A Tunable Impedance Surface Performing as a Reconfigurable Beam Steering Reflector

Dan Sievenpiper, James Schaffner, Robert Loo, Gregory Tangonan, Samuel Ontiveros, and Rick Harold

***Abstract*—We describe a reconfigurable microwave surface that performs as a new kind of beam steering reflector. The surface is textured with an array of tiny resonators, which provide a frequency-dependent surface impedance. By tuning the individual resonators, the surface impedance, and thus the reflection coefficient phase, can be varied as a function of position across the reflector. Using a reflection phase gradient, the surface can steer a reflected beam. As an example, we have built a simple mechanically tuned surface in which physical motion of only $1/100$ wavelength generates a sufficient phase gradient to steer a reflected beam by ± 16 degrees. To steer to greater angles, the surface can be configured as an artificial microwave grating, capable of ± 38 degrees of beam steering. The concept of the tunable impedance surface demonstrated here can be extended to electrically controlled structures, which would permit more elaborate reflection phase patterns, and provide more capabilities, such as the ability to focus or steer multiple beams.**

***Index Terms*—Antenna arrays, impedance sheets, reconfigurable antennas, reflector antennas, scanning antennas.**

I. INTRODUCTION

STEERABLE antennas today are found in two common configurations: those with a single radiator or reflector that is mechanically steered using a gimbal, [1] and those with a stationary array of electronically phased radiating elements. [2] Both have their own shortcomings, and the choice of antenna used is often a tradeoff between cost, speed, reliability, and other aspects of radio frequency (RF) performance. Mechanically steered antennas are inexpensive, but moving parts can be slow and unreliable, and they can require an unnecessarily large volume of unobstructed free space for movement. Active phased arrays are faster and more reliable, but they are much more expensive, and can suffer from significant losses due to the complex feed structure required to supply the RF signal to each radiating element. Losses can be mitigated if an amplifier is included in each element or subarray, but this solution also increases cost.

One alternative to a traditional phased array is to use a reflectarray geometry [3]–[7], and replace the lossy corporate feed network with a free space feed. The actively phased elements operate in reflection mode, and are illuminated by a single feed antenna. The array steers the RF beam by forming an effective reflection surface defined by the gradient of the reflection phase across the array. Using present techniques, such a system still requires a large number of phase shifters or other tuning techniques for active steering. What is needed is a reflective surface, in which the reflection coefficient phase could be arbitrarily defined, and easily varied as a function of position. Ideally, the surface would be less expensive than a comparably sized array of conventional phase shifters, yet maintain comparable RF performance. Such a surface could behave as a generic reconfigurable reflector, with the ability to perform a variety of important functions including steering or focusing of one or more RF beams.

With this goal in mind, we have developed such a reconfigurable reflector based on a high-impedance surface, [8], [9] which is a conductive sheet covered by a two-dimensional array of tiny resonant cavities. This type of surface has previously been applied to low-profile antennas [10], [11] and other RF devices [12]. The concept behind the reconfigurable reflector is to individually tune an array of cavities according to a specified pattern on the surface. We have built a mechanically tuned structure that is capable of steering a reflected beam by nearly ± 40 degrees at 3.1 GHz using a physical motion of only 1 mm, or $1/100$ wavelength. Although we have demonstrated a mechanically tuned structure, the concept can also be used with other methods of tuning, for example, by using electrically tuned capacitors.

II. RESONANT TEXTURED GROUND PLANE

The reconfigurable reflector is based a resonant textured ground plane, often known as a high-impedance surface. This new electromagnetic structure has two important RF properties, which have previously been applied to nonsteerable antennas. It suppresses propagating surface currents, [13], [14] which can improve the radiation pattern of antennas on finite ground planes. It also provides a high-impedance boundary condition, acting as an artificial magnetic conductor, which allows radiating elements to lie in close proximity to the ground plane without being shorted out. It is related to other well-known electromagnetic structures such as the corrugated surface, [15]–[21] and the photonic band gap surface [22], [23]. In this work, we have extended the utility of the high-impedance surface by making it tunable.

The high-impedance surface typically consists of a flat metal sheet, covered with an array of tiny, coupled resonant cavities, each much smaller than a wavelength in size. The cavities often take the form of small thumbtack-like protrusions on the flat metal ground plane. The surface is easily built using standard multi-layer printed circuit board techniques, in which small metal patches on one or more upper layers are connected to a ground plane on a lower layer by metal plated vias. One example of such a structure is shown in Fig. 1. For a more detailed description and analysis of the high-impedance surface, see [8], [9].

Manuscript received May 15, 2001. This work has been supported in part by the DARPA Reconfigurable Aperture (RECAP) program under Contract N6601-99-C-8635.

