Study of a CPW Inductively Coupled Slot Antenna

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***Abstract*—Coplanar waveguide (CPW) fed slot antennas are attractive due to fabrication simplicity and ease of integration with active devices. In this paper, a new concept for exciting slots with a CPW line based on inductive coupling is presented. It is described how this coupling structure can be designed to tune the impedance of the antenna over a wide range. Single elements and arrays designed for 5- and 25-GHz operation are described. Simulated and measured characteristics are presented. This new coupling topology is particularly suitable for series-fed array configurations and broad-band design.**

***Index Terms*—Coplanar waveguide, slot antenna.**

I. Introduction

At microwave- and millimeter-wave frequencies, the interest for the coplanar waveguide (CPW) transmission line has increased significantly in recent years. They are preferable for monolithic microwave integrated circuits (MMIC's) since no backside processing (via holes) are required for integration with devices [1]. Accordingly, many antenna elements suitable for a CPW-fed configuration have been proposed, the slot antenna being one of the most attractive solutions. One of the main issues with CPW-fed slot antennas is to provide an easy impedance matching to the CPW line. So far, several impedance tuning techniques based on a change of slot dimensions, coupling mechanism or both have been reported in the literature. Among them we can quote the one-wavelength center-fed slot antenna [2] [Fig. 1(a)], the one-wavelength offset-fed slot antenna [3] [Fig. 1(b)], the half-wavelength capacitively-fed slot antenna [4] [Fig. 1(c)], and the multiple folded slot antenna [5] [Fig. 1(d)]. In this paper, a new inductively coupled slot antenna configuration is presented. With this technique, the impedance matching with the CPW line is facilitated by the possible adjustment of several geometric parameters.

The paper is structured as follows. In Section II, the topology and concept of the proposed inductively coupled slot are examined. Section III discusses how the antenna impedance varies as a function of the geometric parameters. Section IV presents four antenna prototypes that have been realized, including measured and predicted return loss and radiation patterns. All the predictions were obtained with the method of moments using HP-Momentum™.

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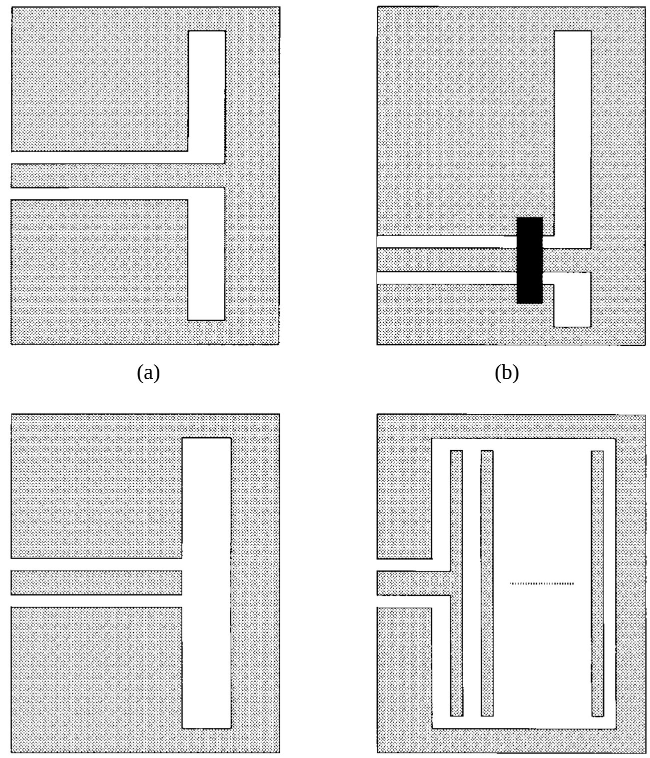


(a) (b) (c) (d)

Fig. 1. Existing topologies of CPW-fed slot antennas. (a) One-wavelength center-fed. (b) One-wavelength offset-fed. (c) Half-wavelength capacitively-fed. (d) Multifolded.

