnections. The transmission coefficient T and reflection coefficient Γ at the nth dipole element $(n = 1, \ldots, N)$ can be written as a function of the self impedance $Z_n = R_n + jX_n$. Assuming that the circuit after the nth element is matched to Z_0, reflection coefficient Γ and transmission coefficient T are

$$\Gamma = \frac{\frac{Z_0 Z_n}{Z_0+Z_n} - Z_0}{\frac{Z_0 Z_n}{Z_0+Z_n} + Z_0} = \frac{1}{1+2\frac{Z_n}{Z_0}} \tag{6a}$$

$$T = \sqrt{1-\Gamma^2} \approx \frac{2}{2+\frac{Z_0}{Z_n}} \tag{6b}$$

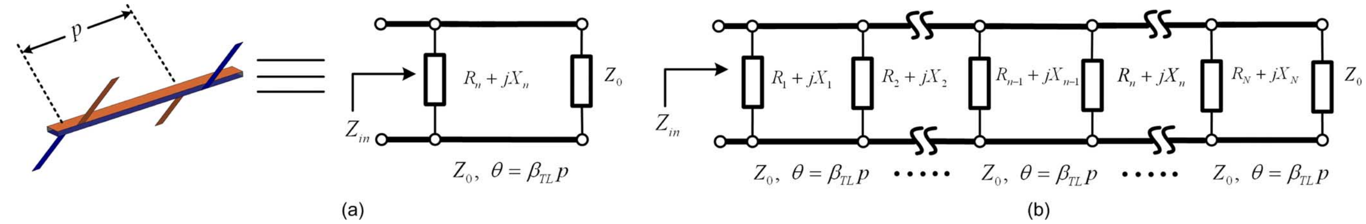

Fig. 3. Transmission line model of conventional phase-reversal array. (a) Exact unit cell structure and equivalent lumped-element circuit model. (b) Overall model.

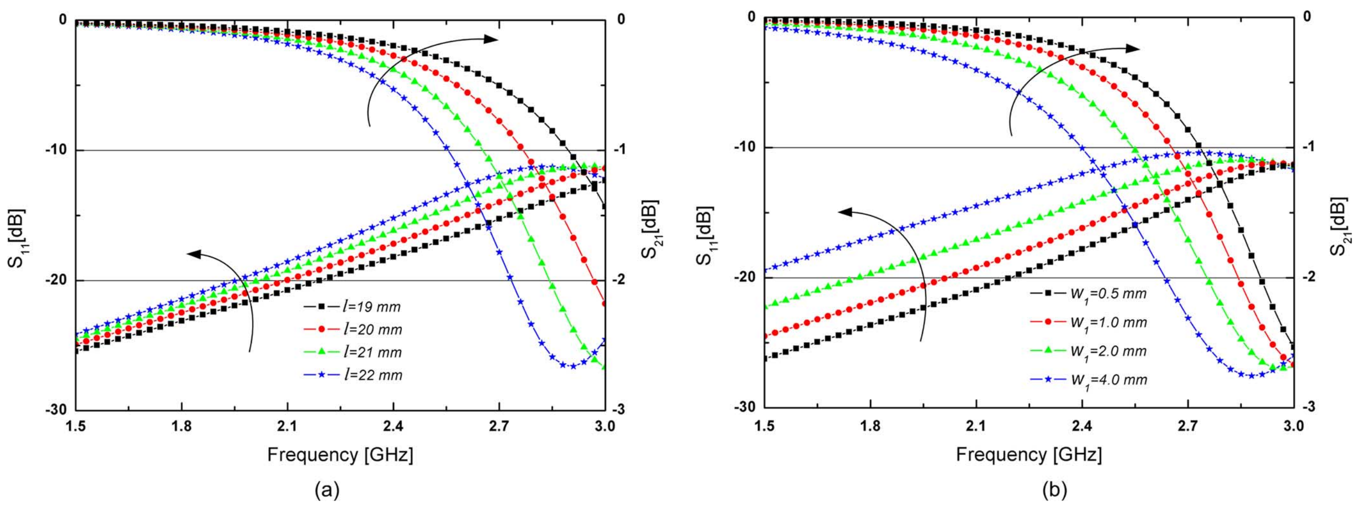

Fig. 4. S-parameters for a one-unit cell conventional phase-reversal antenna simulated from the equivalent transmission lines model of Fig. 3(a). The load and source impedances are of $Z_0 = 50\ \Omega$. (a) Various lengths l of the dipole arm with $w_1 = 1$ mm. (b) Various width w_1 of the dipole arm with $l = 21$ mm.