The authors are with the HRL Laboratories, LLC, Malibu, CA 90265 USA (e-mail: dfsievenpiper@hrl.com).

Publisher Item Identifier S 0018-926X(02)02613-3.

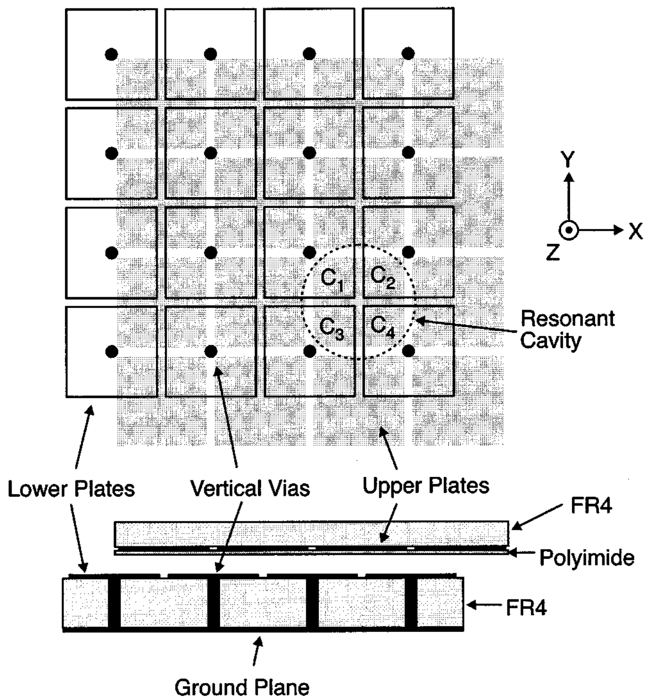


Fig. 1. A mechanically tunable impedance surface consisting of two printed circuit boards designed to slide against each other. The lower board resembles a conventional high-impedance ground plane. The back is solid metal, and the front is covered with a lattice of metal squares connected to the ground plane by metal plated vias. The upper board contains a lattice of metal tuning plates, separated by a polyimide insulator. When the two boards are pressed together, they form a lattice of tunable capacitors.

When the resonant cavities are much smaller than the wavelength of interest, the electromagnetic analysis can be simplified by considering them as lumped LC circuits. The proximity of the neighboring metal plates provides capacitance, while the conductive path that connects them provides inductance. The textured ground plane supports an electromagnetic boundary condition that can be characterized by the impedance of an effective parallel LC circuit, given by $Z_s = (j\omega L)/(1-\omega^2\mathrm{LC})$. The sheet inductance is $L = \mu t$, where μ is the magnetic permeability of the circuit board material, and t is its thickness. For a structure with parallel plate capacitors arranged on a square lattice, the sheet capacitance is $C = \varepsilon A/d$, where ε is the electric permitivity of the dielectric insulator, and A and d are the overlap area and separation, respectively, of the metal plates.

The surface has a frequency-dependent reflection phase given by $\Phi_r = \mathrm{Im}\{Ln((Z_s-\eta)/(Z_s+\eta))\}$, where η is the impedance of free space. Far from the resonance frequency, the surface behaves as an ordinary electric conductor, and reflects with a π phase shift. Near the resonance frequency, the cavities interact strongly with the incoming waves. The surface supports a finite tangential electric field across the lattice of capacitors, and the structure has high, yet reactive surface impedance. At resonance, $\Phi_r = 0$, and the surface provides the effective boundary condition of an artificial magnetic conductor. Scanning through the resonance condition from low to high frequencies, Φ_r varies from π, to zero, to $-\pi$. Thus, by tuning the resonance frequency of the cavities, one can tune Φ_r for a fixed frequency. By varying $\Phi_r(x,y)$ as a function of position on the surface, one can perform a variety of functions. For example, a linear phase gradient is equivalent to a virtual tilt of the reflector. A saw-tooth phase function transforms the surface into a virtual grating. A parabolic phase function could focus a plane wave onto a small feed horn, allowing the flat surface to replace a parabolic dish [24].

III. Tunable Impedance Surface

The tunable impedance surface that we used to demonstrate beam steering consisted of a matching pair of printed circuit boards, shown in Fig. 1. The first board contained a metal sheet covered with a lattice of thumbtack-like structures, resembling an ordinary high-impedance surface as described above. The back of this first board was a solid metal ground plane, and the front was covered with an array of metal patches, which were connected to the ground plane by metal plated vias. The structure was fabricated on FR4, a standard fiberglass-based printed circuit material. The second board contained a matching array of metal tuning plates, which were designed to overlap the metal patches on the first board. The tuning plates were supported on a sheet of FR4, and were covered by an insulating layer of Kapton polyimide. The two boards could be pressed together with the metal plates separated by the polyimide insulator, forming a lattice of parallel plate capacitors. The surfaces were designed to slide against each other, to allow adjustment of the overlap area between the matching sets of metal plates, and thus allow the capacitors to be tuned.