II. Geometry and Concept of the Inductively Coupled Slot Antenna

The proposed antenna is formed by etching two half-wave slots located symmetrically on both sides of the CPW line (Fig. 2), at a distance G from the outer CPW edges. The slots have a bent section of length l in order to increase the coupling to the CPW line. The magnetic field flux of the CPW's propagated wave through the slots excites them by inducing an electric field. Due to the opposite direction of the magnetic field on both sides of the CPW line (Fig. 3, [6]) the electric fields induced in both slots have the same direction in the vertical segments of length L-l and the radiation pattern is broadside. In the horizontal segments of length l, the electric field in the two slots has opposite directions and the radiation from the bent sections cancels in the broadside direction. Actually their radiation only contributes to the cross polarization.

The coupling region identified in Fig. 2 can be seen as a section of two coupled transmission lines, one being the feeding CPW line and the other one being formed by the

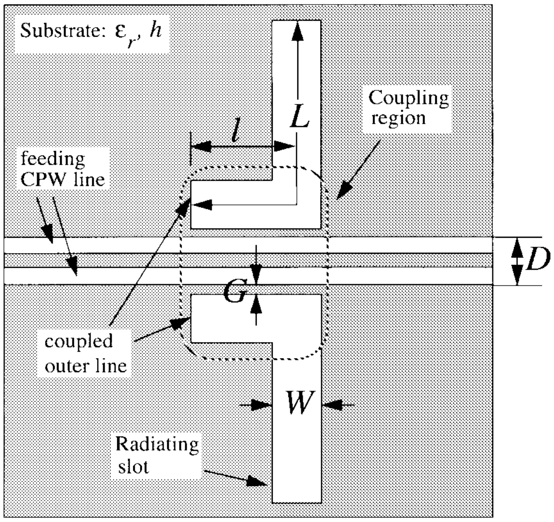


Fig. 2. Geometry of the inductively coupled slot antenna.

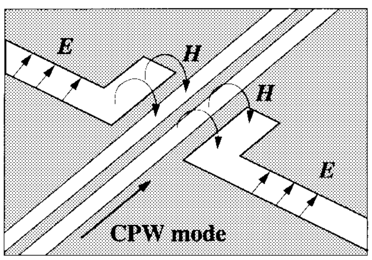


Fig. 3. Electric and magnetic field configuration in the coupling region.

two outer slots of width W. The line formed by the outer slots is terminated at one end by a short circuit and at the other end by the vertical radiating slots forming an antenna with radiation resistance R_s. Each of these vertical sections is approximately half-wavelength long and it can be modeled as lossy resonant circuit. The presence of a short circuit at the end of each outer horizontal slot favors a strong current around the slots and, thus, inductive coupling between these slots and the feeding CPW line. The radiation of the antenna will, therefore, be mostly dependent on the feeding CPW line's current rather than on its voltage. Consequently, the antenna loading can be modeled as an equivalent impedance in series along the feeding line. At resonance, the equivalent resistance is $R_{\rm ant}$. Of course, $R_{\rm ant}$ and its relationship to R_s depend on the dimensions of the coupling region (l, D, G, and W), the dimensions of the vertical radiating slots (W and L-l) as well as on the substrate thickness h and relative permittivity ε_r. Finally, we can see that there is a magnetic plane of symmetry, perpendicular to the plane of the structure in Fig. 3, which is cutting the center of the CPW line's center conductor. Theoretically, this prevents the excitation of the common mode (slot mode) in the CPW line and it minimizes the need for air bridges.