when $|Z_n| \gg Z_0$, the approximation in the (6b) holds. In practice, R_n is in the order of 15 to 30 Ω while X_n is in the order of -200j to -100j, and the corresponding reflection coefficient range from -15 to -30 dB, as confirmed in Fig. 4. Due to the large distance $(p = \lambda_g/2)$ between the adjacent elements, the mutual coupling can be almost neglected. Therefore, the value of the self impedance Z_n can be extracted through simulations of a single dipole by Ansoft simulation software high frequency structure simulator (HFSS). The length l and the width w_1 vary to change the self impedance $Z_{n,}$ which covers the typically required self impedance of the proposed antenna. As shown in Fig. 4, the reflection for a unit cell is very small and typically below -10 dB.

However, as the number of elements increases, the total reflection of conventional phase-reversal array increases at the broadside frequency. Supposing that the phase-reversal periodic leaky-wave antenna has 32 elements along the transmission, the open stopband effect is analyzed by the equivalent transmission lines model shown in Fig. 3(b). Each unit cell is only a single dipole element with $w_1 = 1$ mm, $l = 21$ mm and $p = \lambda_g/2$ at the center frequency 2.44 GHz. In the case of a uniform array, all the dipole elements approximately have the same impedance value $(Z_1 = Z_2 = \cdots = Z_n)$. The equivalent transmission lines model of the conventional phase-reversal array is simulated by AWR Design Environment 2004. Therefore, the phase constant β and the leakage constant α can be calculated from the following equation:

$$S_{21} = e^{-\gamma L}\sqrt{1 - |S_{11}|^2} = e^{-(\alpha + j\beta)L}\sqrt{1 - |S_{11}|^2} \qquad (7)$$

where L is the total length of the antenna.

The total reflection coefficient at the input port is shown in Fig. 5(a) for the case of a uniform array. The return loss near the broadside frequency decrease to 1 dB and the insertion loss show a dip, which means that the radiated power drops quickly at this frequency. The phenomenon is called the *open stopband* region at broadside. It can also be explained from the dispersion diagram of the periodic structure, which is plotted in Fig. 5(b). A sharp drop of the leakage constant occurs near the broadside frequency $(\beta = 0)$, where the electric length between the adjacent elements equals to π. Note that in a narrow frequency region around broadside, the phase constant β also displays a very small variation where it becomes slightly nonlinear. This stopband may be understood in terms of coupling between space harmonics, or equivalently, in terms of the constructive interferences from the reflections occurring at the periodic loads.

In order to obtain effective radiation and linearly-varying phase constant at broadside, self matching technique is normally used. This self matching technique is based on small-perturbation theory [24], [25] and multiple reflections. In this paper, a pair of two crossed dipole elements is introduced to achieve self matching, where the spacing between the elements of each pair is equal to d. The corresponding transmission line model of this structure is presented in Fig. 6. When $d = 0.25\lambda_g$, so that $2\beta_{TL}d = \pi$ at the broadside frequency, and the wave reflected by the first element of each pair is nearly canceled by the wave reflected by the second element at the broadside frequency. It's confirmed by Fig. 7, which shows s-parameters and dispersion diagram versus frequency for various spacings d. It is observed that the curve $d = 0.25\lambda_g$ exhibits almost no variations near broadside while the other curves has a rapid variation

