



Article A Dual-Beam Leaky-Wave Antenna Based on Squarely Modulated Reactance Surface

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Abstract: In this paper, a novel dual-beam leaky-wave antenna (LWA) based on squarely modulated reactance surface (SquMRS) is proposed. The equivalent transmission lines model is utilized to characterize the field distributions of surface wave guided by the SquMRS. The calculated dispersion characteristics of SquMRS are verified by the simulated results, and it is demonstrated that SquMRS exhibits a more flexible control of phase constant and attenuation constant compared with traditional sinusoidally modulated reactance surface (SinMRS), which means SquMRS has a great potential for near-field focusing and far-field beam shaping. On this basis, a versatile method, based on a superposition of individual modulation patterns, was used to generated two beams with almost identical gain at 8.5 GHz. The measured results show that the gains are 10 dBi and 8.2 dBi at $\theta_1 = -30^\circ$ and $\theta_2 = 18^\circ$, respectively, and the radiation efficiency is 83%, which shows good agreement with the simulated results.

Keywords: leaky-wave antenna (LWA); squarely modulated reactance surface (SquMRS); surface wave

1. Introduction

Due to the advantages of high gain, low profiles and simple feeding network, leaky-wave antennas (LWAs) have attracted a lot of attention since being proposed in the 1940s [1]. According to the operating principles, LWAs can be generally divided into two categories, uniform LWAs and periodic LWAs [2–4]. Compared with uniform LWAs, periodic LWAs exhibit a large scanning range and flexible design schemes, which make them widely used in millimeter-wave bands [5,6]. One approach to achieving beam-scanning capability is utilizing the periodical shape changing of transmission line (such as triangle-truncated double-side parallel-strip lines [7] and bends of sharpness [8]). Another demonstrated method is sinusoidally modulating the surface impedance of the LWA [9–12]. Some previous research about sinusoidally modulated reactance surface (SinMRS) has shown the great advantages in the synthesis of radiation patterns. For instance, a simultaneous tapering of the leaky-wave leakage rate and pointing angle along the leaky radiator was used to synthesize radiation beams [13]. Cosine-tapered designs were performed to reduce sidelobe levels and tune the pointing direction over a wide range [14]. Recently, an array of non-uniform SinMRSs with application to near-field focused leaky-wave radiation in the backward Fresnel zone were proposed [15].

For LWAs based on SinMRS, at least five kinds of unit cells have to be selected to mimic the sinusoidal distribution of surface impedance [16], which makes the construction of LWAs extremely complex. More recently, SinMRS, constructed with five kinds of tunable unit cells, was utilized to validate beam steering at fixed frequency [17,18]. In theory, the bias voltage of each kind of unit cell should be independently controlled, which leads to the complex bias network and additional energy loss. To avoid this situation, different kinds of unit cell are designed with various dimensions so that

their surface impedances can satisfy the requirement under a uniform bias voltage control [17,18]. However, the distribution of surface impedance during one period is restricted in several fixed patterns, and the phase constant and attenuation constant of the SinMRS cannot be controlled independently. To overcome this, squarely modulated reactance surface (SquMRS), which is composed of only two kinds of unit cells, is proposed in [19]. Since each kind of unit cell can be treated as a "macro cell" (a macro cell is loaded with only one tunable component) [20], SquMRS can load much less tunable components, which makes it convenient to realize the independent control of phase constant and attenuation constant with the simpler biasing network.

In this paper, a novel SquMRS structure is proposed to generate the leaky-wave radiation. Firstly, the propagation properties of surface wave travelling along the SquMRS are discussed by using a numerical method based on the equivalent transmission lines model proposed in [19]. In order to verify the feasibility of SquMRS, two LWAs have been designed based on SquMRS and SinMRS, respectively. Finally, a dual-beam LWA is presented that uses an impedance superimposing approach. The measured results show a good agreement with the simulated ones.

2. Dispersion Characteristics Analysis

According to the Floquet theory, leaky-wave radiation can be generated by the periodically modulated structure. Therefore, for the construction of spatial distribution of surface impedance, squarely modulated reactance surface (SquMRS), instead of SinMRS, can be adopted to achieve the conversion from surface wave (electromagnetic wave that propagates along the interface between different media) to leaky-wave (electromagnetic wave that is coupled or transferred to a propagation medium outside the interface). The distributions of surface impedances along one period for SquMRS and SinMRS under different modulations, which have identical average value X_s , modulation depth M and modulation period p, are shown in Figure 1. Compared with SinMRS, the structure of SquMRS is much simpler since it needs fewer kinds of unit cells.