III. Antenna Impedance Tuning

The sensitivity of input impedance to the geometry was studied for three parameters that were found to have most critical influence. These parameters are:

TABLE I
TRENDS OF RADIATION RESISTANCE AS A FUNCTION OF THE GEOMETRICAL PARAMETERS

G ↗	⇨	R_{ant} ↘	
l/L ↗	⇨	R_{ant} ↗	Q ↗
W ↗	⇨	R_{ant} ↘	Q ↘

- G, the width of the gap between the slot and the CPW line;
- l/L, the ratio between the CPW-coupled section and the overall length of the slots;
- W, the slot width.

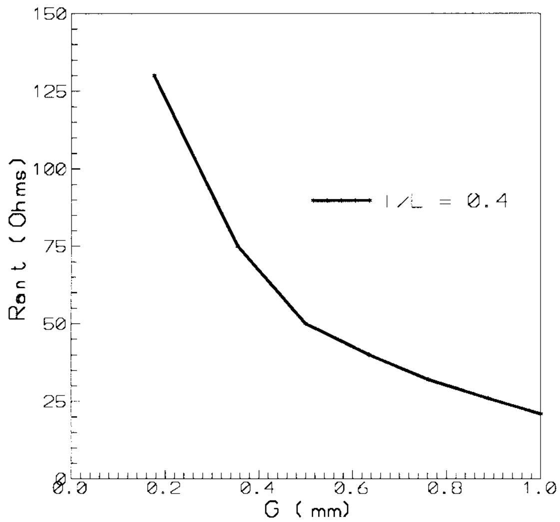
Given the complexity of the structure, these dependencies were studied quantitatively using numerical modeling. Qualitative descriptions of the observed trends are summarized in Table I. The propagated wave in the feeding CPW line has its fields well confined around the line with a rapid decay in the plane transverse to the direction of propagation. Therefore, the coupling region is limited to a small area around the line. The coupling to the outer horizontal slots is proportional to the flux of the magnetic field generated by the incident CPW wave through the aperture of the outer slots in the coupling region. Thus, it is directly affected by the metallic strip of width G. Increasing G decreases the coupling, which decreases the radiation resistance $R_{\rm ant}$.

The coupling section of length l does not contribute significantly to the radiation. Therefore, for a given slot length L, increasing l decreases the length of the vertical radiating sections, which leads to an increase of the intrinsic radiation resistance R_s of the vertical slot and, therefore, an increase of the equivalent series resistance $R_{\rm ant}$ on the CPW line. The Q factor of the antenna at resonance will also increase. It should be recalled that l must be kept to a minimum value in order to minimize the level of cross polarization.

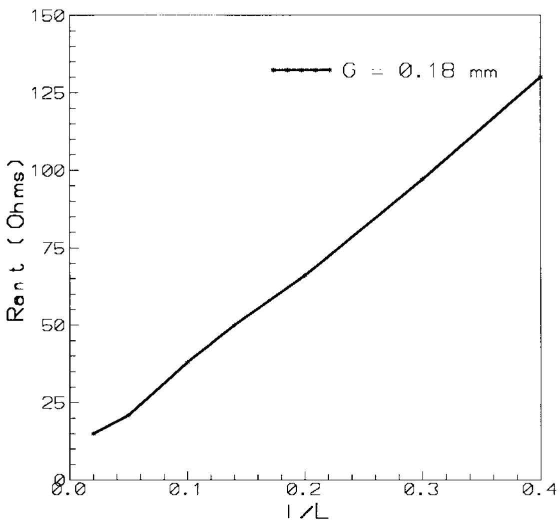
Increasing the slot width W leads to two opposite effects on $R_{\rm ant}$. First, it favors a better coupling between the CPW line and the antenna, which should lead to a larger $R_{\rm ant}$. On the other hand, a wider slot has a smaller radiation resistance (R_s), which also means a smaller $R_{\rm ant}$. In most of the cases studied, the second effect is predominant and the radiation resistance and the Q factor of the antenna decrease.