For our experimental structure, the square metal plates on both boards measured 6.10 mm, and they were distributed on a 6.35-mm lattice. The fixed board was 6.35-mm thick, and the conducting vias were 0.5-mm in diameter, centered on the square metal plates. The movable board was 1.57-mm thick, and the polyimide insulator that covered the tuning plates was 0.05-mm thick. Both boards measured 25.4 cm on each edge. To ensure uniform, intimate contact between the two matching surfaces, a vacuum pump was attached to the back of the fixed board. This evacuated the space between the boards by way of the hollow vias, and forced the two together.

By sliding the upper board across the lower one, the overlap area of the capacitors is changed. Since the resonance frequency of the individual cavities formed by the metallic plates and via structures depends on the capacitance, changes in the overlap area lead to changes in the resonance frequency. However, only motion that is parallel to the applied electric field contributes to a change in resonance frequency. This can be understood from the following argument: The resonance frequency of the cavities is given by $\omega = (1/(\sqrt{\mathrm{LC}}))$, where C is the effective capacitance produced by a combination of four separate capacitors $C_1 \ldots C_4$ indicated in Fig. 1. The mode that is excited in the cavities, and the circuit topology that produces the effective capacitance, depends on the polarization of the incoming wave. The circuit topology for two cases is shown in Fig. 2.

As an example, consider an incoming wave polarized along Y, referring to Fig. 1 for orientation. The effective capacitance is (C_1+C_2) in series with (C_3+C_4). If the top plate is moved in the $+Y$ direction, parallel to the applied field, then C_1 and C_2 are increased while C_3 and C_4 are decreased by the same amount,

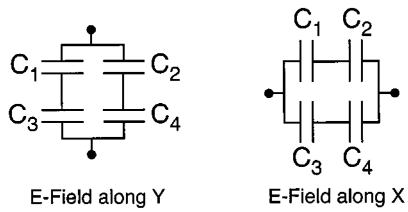


Fig. 2. The circuit topology that determines the effective capacitance depends on the direction of the applied electric field. As the tuning plate is moved, the values of the four capacitors are changed. The resonance frequency is only affected when the tuning plate is moved along the direction of a series connection.

as shown in Fig. 3. Since the motion occurs along the direction of pairs that are in series, the result is a net change in capacitance, and thus, a change in resonance frequency. Conversely, if the top plate is moved in the $+X$ direction, perpendicular to the applied field, then C_2 and C_4 are increased while C_1 and C_3 are decreased by the same amount. Since the motion occurs along the direction of pairs that are in parallel, there is no net change in capacitance, and no change in resonance frequency. The maximum effective capacitance, and thus the lowest resonance frequency, occurs when the upper plate is centered such that capacitors that are in series have equal value.

The resonance frequency of the textured surface is the frequency where Φ_r crosses through zero. For a fixed test frequency, a change in the resonance frequency of the surface appears as a change in Φ_r. To measure Φ_r as a function of frequency, we used a network analyzer and a pair of horn antennas, one for transmitting and the other for receiving. The use of separate transmitting and receiving horns is preferred for this experiment because it eliminates interference from internal reflections within the antennas. The horns were placed next to each other, both aimed at the tunable surface, and separated by a sheet of microwave absorber. Microwave energy was transmitted from one horn, reflected by the surface, and received with the other horn, while the phase of the reflected wave was monitored for various positions of the movable board. A reference measurement was taken using a flat metal surface, which has a reflection coefficient phase of π.

The reflection coefficient phase Φ_r of the tunable surface is shown in Fig. 4 as a function of frequency for ten different positions of the upper board, displaced in the direction of the applied electric field. By varying the overlap area of the capacitor plates, the resonance frequency is tuned from roughly 1.7–3.3 GHz. The series of scans shown corresponds to a total translation of one-half period of the textured surface, or 3.2 mm. The tuning range is limited by the maximum and minimum achievable capacitance, which depend on the area of the plates, the thickness of the insulator, and the fringing field in the surrounding medium.