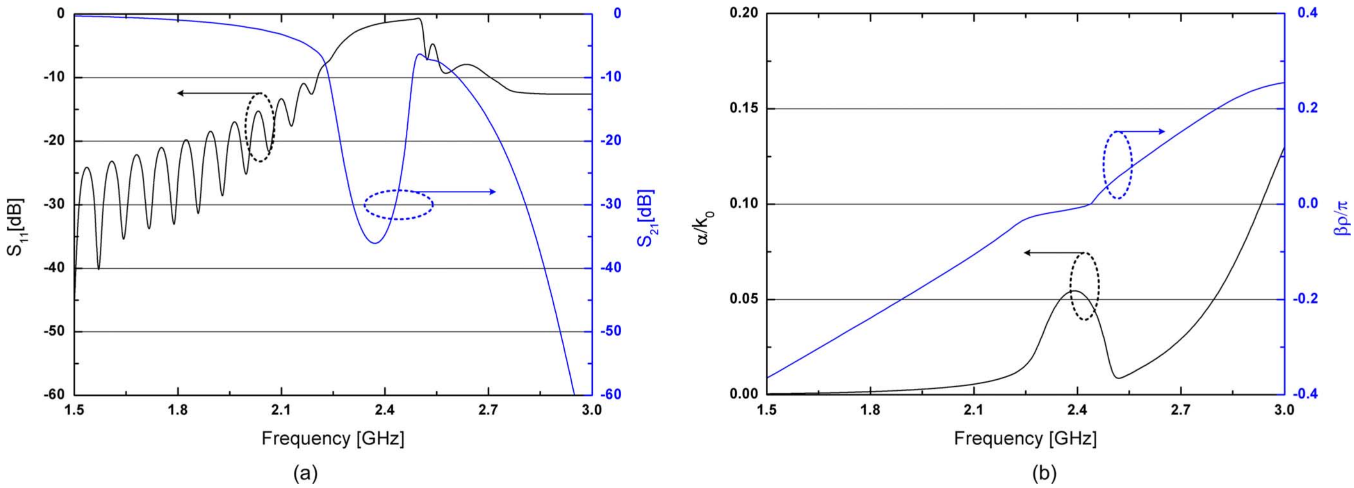

Fig. 5. The open stopband effect of a 32-element conventional phase-reversal array. (a) S-parameters. (b) Dispersion diagrams.

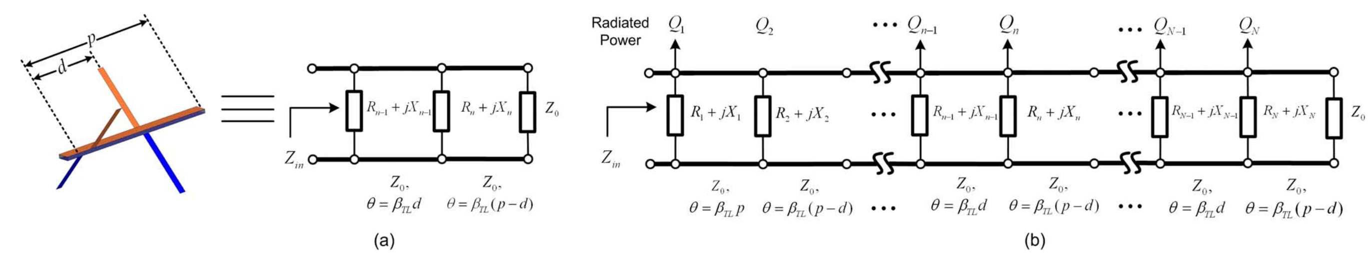

Fig. 6. Transmission line model of the phase-reversal array consisting of a crossed dipole pair per unit cell proposed in Fig. 1, where Q_n is power radiated from the nth radiator. (a) Exact unit cell structure and equivalent lumped-element circuit model. (b) Overall model.