Table I shows no relationship between G and Q. In fact, the influence at a given of G on Q was found to be very small. This is probably due to the fact that G has a relatively small influence on R_s compared to the two other parameters. We can also notice that modifications of the geometry can have countering effects on the antenna performance, which allows for design flexibility. For instance, if we try to increase the gain of the antenna by decreasing the ratio l/L, $R_{\rm ant}$ will be modified. However, this modification can be compensated by reducing G. To a certain extent, it is also possible, by modifying the geometry, to adjust the bandwidth and reduce the cross-polarization level while maintaining the same value of $R_{\rm ant}$.

Numerical examples are shown in Fig. 4 for a specific type of substrate ($\varepsilon_r = 10.2$, $h = 1.25$ mm) and fixed slot



(a)



(b)

Fig. 4. (a) Variation of $R_{\rm ant}$ with respect to G for l/L fixed and (b) with respect to l/L for fixed G ($\varepsilon_r = 10.2$, $h = 1.25$ mm, $D = 1.15$ mm, $L = 12$ mm, $W = 0.5$ mm).

antenna and CPW line widths ($W = 0.5$ mm, $D = 1.15$ mm, $L = 12$ mm). In Fig. 4(a), it can be seen that for a length ratio $l/L = 0.4$, it is possible by varying G to adjust the antenna resistance at resonance anywhere between 20 Ω and 130 Ω. Similarly, for $G = 0.18$ mm, varying l/L caused a nearly linear increase of $R_{\rm ant}$ between 15 Ω and 130 Ω [see Fig. 4(b)]. In this case, l/L was varied but with a fixed value of L in order to keep the resonance frequency almost constant. The possibility to easily tune the antenna resistance over such a broad range is an interesting feature from the point of view of array feed design and active circuit matching.

IV. Experimental Validation

Three antenna prototypes have been designed using this new topology and built on 50-mil substrate ($\varepsilon_r = 10.2$). In order to characterize the antennas on a HP8753D network analyzer, a coax-to-CPW transition was fabricated and a de-embedding procedure using three different short circuit CPW lines was implemented to measure the input return loss of the antennas. Finally, their radiation patterns were measured in an anechoic chamber in the E and the H plane. The circuits are a single-slot element (5 GHz), a three-element uniformly excited linear array (5 GHz), and a three-element log-periodic array (4–5 GHz). The slots are bent over one fifth of their length ($l/L = 0.2$) with a distance between the slots ($2G+D$) kept small enough to keep the cross-polarization level under −30 dB. The characteristic impedance of the feeding CPW transmission lines is 50 Ω.

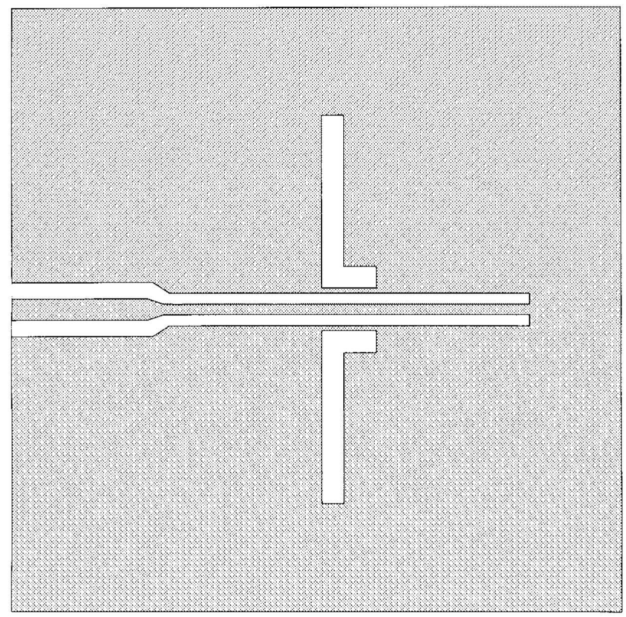
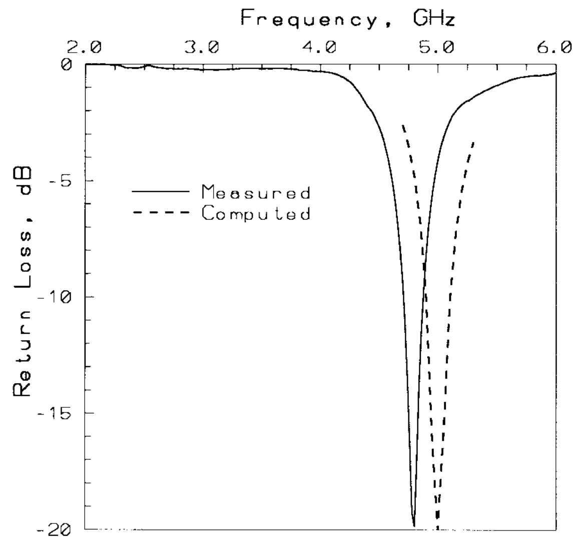