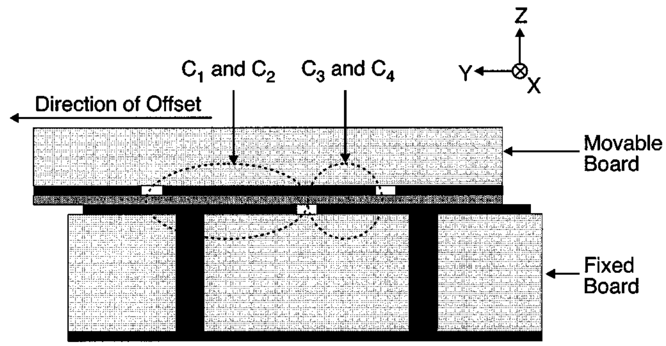


Fig. 3. As the top plate is moved, half of the capacitors are increased, while the other half are decreased by the same amount. If these capacitors appear in series to the applied electric field, the result is a net decrease in effective capacitance, and an increase in resonance frequency.

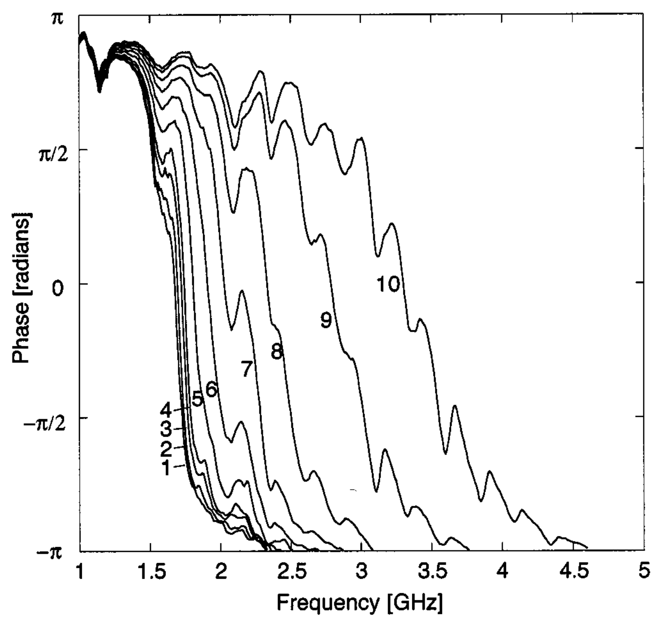


Fig. 4. The reflection coefficient phase of the tunable impedance surface for various positions of the top tuning board. Each successive curve represents a physical translation of 0.32 mm of the top metal plates. For curve 1, the two boards are aligned as shown in Fig. 1. For curve 10, the top board is translated by one-half period in the Y direction. The phase crosses through zero at the resonance frequency, which is tuned by varying the overlap area of the capacitor plates. The ripples in the curves are caused by spurious reflections from inside the anechoic chamber.

IV. Reflective Beam Steering

If the resonance frequency of the cavity array can be changed locally, then the reflection coefficient phase can be programmed to a specified function, $\Phi_r(x,y)$. By creating a monotonic, preferably linear phase gradient, the surface can steer a reflected beam. For our mechanically tuned reflector, this can be accomplished by a rotation of the upper printed circuit board with respect to the lower one, as shown in Fig. 5. From the discussion in the previous section, $\Phi_r(x,y)$ is only affected by translation of the capacitor plates in the direction parallel to the applied electric field. Specifically, for a wave polarized along Y, only the component of translation in the Y direction is relevant, and the translation along X has no effect. For each individual capacitor plate, a small rotation of the upper board results in a translation in Y that is roughly a linear function of X, but is largely independent of Y. Thus, rotation generates a monotonic gradient in $\Phi_r(x,y)$ in the direction perpendicular to the applied electric field, which is equivalent to a virtual tilt of the surface. Only a small mechanical motion

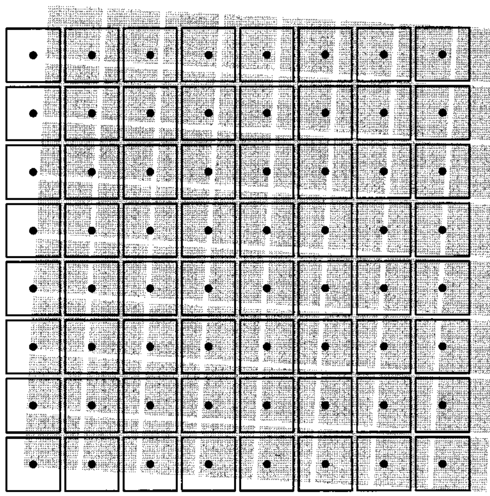


Fig. 5. Method of creating a phase gradient by rotating the upper tuning board. The overlap area in the vertical direction varies linearly as a function of horizontal position, creating a horizontal gradient in reflection coefficient phase across the surface.