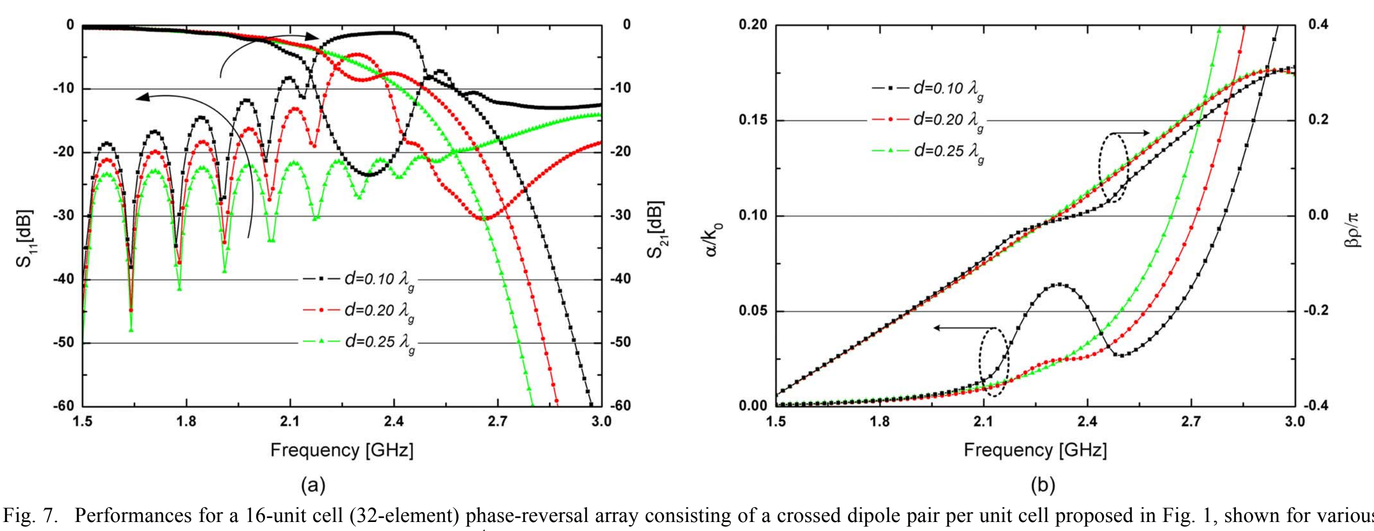

Fig. 7. Performances for a 16-unit cell (32-element) phase-reversal array consisting of a crossed dipole pair per unit cell proposed in Fig. 1, shown for various spacings d, with $w_1 = 1$ mm, $l = 21$ mm and $p = \lambda_g/2$ at the center frequency 2.44 GHz. (a) S-parameters. (b) Dispersion diagrams.

in this region. The phase constant and leakage constant of the proposed unit cell are unaltered compared to the case of Fig. 5 (conventional phase-reversal array) at frequencies away from broadside. The self matching technique by introducing a crossed dipole pair per unit cell nearly eliminates this stopband behavior. Therefore, the proposed periodic leaky-wave antenna array design in Fig. 1 is suitable for scanning through broadside without any large frequency regions of high attenuation.

V. Antenna Design

As discussed in Section III, the rotating electric field method is realized by properly designing a unit cell with a pair of crossed dipole elements spaced by a distance equal to $\lambda_g/4$ at the broadside frequency, which is able to obtain a horizontally polarized omnidirectional pattern at the azimuth plane. Moreover, the open stopband effect in periodic leaky-wave antenna array is completely suppressed by a self matching technique in Section IV, which introduces a pair of crossed dipole elements spaced by a distance approximately equal to $\lambda_g/4$ per unit cell. Therefore, the two requirements of the periodic leaky-wave antenna array in this paper are identical.

To validate the proposed periodic leaky-wave antenna for broadside radiation with horizontally polarized omnidirectional pattern, an example is presented with the following specifications: 2.44 GHz for the broadside radiation, a radiation efficiency of 90% and 3-dB beamwidth of 10° (approximately corresponding to 10.5 dBi gain). The antenna length L is usually selected for a given value of α, so that 90 percent of the power is radiated, with the remaining 10 percent absorbed by a match load. For a leaky-wave structure that is maintained uniform along its length, the approximate design formula of the antenna length L and 3-dB beamwidth $\Delta\theta_{3\ \mathrm{dB}}$ will be [17]

$$\frac{L}{\lambda_0} \approx \frac{0.18}{\alpha/k_0}, \quad \Delta\theta_{3\ \mathrm{dB}} \approx \frac{0.91}{(L/\lambda_0)\sin\theta_m} \tag{9}$$

where θ_m is the scan angle and k_0 is the free-space wave-number. Hence, from the formula (9), the antenna length $L \approx 5.2\lambda_0$ and the leakage constant $\alpha/k_0 \approx 0.035$ to satisfy the specifications.

The leakage constant α of the antenna is proportional to the self impedance distributed over the extent of the structure. As shown in Fig. 8, the leakage constant α increases as the width w_1 and lengths l of the dipole arm increases. This is due to the increasing of the radiation resistance of each dipole element. By tuning the width w_1 and lengths l of the dipole arm, an appropriate leakage constant α will be obtained. The leakage constant $\alpha/k_0 \approx 0.035$ is achieved with $w_1 = 1$ mm and $l = 21$ mm. The overall antenna is designed on a low-cost teflon substrate with a height $h = 0.8$ mm and a dielectric constant $\varepsilon_r = 2.65$, leading to a period $p = 39$ mm and $d = 19.5$ mm. According to the antenna length $L \approx 5.2\lambda_0, N = 16$ unit cells are required to provide $\Delta\theta_{3\ \mathrm{dB}} = 10°$. The width w_0 of parallel strip transmission line is designed to be 2.8 mm with characteristic impedance $Z_0 = 50\ \Omega$. To verify the horizontally polarized omnidirectional characteristic of the proposed antenna, the 16-unit cell (32-element) periodic leaky-wave antenna with crossed dipole pair was full-wave simulated using Ansoft simulation software high frequency structure simulator (HFSS). The corresponding results are compared with theoretical analytic results of equivalent lumped-element circuit model in Fig. 3. As the spacing $d = p/2 = \lambda_g/4$, the theoretical analytical radiation pattern $F(\theta, \varphi)$ is given by

$$F(\theta,\phi) = \sum_{n=1}^{N} Q_n e^{-\frac{jn\pi}{2}} e^{jk_0 nd\cos\theta} f\left(\theta, \phi + \frac{n\pi}{2}\right) \tag{10}$$