Fig. 5. Layout of single-slot element for 5-GHz operation ($\varepsilon_r = 10.2$, $h = 1.25$ mm, $D = 1.15$ mm, $L = 12.42$ mm, $l/L = 0.27$, $G = 0.2$ mm, and $W = 1$ mm).

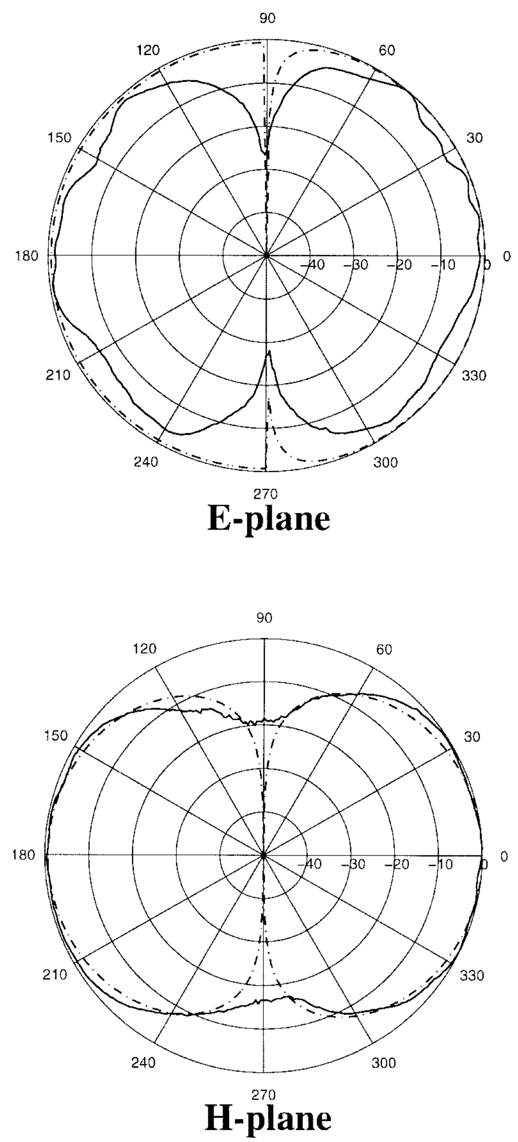
Another single-slot element was designed for operation at 25 GHz. The substrate in this case was 10-mil Alumina ($\varepsilon_r = 9.9$). The input return loss was measured with a network analyzer HP8510C and a probe station.