Figure 1. The distributions of surface impedances along one period under different modulations.

Assume the SquMRS is placed in the *xoy* plane, while the direction of the propagation along the surface is the *x*-axis, as shown in Figure 2. For simplicity, only transverse magnetic (TM) polarization is discussed. The surface impedance distribution in one period exhibits a form:

$$Z_s(x) = \begin{cases} jX_s(1+M) & \text{if } 0 \le x \le p/4 \text{ or } 3p/4 \le x \le p\\ jX_s(1-M) & \text{if } p/4 < x < 3p/4 \end{cases}$$
(1)

where X_s is the average value, M is the modulation depth and p is the modulation period.



Figure 2. The equivalent transmission lines model.

Due to the periodic function of Equation (1), it can be expressed as a Fourier series expansion, and the impedance profile of squared modulation can be rewritten as:

$$Z_s(x) = jX_s + jMX_s \sum_{m=1}^{\infty} \frac{4}{(2m-1)\pi} e^{-j(2m-1)\frac{2\pi}{p}x}$$
(2)

Here, an equivalent transmission lines model is established to analyze the dispersion characteristics, as shown in Figure 2 and originally proposed in [19]. An infinite number of independent transmission lines along the normal direction are coupled together at the impedance surface. $Z_{eff}^{(n)} = V_n/I_n$ is interpreted as the effective impedance of mode n. The mode voltage V_n and mode current I_n are used to represent the field energy above the surface:

$$\vec{E}_t = (1/\sqrt{2\pi})\vec{x}\sum_{n=-\infty}^{\infty} V_n e^{-jk_{xn}x}$$
(3)

$$\vec{H}_t = (1/\sqrt{2\pi})\vec{y}\sum_{n=-\infty}^{\infty} I_n e^{-jk_{xn}x}$$
(4)

According to the Floquet theory, spatial harmonics can be generated by adjusting the modulation period *p*:

$$k_x^{(n)} = k_x^{(0)} + 2\pi n/p = \beta_{eff}^{(n)} + 2\pi n/p + j\alpha_{eff}$$
(5)

in which $k_x^{(n)}$ and $\beta_{eff}^{(n)}$ are the wave number and phase constant of the *n*th spatial harmonic, α_{eff} is the attenuation constant. It has to be noticed that the periodical modulation has few influences on the attenuation constant according to (5). Therefore, all space harmonics share an identical α_{eff} and possess different $\beta_{eff}^{(n)}$ for each mode *n*. The effective impedance of mode *n* $Z_{eff}^{(n)}$ can be rewritten as:

$$Z_{eff}^{(n)} = j\eta_0 \sqrt{\left(k_x^{(n)}/k_0\right)^2 - 1}$$
(6)

where η_0 is the free-space wave impedance and k_0 is the free-space wavenumber. Using circuit theory, the effective impedance of mode n = 0 can be derived as an infinite continued fraction:

$$Z_{eff}^{(0)} = jX_s + (4X_sM/\pi)^2 \sum_{m=1}^{\infty} \frac{1}{(2m-1)} \left[\frac{1}{jX_s + Z_{eff}^{(2m-1)}} + \frac{1}{jX_s + Z_{eff}^{-(2m-1)}} \right]$$
(7)

Substituting the Equations (5) and (6) into Equation (7):