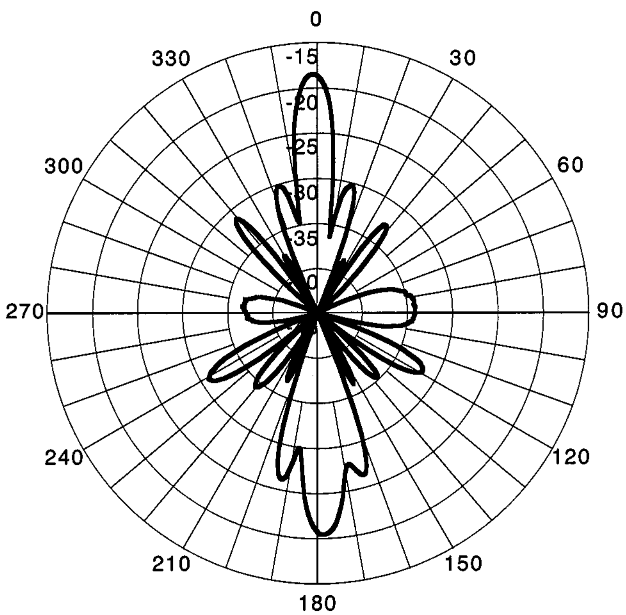
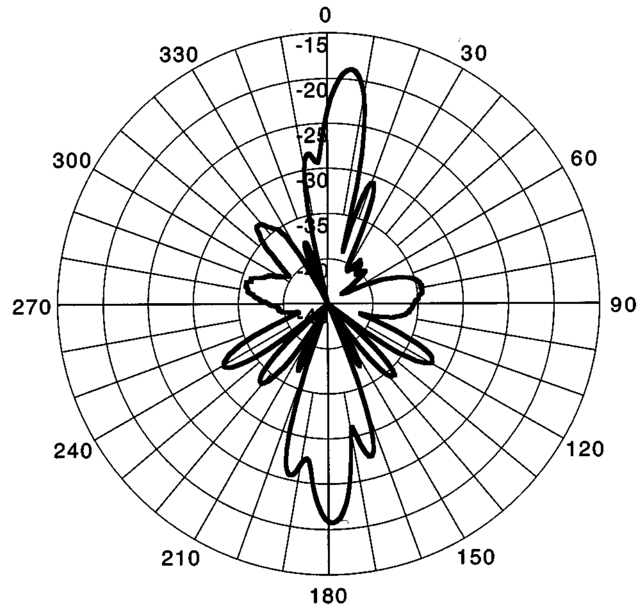


Fig. 6. Reflection pattern of the tunable surface with no phase gradient. The two lobes at 0 and 180 degrees occur when the front or back of the surface is directly facing the transmitting and receiving horns.

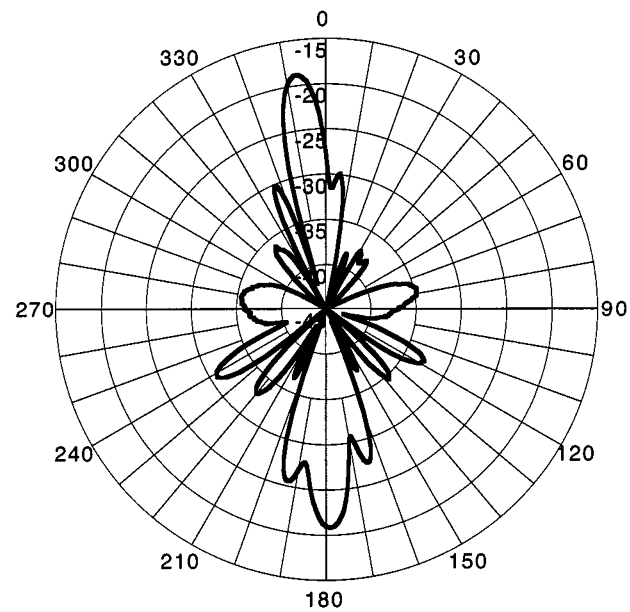
is required, since the maximum displacement needed at the edge of the board is only one-half of the lattice period.

To measure the beam steering properties of the tunable reflector, it was mounted vertically on a rotating pedestal and the reflection magnitude was measured as a function of incidence angle using two stationary horn antennas. Adjustment screws placed at two corners of the surface allowed independent control of both the relative orientation and the relative vertical displacement of the two boards. Measurements of the reflection pattern were taken for various positions of the movable board, at around 3.1 GHz near the center of the tuning bandwidth.

With the two boards aligned, $\nabla\Phi_r(x,y) = 0$, and the angle of reflection is equal to the angle of incidence. The reflection magnitude as a function of incidence angle is shown in Fig. 6.