A. Single-Element Antenna—5 and 25 GHz

The antenna in Fig. 5 resonates at 5 GHz with an equivalent radiation resistance matched to the CPW line ($R_{\rm ant} = 50$ Ω). Its dimensions are: $L = 12.42$ mm, $l/L = 0.27$, $G = 0.2$ mm, $W = 1$ mm, and $D = 1.15$ mm. The transmission line is terminated by a short circuit located at a half-guided wavelength from the coupling region in order to present a short circuit in series with $R_{\rm ant}$ in this region. The return loss and radiation patterns are presented in Fig. 6(a) and (b). Except for deep minima along the directions parallel to the substrate, the predicted radiation pattern in the E plane is quasi-isotropic. Notice that in these directions, discrepancies between the predicted and measured E-plane patterns are expected due to the finite dimensions of the dielectric substrate and ground plane that are not taken into account in the simulation. In the H plane, the HPBW of 70° is comparable to the E-plane HPBW of a half-wave electric current dipole. The measured resonance frequency is slightly shifted (<4%) compared to



(a)



(b)

Fig. 6. (a) Return loss and (b) radiation patterns of the single-slot element for 5-GHz operation (·-·-·-: computed, ——: measured).

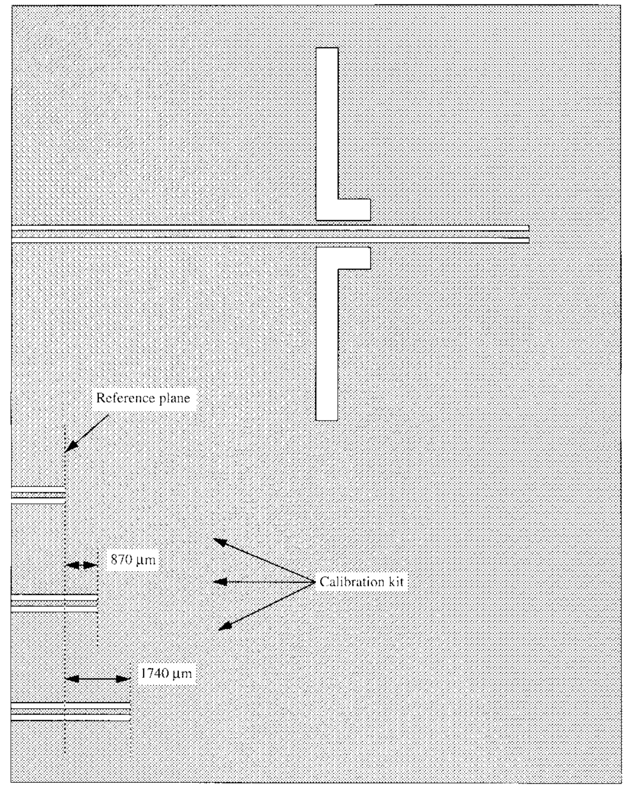


Fig. 7. Layout of single-slot element for 25-GHz operation ($\varepsilon_r = 9.9$, $h = 0.25$ mm, $D = 0.25$ mm, $L = 2.5$ mm, $l/L = 0.25$, $G = 0.04$ mm, and $W = 0.2$ mm).

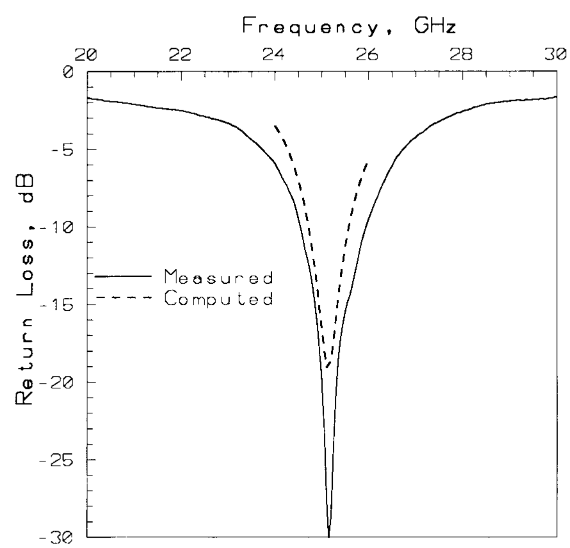


Fig. 8. Return loss and radiation patterns of the single-slot element for 25-GHz operation (·-·-·-: computed, ——: measured).

the prediction. The bandwidth for a −10-dB return loss is approximately 4%.