$$j\eta_{0}\sqrt{\left(\frac{k_{x}^{(0)}}{k_{0}}\right)^{2}-1} = jX_{s} + \left(4X_{s}M/\pi\right)^{2}\left\{\left[\frac{1}{jX_{s}+j\eta_{0}}\sqrt{\left(\frac{k_{x}^{(1)}}{k_{0}}\right)^{2}-1} + \frac{1}{jX_{s}+j\eta_{0}}\sqrt{\left(\frac{k_{x}^{(1)}}{k_{0}}\right)^{2}-1}\right] + \frac{1}{jX_{s}+j\eta_{0}}\sqrt{\left(\frac{k_{x}^{(1)}}{k_{0}}\right)^{2}-1} + \frac{1}{jX_{s}+j\eta_{0}}\sqrt{\left(\frac{k_{x}^{(1)}}{k_{0}}\right)^{2}-1}\right] + \frac{1}{5}\left[\frac{1}{jX_{s}+j\eta_{0}}\sqrt{\left(\frac{k_{x}^{(3)}}{k_{0}}\right)^{2}-1} + \frac{1}{jX_{s}+j\eta_{0}}\sqrt{\left(\frac{k_{x}^{(3)}}{k_{0}}\right)^{2}-1}\right] + \frac{1}{5}\left[\frac{1}{jX_{s}+j\eta_{0}}\sqrt{\left(\frac{k_{x}^{(5)}}{k_{0}}\right)^{2}-1} + \frac{1}{jX_{s}+j\eta_{0}}\sqrt{\left(\frac{k_{x}^{(0)}-2\pi/p}{k_{0}}\right)^{2}-1}\right] + \frac{1}{5}\left[\frac{1}{jX_{s}+j\eta_{0}}\sqrt{\left(\frac{k_{x}^{(0)}+6\pi/p}{k_{0}}\right)^{2}-1} + \frac{1}{jX_{s}+j\eta_{0}}\sqrt{\left(\frac{k_{x}^{(0)}-6\pi/p}{k_{0}}\right)^{2}-1}\right] + \frac{1}{5}\left[\frac{1}{jX_{s}+j\eta_{0}}\sqrt{\left(\frac{k_{x}^{(0)}+10\pi/p}{k_{0}}\right)^{2}-1} + \frac{1}{jX_{s}+j\eta_{0}}\sqrt{\left(\frac{k_{x}^{(0)}-6\pi/p}{k_{0}}\right)^{2}-1}\right] + \frac{1}{5}\left[\frac{1}{jX_{s}+j\eta_{0}}\sqrt{\left(\frac{k_{x}^{(0)}+10\pi/p}{k_{0}}\right)^{2}-1} + \frac{1}{jX_{s}+j\eta_{0}}\sqrt{\left(\frac{k_{x}^{(0)}-10\pi/p}{k_{0}}\right)^{2}-1}\right] + \frac{1}{5}\left[\frac{1}{jX_{s}+j\eta_{0}}\sqrt{\left(\frac{k_{x}^{(0)}+10\pi/p}{k_{0}}\right)^{2}-1} + \frac{1}{jX_{s}+j\eta_{0}}\sqrt{\left(\frac{k_{x}^{(0)}-10\pi/p}{k_{0}}\right)^{2}-1}\right] + \frac{1}{5}\left[\frac{1}{jX_{s}+j\eta_{0}}\sqrt{\left(\frac{k_{x}^{(0)}+10\pi/p}{k_{0}}\right)^{2}-1} + \frac{1}{jX_{s}+j\eta_{0}}\sqrt{\left(\frac{k_{x}^{(0)}-10\pi/p}{k_{0}}\right)^{2}-1}\right] + \frac{1}{5}\left[\frac{1}{jX_{s}+j\eta_{0}}\sqrt{\left(\frac{k_{x}^{(0)}-10\pi/p}{k_{0}}\right)^{2}-1}\right] + \frac{1}{5}\left[\frac{1}{jX_$$

Due to the presence of an additional factor of $4X_S M/\pi$ in each successive term, the continued fractions converge rapidly when the modulation depth *M* has a small value. It is sufficient to yield the solution for $k_x^{(0)}$ to a high degree of accuracy by using (8) with only 3 modes considered, and then the effective phase constant $\beta_{eff}^{(0)}$ and attenuation constant α_{eff} of SquMRS can be calculated as:

$$\begin{cases} \beta_{eff}^{(0)} = \operatorname{Re}\left[k_x^{(0)}\right] \\ \alpha_{eff} = \operatorname{Im}\left[k_x^{(0)}\right] \end{cases}$$
(9)