(a)



(b)

Fig. 7. (a) Reflective beam steering using a phase gradient. The main lobe at 8 degrees indicates that the surface reflects an incoming wave back toward its source when it is rotated 8 degrees from normal. The same phase gradient would reflect a normally incident wave to an angle of 16 degrees. (b) Beam steering to the opposite direction using an opposite phase gradient. The surface can steer to any angle between these extremes by adjusting the direction and magnitude of the gradient.

As expected, the reflection is strongest at 0 and 180 degrees when the front and back surfaces of the reflector are directly facing the horns. The lobes at other angles are due to reflections from the rotating stage, the edges of the boards, the adjustment screws, the walls of our anechoic chamber, and other objects. The asymmetry in the reflection magnitude and angular profile between the front and back sides of the pattern is due to an acrylic vacuum plate attached to the back of the reflector, which holds the two printed circuit boards together. The difference in reflection phase between the two surfaces also contributes to

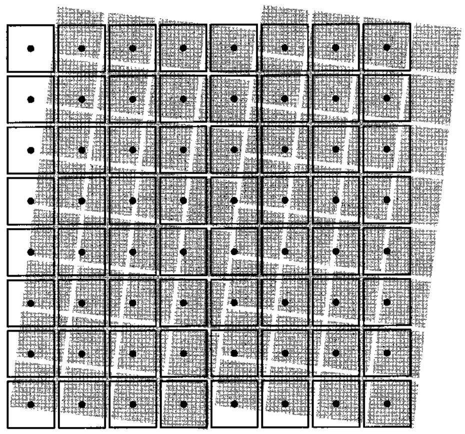


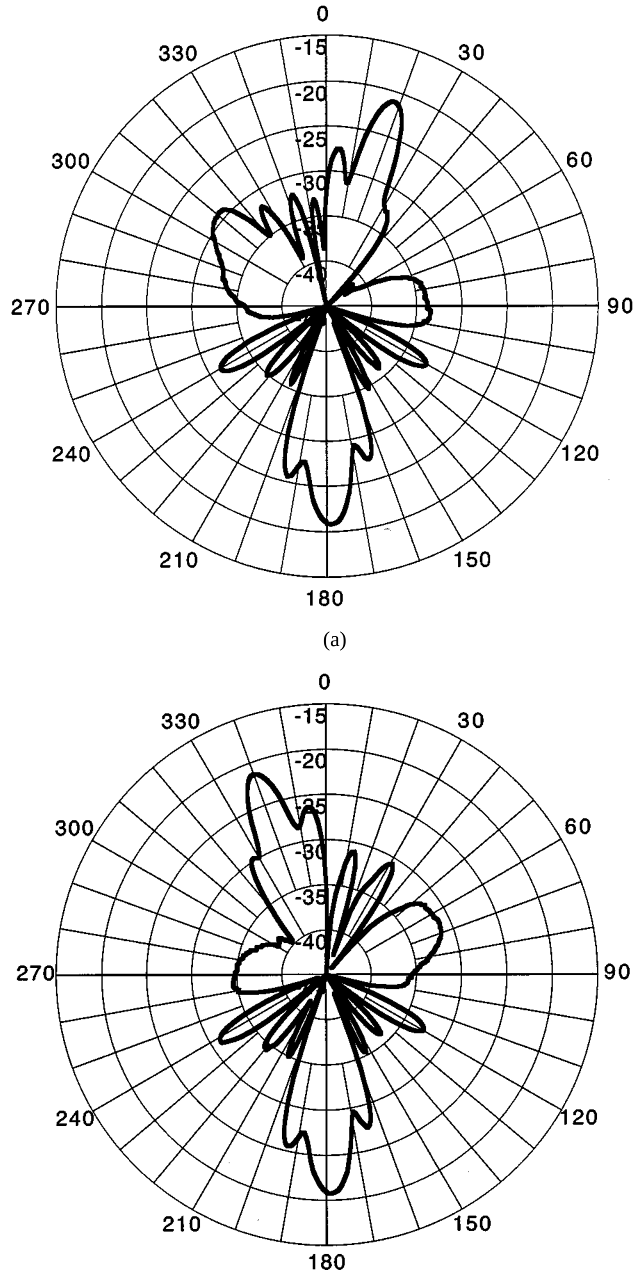
Fig. 8. Method of creating a steeper phase gradient by adding a discontinuity of 2π. The surface resembles a two-period microwave grating.

this asymmetry, because it affects the way the reflected waves interfere with other reflections from the surroundings.