where $f(\theta, \varphi)$ is the radiation pattern of the dipole element, which can be estimated from a parameter extracting technique. Q_n is power radiated from the nth radiator as shown in Fig. 6(b) and can be calculated from the leakage constant α. Full-wave simulated and analytical radiation pattern of the 16-unit cell periodic leaky wave antenna are shown in Fig. 9. It is clear that the analytical results of equivalent lumped-element circuit model follow the full-wave simulation very well, which confirms the proposed equivalent circuit model and validates its accuracy. The beam direction in H-plane points towards to broadside radiation with about 10.4 dBi gain at 2.44 GHz. Good horizontally polarized omnidirectional radiation in the H-plane (x-z plane) with small gain variation less than 1.2 dBi is obtained. The sidelobe level (SLL) of radiation pattern is -12.6 dB at broadside

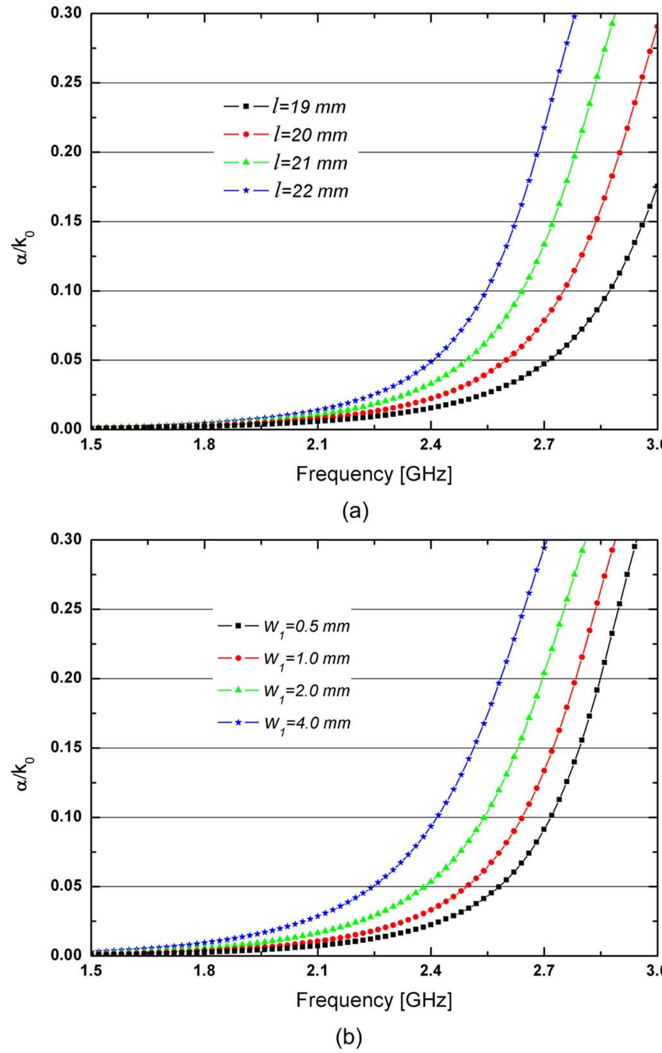

Fig. 8. The leakage constant α of the proposed antenna from the equivalent transmission lines model of Fig. 6, with $d = \lambda_g/4$ and $p = \lambda_g/2$ at the center frequency 2.44 GHz. (a) Various lengths l of the dipole arm ($w_1 = 1$ mm). (b) Various width w_1 of the dipole arm ($l = 21$ mm).

frequency. The antenna in this paper is a uniform array, which is not the optimized structure for SLL. The SLL of the uniform aperture antenna can be decreased by an efficient array synthesis procedure. The technique to taper the LWA illumination [26], [27] is a well known method to reduce the SLL. In our case, we should first taper the parameters w_1 and l along the LWA length to obtain the requested variation of the leakage rate, and then slightly tune the period p to keep the broadside radiation condition ($\theta_{-1} = 90°$) for all longitudinal sections of the antenna. It must be noticed that an iterative process of the two procedures (tune the parameters w_1, l and tune the period p) is needed.