By using the equations mentioned above, the comparison between SinMRS and SquMRS is presented with parameters $X_s = 430 \ \Omega$ and $p = 30 \ mm$. Figure 3a shows the relations between the dispersion characteristics and modulation depth *M* at operating frequency of 8.5 GHz. It can be seen that both $\beta_{eff}^{(0)}$ and α_{eff} are increased as *M* raises from 0 to 0.4, and the changes of SquMRS are more drastic than SinMRS. For both modulation methods, the variation of $\beta_{eff}^{(0)}$ is slight when the value of *M* is less than 0.2. Meanwhile, SquMRS allows for more design flexibility to choose any required value of α_{eff} , since it has a larger variation range of the effective attenuation constant. Figure 3b presents $\beta_{eff}^{(0)}$ and α_{eff} as a function of the average impedance X_s with M = 0.3. As can be seen, $\beta_{eff}^{(0)}$ is mainly affected by X_s when *M* is fixed. In conclusion, the values of X_s , *M* and *p* should be modified to assure the independent control of attenuation constant and phase constant.



Figure 3. The calculated effective phase constant $\beta_{eff}^{(0)}$ and attenuation constant α_{eff} of SinMRS and SquMRS (**a**) as the function of modulation depth *M* with $X_s = 430 \Omega$ and p = 30 mm, (**b**) as the function of average value X_s with M = 0.3 and p = 30 mm.

The spatial harmonics with $n \le 0$ can be fast wave ($\beta_{eff}^{(n)} < k_0$) by choosing suitable modulation period p as $\beta_{eff}^{(0)}$ is larger than k_0 . In this paper, the n = -1 spatial harmonic is designed to be radiated. The radiation angle θ can be derived as:

$$\theta = \arcsin(\beta_{eff}^{(0)} / k_0 - 2\pi / k_0 p) \approx \arcsin[\sqrt{(X_s / \eta_0)^2 + 1} - 2\pi / k_0 p]$$
(10)

Therefore, SquMRS can generate the radiating beam with the same angle as SinMRS with a relatively simple structure.

In order to construct the LWAs based on SquMRS for verification, the unit cell shown in Figure 4a is proposed and analyzed in detail. The "H"-shaped unit cell was printed on a dielectric substrate with a relative permittivity of 2.65 and a thickness of 1 mm. The advantage of this unit cell is that its surface impedance can be easily modulated by changing the width of the central bar *b*. The length of this unit cell *a* was chosen to be 3 mm (0.085 λ_0 at 8.5 GHz) to guarantee the sub-wavelength structure. Other dimensions of unit cell were d = 1.1 mm, l = 10 mm. The dispersion curves of the unit cell with different *b* were obtained by using the commercial software CST Microwave Studio (Computer Simulation Technology TM, Darmstadt, Germany) [21], as shown in Figure 4b. The phase difference across the unit cell βa was simulated by eigen mode solver under periodic boundary conditions [22]. When *b* decreases from 8 mm to 2 mm, the dispersion curves gradually bend away from the air line, giving rise to a much slower travelling wave. The results show that the phase constant β of the dominant mode (n = 0) becomes higher as *b* decreases. Then, the surface impedance of the unit cell can be calculated as $Z_s = j\eta_0 \sqrt{(\beta a/k_0 a)^2 - 1}$. The relation between the surface impedance and *b* is demonstrated in Figure 4b. The surface impedance ranges from the minimum value of 137.5 $j\Omega$ to the maximum value of 752.3 $j\Omega$ at 8.5 GHz as *b* decreases from 9 mm to 1 mm.



Figure 4. (**a**) Structure of the proposed unit cell. (**b**) The dispersion curves with different *b* and surface impedance of the unit cell with different *b* at 8.5 GHz.

LWAs, based on these two modulation methods, were designed with a similar structure. The schematic of the LWA based on SquMRS is shown in Figure 5, which consists of coplanar waveguides (CPWs), matching transitions and modulation pattern. The proposed LWAs were printed on dielectric substrates with a relative permittivity of 2.65 and a thickness of 1 mm. The dimensions of the CPWs shown in Figure 5 are $w_0 = 3 \text{ mm}$, w = 20 mm and g = 0.1 mm. In order to feed and receive energy, both ends of CPWs are designed to achieve 50Ω impedance. The matching transition, which consists of a flaring ground and gradient grooves, provides a good matching of impedance and momentum between CPWs and modulation pattern [23]. Therefore, a high-efficiency transmission is ensured by the matching transitions. The function of the modulation pattern is to generate the desired leaky-wave radiation by using the periodically modulated reactance surface, which is composed of 10 periods, and each period contains 10 unit cells. The distribution of surface impedance within one period is modulated as described in Figure 1 with $X_s = 430 \Omega$ and p = 30 mm (10 unit cells per period). The