When one board is rotated against the other, a Y-polarized wave will see a gradient in the reflection coefficient phase $\partial\Phi_r/\partial x$ which causes a normally incident beam with wavelength λ to be reflected to an angle in the XZ plane given by $\theta = 2\tan^{-1}((\lambda/2\pi)(\partial\Phi_r/\partial x))$. In our experimental setup previously described, the transmitting and receiving horns are stationary while the surface itself is rotated, so the reflection is seen when the surface is at an angle of $\theta/2$.

The reflected beam patterns for two different relative orientations of the two plates are shown in Fig. 7. The main lobes can be seen at angles of about ± 8 degrees and by rotating the upper surface between these extremes, the reflection angle can be tuned in an analog fashion. Of course, the lobe in the backward direction still appears at 180 degrees, because the back of the surface is untextured. Since the reflection phase of the front surface is a function of position, it forms a virtual reflection surface that is tilted with respect to the actual surface. It should be recalled that because the transmitting and receiving horns are stationary and mounted next to each other, the main lobes of the reflection pattern indicate angles at which a plane wave is reflected directly back toward its source. A normally incident plane wave would therefore be reflected to twice the angle measured in this experiment, and could be steered over a range of ± 16 degrees.

Because the resonance frequency is not a perfectly linear function of the displacement, as seen in Fig. 4, the maximum useful range of motion is actually less than one-half period. For the results described above, the difference in displacement between the two edges of the structure was roughly 1 mm, or $1/100$ wavelength. The two circuit boards were positioned so that over a path traversing from one side of the surface to the other, Φ_r varied over the range indicated by curves $7\ldots 10$ of Fig. 4. The higher frequency range between 2.3–3.3 GHz is preferred because the resonance frequency is roughly a linear function of displacement. This range also defines the bandwidth over which the surface can effectively steer a beam.



(a)

(b)

Fig. 9. (a) Beam steering using the two-period grating. The surface reflects an incoming wave back toward its source at an angle of 19 degrees from normal. It would reflect a normally incident wave to 38 degrees. (b) Beam steering to the opposite direction using an opposite phase gradient.

V. MICROWAVE GRATING

The maximum variation of Φ_r is limited to 2π when a monotonic phase function is used. This limits the beam steering capabilities of a surface with width W to $\theta = 2\tan^{-1}(\lambda/W)$. In order to steer to larger angles, a larger phase gradient must be used. Since phase can only be defined modulo 2π, periodic discontinuities of 2π must be included in the reflection coefficient phase function. Such a surface can be considered an artificial grating.

In order to build a microwave grating using our mechanically tuned surface, we sheared the movable board down the

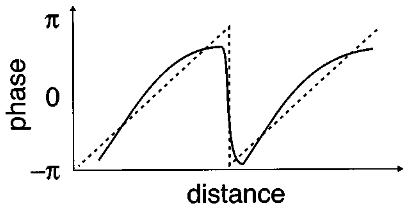


Fig. 10. Phase profile of the artificial grating, showing the reason for scattering to other directions besides the main lobe. The dotted line shows the ideal saw-tooth profile. The solid line indicates a more realistic profile for the mechanically tuned surface, including a nonlinear phase slope, and an imperfect discontinuity in the center.

center, and offset the two sides as shown in Fig. 8. This produced the necessary phase discontinuity to produce a two-period grating, which can provide a total phase variation of 4π across the surface. As seen in Fig. 9, the maximum reflection angle now occurs at ± 19 degrees. For a normally incident plane wave this corresponds to beam steering of ± 38 degrees. The beam could be steered to any angle within this range by adjusting the phase gradient, while maintaining the 2π discontinuity. A grating with a larger number of periods would provide beam steering to greater angles.

The patterns shown for this experiment exhibit significant scattering to other angles besides the main beam. This is not a grating lobe, but instead occurs because rotation of the upper board does not produce a perfectly linear phase function, as dictated by the functional dependence of the resonance frequency on the displacement of the capacitor plates. The problem is most severe at the 2π phase discontinuities, as shown in Fig. 10. This is not an intrinsic problem with artificial microwave gratings, but rather it is an artifact of the inexact phase function produced by our mechanical steering method. This effect could be diminished by having independent control over the resonance frequency of each individual cavity.

VI. Conclusion

We have described a tunable reflector that can provide beam steering over a wide angular range using a mechanical motion of much less than the operating wavelength. We have used our tunable surface to demonstrate beam steering using a simple phase gradient, and have also shown a two-period microwave grating that can achieve greater steering angles. Although our structure was mechanically tuned, it demonstrates the concept of using a two-dimensional array of small tunable resonant cavities to perform beam steeing functions. The same concept could also be used to build electrically tuned versions, by using electrically tuned capacitors for example. Other phase functions could also be applied, such as a parabolic function for focusing. With independent control over each individual capacitor, the surface could be programmed to perform a variety of functions, such as steering over wide angles, or subdividing the surface into smaller units to focus multiple beams onto one or more feed antennas.