VI. Experimental Demonstration

The 16-unit cell (32-element) periodic leaky wave antenna presented in Section V is fabricated and experimentally tested. A simple prototype is developed to provide a verification of the new design method of high gain horizontally polarized omnidirectional antenna. Photograph of the simple prototype is shown in Fig. 10. The substrate of the parallel striplines is low-cost telfon with dielectric constant of $\varepsilon_r = 2.65$ and height of

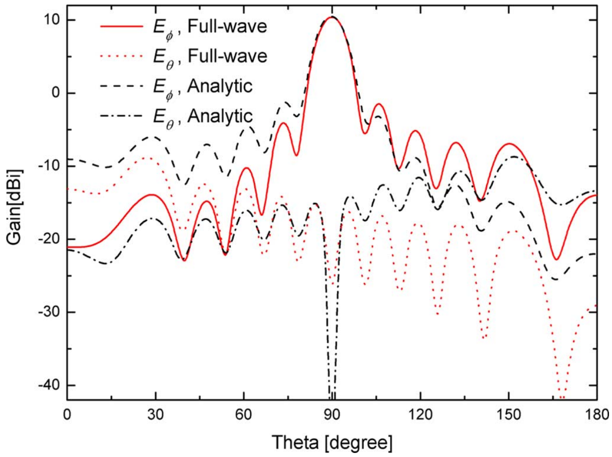

Fig. 9. Full-wave simulated and analytical H-plane (x-z plane) radiation pattern of the 16-unit cell (32-element) periodic leaky-wave antenna at 2.44 GHz, with $\varepsilon_r = 2.65, h = 0.8$ mm, $p = 39$ mm, $d = 19.5$ mm, $w_0 = 2.8$ mm, $w_1 = 1$ mm and $l = 21$ mm.

Fig. 10. Photograph of the developed periodic leaky-wave antenna.

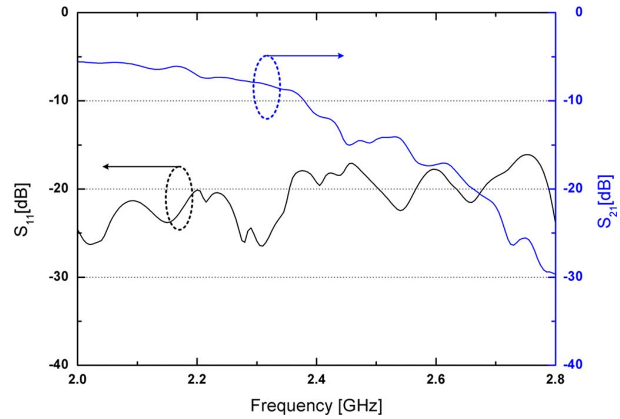

Fig. 11. Measured s-parameters of the prototype.

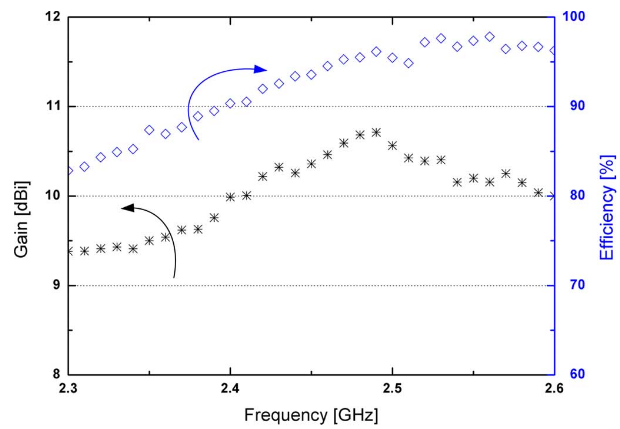

Fig. 12. Measured antenna gain and efficiency of the prototype.

0.8 mm. The overall antenna including the input and matching load is about 640-mm long.

Fig. 11 shows the measured s-parameters of the fabricated antenna. Reflection level of the fabricated antenna is -18 dB, while transmission level is below -11 dB at the design frequency 2.4 GHz for WLAN service. To verify that periodic leaky-wave antenna has a horizontally polarized omnidirectional radiation pattern, the radiation characteristics of the prototype was also studied. An ETS 3-D chamber was used to measure radiation pattern of the fabricated antenna. Fig. 12 shows the measured antenna gain and efficiency as a function of the operating frequency. The measured gain of the fabricated antenna with the case varied from 9.9 to 10.6 dBi at the 2.4 GHz WLAN band. The ETS chamber can also provide an estimated value of the radiation efficiency of the measured antenna. The efficiency is defined as the ratio of radiated power versus total available power from power source. Thus the efficiency value includes all impacts from mismatch loss, dielectric loss, conductor loss and matching component loss. The efficiency of the fabricated antenna was found to exceed 90% in the covering bands.