desired radiation angle is set to be $\theta = 20^{\circ}$ by using Equation (8). The simulated radiation patterns of different modulation methods at 8.5 GHz are shown in Figure 6. Table 1 reports the values of radiation angle θ and beam width $\Delta \theta$ with different *M*. For both modulation methods, the radiation angle changes a little when the modulation depth *M* varies from 0.2 to 0.4 and the beam width increases with *M* rising so that the feasibility of LWAs based on squared modulation is proved. It is worth noting that the broader beam is just a characteristic rather than an advantage of SquMRS.



Figure 5. The schematic of LWA based on SquMRS, in which $w_0 = 3$ mm, w = 20 mm and g = 0.1 mm.



Figure 6. The simulated radiation patterns with different modulation depths at 8.5 GHz.

M	θ (SinMRS)	θ (SquMRS)	$\Delta \theta$ (SinMRS)	$\Delta \theta$ (SquMRS)
0.2	22°	21°	6.1°	6.7°
0.3	21°	19°	6.4°	7.2°
0.4	21°	20°	6.9°	8.6°

Table 1. Radiation angles and beam width for various values of *M*.

3. Dual-Beam Leaky-Wave Antenna Implementation

A dual-beam LWA pointing at $\theta_1 = -30^\circ$ (beam 1) and $\theta_2 = 18^\circ$ (beam 2) was designed to operate at 8.5 GHz. In previous studies about producing multi-beams, the modulation pattern was divided into several regions, and each region provided a desired radiation beam [24]. Usually, a monopole antenna was chosen as the feed source so that it was difficult to reduce the height of the antenna. In addition, a holographic synthesizing method, which translated any requested electromagnetic modulation of β and α along the antenna into the corresponding geometrical modulation, was used to produce multiple beams [25]. Several previous designs, based on superposing multiple objective field patterns on the SinMRS, were proposed to produce multi-beams and achieve near-field focusing [26–28]. In this work, a superposition method based on SquMRS was used to design a dual-beam LWA, due to the capability of flexible control of phase constant and attenuation constant. Unlike using n = -1 and n = -2 spatial harmonics to generate dual-beam ($\beta_{eff}^{(-1)} < k_0$, $\beta_{eff}^{(-2)} < k_0$) [29–31], the produced two beams in this work are both excited by the n = -1 space harmonic, and other space harmonics are all slow waves (only $\beta_{eff}^{(-1)} < k_0$). Assuming these two beams possess the identical X_s and M, the radiation angles are decided by the modulation period p_1 and p_2 , respectively. The individual modulation patterns for each beam are added up at each unit of the proposed antenna [22,24]. According to the holographic antenna theory, once the interference pattern is recorded by the interaction between reference wave Ψ_{ref} and multi-beam radiation wave Ψ_{rad} on the modulated surface, the multiple beams can be reconstructed. The multi-beam radiation wave can be written as the sum of two individual beams: $\Psi_{rad} = \Psi_{rad}^{(1)} + \Psi_{rad}^{(2)}$. The distribution of surface impedance is used to embody the whole surface interferogram:

$$Z_{s}(x) = j[X_{s} + MRe(\Psi_{rad} \cdot \Psi_{rad}^{*})] = j\{X_{s} + MRe[(\Psi_{rad}^{(1)} + \Psi_{rad}^{(2)}) \cdot \Psi_{rad}^{*}]\}$$