A reconfigurable reflector could be used in many applications as a replacement for conventional steerable antennas. For example, a mechanically tuned surface could replace a gimbaled antenna, to significantly reduce the required movement and provide faster steering. A surface with electronically tuned capacitors could be fast enough to compete with the performance of phased arrays, with the potential of much lower cost. Furthermore, If the capacitors have sufficient tunability, such a surface could be reconfigured to work over a broad range of frequencies by simply applying a uniform capacitance offset. With the ability to produce multiple beams, and cover multiple frequencies, it could potentially replace several existing antenna systems with a single aperture.

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Dan Sievenpiper received the B.S. and Ph.D. degrees in electrical engineering from the University of California at Los Angeles, in 1994 and 1999, respectively. His doctoral research included photonic crystals, high-impedance electromagnetic surfaces, and novel antennas.

He is currently working as a Senior Research Staff Scientist at HRL Laboratories, Malibu, CA, in the areas of mobile wireless communication, reconfigurable antennas, and novel electromagnetic materials. He has 11 refereed journal publications.

James Schaffner (S'74–M'79–SM'97) received the B.S., M.S., and Ph.D. degrees in electrical engineering from the University of California at Los Angeles, in 1978, 1979, and 1988, respectively.

He is a Senior Research Scientist at HRL Laboratories, Malibu, CA, where he has worked since 1988. From 1978 to 1988, he was with the Hughes Aircraft Company Missile System Group, where he worked on microwave and millimeter wave antenna arrays, antenna feed networks, and MMIC circuits for transmit/receive modules. His current research interests are in components, systems, and metrology for optical and millimeter wave broadband communications channels.

Robert Y. Loo received the Ph.D. degree in electrical engineering from the University of California at Los Angeles in 1976.

He joined HRL Laboratories, Malibu, CA, in 1976, and has been responsible for the development of GaAs solar cell technology, as well as having developed a model that suggested methods to improve the radiation resistance of the cell. He has also been developing high speed InGaAs–InP P-I-N photodetector arrays and optoelectronic selector switches for the phased array antenna and WDM optical networks for subcarrier multiplexing signal applications. During recent years, he led an RF MEMS team at HRL in developing metal-to-metal contact RF MEMS switches for antenna and microwave applications. He has 29 years experience in semiconductor materials and photonic device technology, and has co-authored 45 technical papers on photonics devices, optical control of phased arrays, GaAs solar cells, and RF MEMS technologies.

Gregory Tangonan received the B.S. degree in physics from Ateneo de Manila University, Philippines, and the Ph.D. degree in applied physics, from California Institute of Technology, Pasadena, in 1969 and 1975, respectively.

He joined HRL Laboratories, Malibu, CA, in 1971 after receiving the Howard Hughes Doctoral Fellowship for studies at the California Institute of Technology. He led the effort which built the first optoelectronic (OE) crossbar (8×8) switch; these OE switches are presently used in radar beam steering, all-optical networks, and signal processing programs within Hughes. He led the team that demonstrated the world's first optically controlled phased array antenna, and he initiated the shift in emphasis at HRL into RF MEMS and reconfigurable antennas for commercial wireless access. The Communications and Photonics Laboratory that he manages focuses on the exploitation of novel antennas for mobile wireless access, wideband fixed wireless for broadband access to interactive services, laser technology for communications and high power applications, and photonic beamsteering and signal distribution for RF and microwave system. He is the coauthor of many published papers and presentations in the fields of fiber optics, integrated optics, laser spectroscopy and materials, and he has 32 U.S. patents.

Samuel C. Ontiveros is an Engineering Specialist at HRL Laboratories, Malibu, CA. He joined HRL in 1979, working on MOVPE materials growth for GaAs solar cell projects. Currently, he is working on RF MEMS switches for microwave and antenna projects for wireless communications, and is also doing packaging for photonic devices. Before joining HRL, he worked at Raytheon Missile Systems, Oxnard, CA, building a missile test station and in telemetering assembly. After that, he worked in Bunker Ramo in Westlake Village, CA, on precision navigation buoys for the navy.

Rick Harold received the B.S. degree in applied physics from California State University, Northridge, in 1992.

He has worked with electro-optics and laser-based ultrasonic systems. He is currently working with staff members on the development of advanced antenna configurations, and is a Senior Development Engineer at HRL Laboratories, Malibu, CA, where he has worked since 1984.