The measured H-plane (x-z plane) radiation patterns of the fabricated periodic leaky-wave antenna are presented in Fig. 13(a). To better evaluate the results, three scanned radiation patterns at 2.30, 2.44 and 2.60 GHz are chosen to shown, which corresponds to the main beam angle at $\theta = 84°, \theta = 90°$ (broadside), and $\theta = 96°$ respectively. Broadside radiation occurs at 2.44 GHz. At this point, the measured gain is 10.35 dBi and the measured 3-dB beamwidth is of 10°. The SLL of radiation pattern in H-plane is of -13.2 dB at broadside frequency. The reason of this small discrepancy between measured and simulated SLL may be due to re-radiation from the signal reflected for the end matching load at mirrored angles. The measured and full-wave simulated azimuth plane (x-y plane) patterns are compared in Fig. 13(b). Good agreement between the measurement and the simulation has been obtained. It is clear that good omnidirectional radiation with horizontal polarization in the azimuth plane is obtained. Due to the existence of the parallel striplines, the radiation pattern in azimuth plane slightly tilts towards to x-axis. The gain variation in the azimuth plane is below 1.8 dB over the WLAN band, which represents a stable omnidirectional coverage.

VII. Conclusion

A novel full-space novel periodic leaky-wave antenna for broadside radiation with horizontally polarized omnidirectional pattern, radiating in the $m = -1$ space harmonic, has been proposed, designed, fabricated, and tested. In contrast to previously reported leaky-wave antennas, this antenna radiates from its phase-reversing unit cells interconnected by

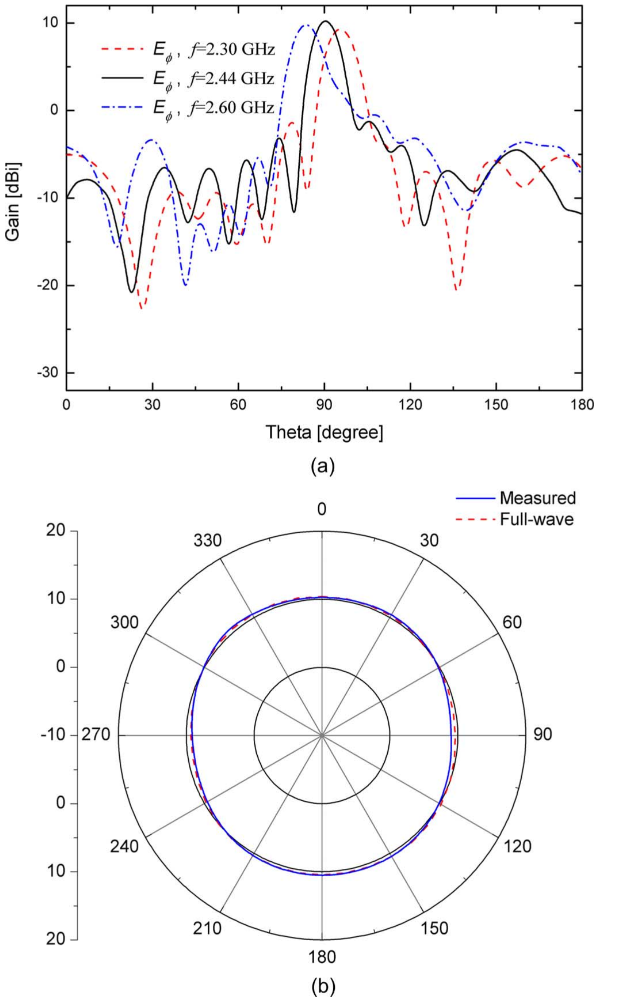

Fig. 13. Measured radiation pattern of the fabricated periodic leaky-wave antenna, with identical dimensions as discussed in Fig. 8. (a) H-plane (x-z plane). (b) E-plane (x-y plane).

balanced parallel stripline sections, which allows for small leakage constant and subsequently large directivities. The rotating electric field method and leaky-wave antenna are combined in this structure to obtain a horizontally polarized omnidirectional antenna array with highly directive beams. It can be seen that, by properly designing a unit cell with a pair of crossed dipole elements spaced by a distance approximately equal to $\lambda_g/4$ at the broadside frequency, not only omnidirectional radiation pattern with horizontal polarization is obtained by crossed dipoles fed in phase quadrature, but also the open-stopband effect exhibited at broadside scan is significantly reduced. A 16-unit cell (32-element) periodic leaky wave antenna which is maintained uniform along its length, was designed and simulated to validate the proposed approach. A simple prototype that follows the design was also constructed and experimentally tested. The measured beam-width and gain at the broadside frequency of 2.44 GHz are of 10° and 10.35 dBi, respectively. Good horizontally polarized omnidirectional radiation in the E-plane (x-y plane) with small gain variation less than 1.8 dBi is obtained.