= $jX_{s}[1 + a_{1}Mf_{1}(x) + a_{2}Mf_{2}(x)] = a_{1}jX_{s}[1 + Mf_{1}(x)] + a_{2}jX_{s}[1 + Mf_{2}(x)]$ (11)

where $f_1(x)$ and $f_2(x)$ are periodic functions of the desired two beams with the surface impedance coefficient a_1 and a_2 ($a_1 + a_2 = 1$), respectively. Therefore, the precondition of this superposing method is that two beams possess the identical X_s and M. Figure 7 shows the simulated radiation patterns of the LWA with different surface impedance coefficients. In theory, two beams should obtain an identical gain when $a_1 = a_2 = 0.5$, but the simulated result shows that a gain difference (about 1 dB) existed. As a_1 raises to 0.6, the gain of beam 1 is 3 dB higher than the gain of beam 2. In these two cases, the changing of surface impedance coefficients has almost no effect on the radiation angles. Finally, the gain difference between two beams is about 6.5 dB when $a_1 = 0.7$, $a_2 = 0.3$. This means that this antenna can be considered as a single-beam antenna pointing at -34° , due to beam 2 actually being a side lobe ($\theta_2 = 16^\circ$). It can be concluded that the radiation power of each beam is related to the surface impedance coefficients. For the last case, the direction and gain of the radiation beams are obviously deviated from the theoretical values. These problems are not only caused by unequal surface impedance coefficients but also by the coupling between the two coexisted radiation patterns. In order to solve these problems, the surface impedance coefficients $(a_1 \text{ and } a_2)$ and modulation periods $(p_1 \text{ and } p_2)$ should be optimized from the initial values. The radiation efficiency from CST Microwave Studio simulation under these conditions is 84% ($a_1 = a_2 = 0.5$), 71% ($a_1 = 0.6$, $a_2 = 0.4$) and 48% $(a_1 = 0.7, a_2 = 0.3)$. This means that the radiation efficiency is related to the balance of radiation energy from each beam.



Figure 7. The simulated radiation patterns with different modulation depth at 8.5 GHz.

The prototype of the dual-beam LWA based on squared modulation was fabricated, as shown in Figure 8a. The modulation pattern was designed with the parameters $X_s = 400 \Omega$, M = 0.4, $p_1 = 18 \text{ mm}$ (6 unit cells per period) and $p_2 = 30 \text{ mm}$ (10 unit cells per period). Here $a_1 = a_2 = 0.5$ is applied to ensure the equal radiation energy of each beam. According to Equation (11), the modulation pattern

with varying *b* was designed to match the distribution of surface impedance, as shown in Figure 8b. The measurement of S-parameters was conducted by using an Agilent N5227A microwave vector network analyzer (VNA, Agilent Technologies Inc., Santa Clara, CA, USA), and the simulated and measured results are presented in Figure 8d. The reflection coefficient of the proposed antenna is almost below -10 dB in the frequency range of 8 GHz to 8.6 GHz, which means a good impedance match is provided. Meanwhile, S₂₁ keeps below -10 dB, indicating that little energy is transmitted to the output port. The far-field pattern was measured in the anechoic chamber with a linearly polarized horn antenna as the transmitting antenna, as shown in Figure 8c. The measured gain of the antenna at 8.5 GHz is shown in Figure 8e. The directions and beam widths of the measured main beams are in close agreement with the simulated patterns. The measured gains are 10 dBi and 8.2 dBi at

 $\theta_1 = -30^\circ$ and $\theta_2 = 18^\circ$, respectively, and the radiation efficiency is about 83%. This may be caused by

the additional loss brought by manufacturing error.



Figure 8. (a) The photograph of the fabricated antenna, (b) the schematic diagram of the measurement system, (c) modulated geometrical parameter *b* along the aperture, (d) the measured and simulated S-parameters, (e) the measured and simulated gain at 8.5 GHz.

4. Conclusions

The design method of LWA utilizing SquMRS is proposed and verified in this paper. Compared with SinMRS, SquMRS exhibits the advantage of simple structure with few kinds of unit cells. A

novel one-dimensional LWA based on SquMRS was designed to generate dual-beam radiation by superposition of individual modulation pattern for each beam. This approach can not only produce multi-beams but also control the radiation energy of each beam. Due to the coupling between the two coexisted radiation patterns, some discrepancies can be found between the simulated and theoretical results, which can be solved by optimizing the surface impedance coefficients and modulation periods from the theoretical values. Good agreement between experiment and simulation has been observed. It is much more convenient to control the tunable unit cells simultaneously when SquMRS is applied to achieve beam scanning capability at fixed frequency. Meanwhile, it also effectively reduces the insertion loss and cost by decreasing the number of tunable components.

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