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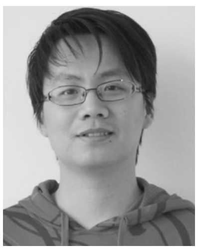


Kunpeng Wei was born in Nanchang, China, in 1987. He received the B.S. degrees in electronic and information engineering from the HuaZhong University of Science and Technology, Wuhan, China, in 2008. He is currently working toward Ph.D. degree in electrical engineering from Tsinghua University, Beijing, China.

His current research interests include antenna design and theory, particularly in omni-directional antennas and base station antennas.

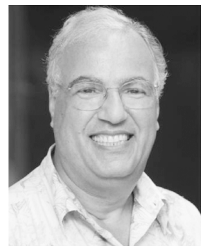
Zhijun Zhang (M'00–SM'04) received the B.S. and M.S. degrees from the University of Electronic Science and Technology of China, in 1992 and 1995, respectively, and the Ph.D. degree from Tsinghua University, Beijing, China, in 1999.

In 1999, he was a Postdoctoral Fellow with the Department of Electrical Engineering, University of Utah, where he was appointed a Research Assistant Professor in 2001. In May 2002, he was an Assistant Researcher with the University of Hawaii at Manoa, Honolulu. In November 2002, he joined Amphenol T&M Antennas, Vernon Hills, IL, as a Senior Staff Antenna Development Engineer and was then promoted to the position of Antenna Engineer Manager. In 2004, he joined Nokia Inc., San Diego, CA, as a Senior Antenna Design Engineer. In 2006, he joined Apple Inc., Cupertino, CA, as a Senior Antenna Design Engineer and was then promoted to the position of Principal Antenna Engineer. Since August 2007, he has been with Tsinghua University, where he is a Professor in the Department of Electronic Engineering. He is the author of *Antenna Design for Mobile Devices* (Wiley, 2011).

Dr. Zhang is serving as an Associate Editor of the IEEE TRANSACTIONS ON ANTENNAS AND PROPAGATION and the *IEEE Antennas and Wireless Propagation Letters*.

Zhenghe Feng (M'00–SM'08–F'12) received the B.S. degree in radio and electronics from Tsinghua University, Beijing, China, in 1970.

Since 1970, he has been with Tsinghua University, as an Assistant, Lecture, Associate Professor, and Full Professor. His main research areas include numerical techniques and computational electromagnetics, RF and microwave circuits and antenna, wireless communications, smart antenna, and spatial temporal signal processing.



Magdy F. Iskander (F'93) is the Director of the Hawaii Center for Advanced Communications (HCAC), College of Engineering, University of Hawaii at Manoa, Honolulu (http://hcac.hawaii.edu). He is Co-Director of the NSF Industry University Cooperative Research Center with four other universities. From 1997–99 he was a Program Director at the National Science Foundation, where he formulated a "Wireless Information Technology" Initiative in the Engineering Directorate. He authored the textbook *Electromagnetic Fields and Waves* (Prentice Hall, 1992 and Waveland Press, 2001; 2nd edition 2012), edited the *CAEME Software Books*, Vol. I, II 1991–94; and edited four books on *Microwave Processing of Materials* (Materials Research Society, 1990–96). He has published over 230 papers in technical journals, has eight patents, and made numerous presentations in national and International conferences. He is the founding editor of the *Computer Applications in Engineering Education* (CAE) journal, (Wiley, 1992–present). His research focus is on antenna design and propagation modeling for wireless communications and radar systems. His areas of research include phased antenna array design for communications and radar applications, wireless propagation modeling in complex communication environments, computational electromagnetics techniques and applications.

Dr. Iskander received the 2010 University of Hawaii Board of Regents' Medal for Teaching Excellence, the 2010 Northrop Grumman Excellence in Teaching Award, the 2011 Hi Chang Chai Outstanding Teaching Award, and the University of Utah Distinguished Teaching Award in 2000. He also received the 1985 Curtis W. McGraw ASEE National Research Award, 1991 ASEE George Westinghouse National Education Award, 1992 Richard R. Stoddard Award from the IEEE EMC Society. He was a member of the 1999 WTEC panel on "Wireless Information Technology-Europe and Japan", and Chaired two International Technology Institute panels on "Asian Telecommunication Technology" sponsored by NSF/DoD in 2001 and 2003. He edited two special issues of the IEEE TRANSACTION ON ANTENNAS AND PROPAGATION in 2002 and 2006, co-edited a special issue of the *IEICE Journal* in Japan in 2004. He is the 2002 President of IEEE Antennas and Propagation Society, Distinguished Lecturer for IEEE AP-S (1994–97) and Fellow of IEEE, 1993.