Due to the growing interest for millimeter-wave applications, an attempt was made to design a CPW slot element operating at such frequencies. The material used for our design is a 10-mil-thick alumina substrate (dielectric constant of 9.9), which is widely used in the industry for the fabrication of miniature hybrid microwave integrated circuits (MHMIC). A layout of the antenna and of the calibration standards used for the reflection coefficient measurements is shown in Fig. 7.

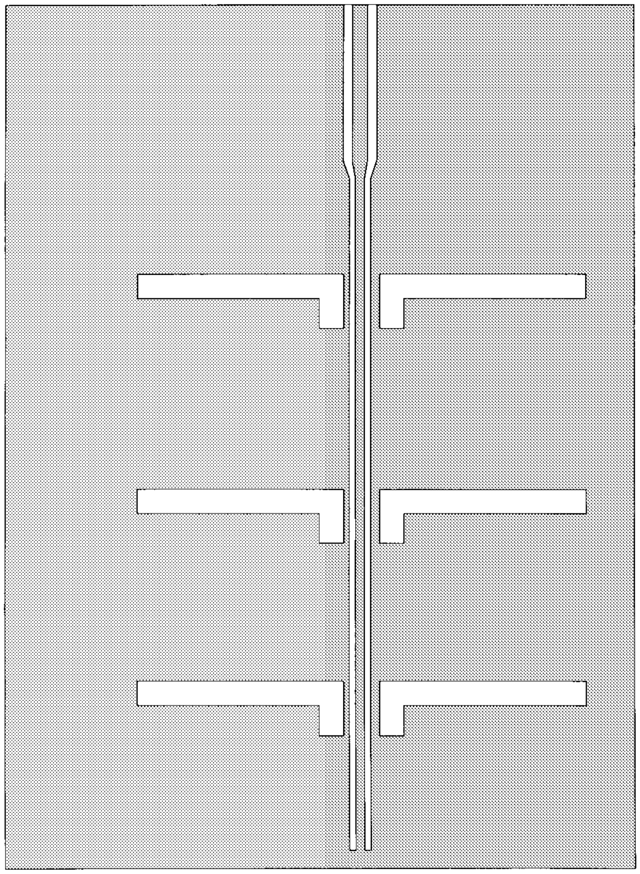


Fig. 9. Layout of the three-element uniformly excited linear array designed for 5-GHz operation ($\varepsilon_r = 10.2$, $h = 1.25$ mm, $D = 0.81$ mm, $L = 12.7$ mm, $l/L = 0.24$, $G = 0.4$ mm, and $W = 1$ mm).

Measurements were done with an HP8510C network analyzer. The predicted and measured return loss curves are presented in Fig. 8. The agreement between the predicted and measured resonance frequencies is excellent. However, the measured bandwidth for a 10-dB return loss is of 1.4 GHz (5.6%) while the predicted bandwidth is 0.8 GHz (3.2%). This may indicate a loss of efficiency.

B. Uniformly Excited Three-Element Array—5 GHz

In order to demonstrate that the proposed feeding technique is suitable for series-fed arrays, we designed a linear three-element antenna whose layout is shown in Fig. 9. The slots are excited in phase (broadside array) since the electrical length in the CPW line between the slots is λ_{cpw} at 5 GHz ($\lambda_{\text{cpw}} \approx 0.42\lambda_{\text{free-space}}$). All the lines have characteristic impedance of 50 Ω. At the resonance frequency, the radiation resistance of each simple antenna presents a series load on the transmission line. So, in order to have an antenna matched to a 50-Ω generator and a uniform array excitation, the value of each radiation resistance must be $R_{\text{ant}} = 50/3\ \Omega$. The chosen dimensions for the elements are: $L = 12.7$ mm, $l/L = 0.24$, $G = 0.4$ mm, $W = 1$ mm and $D = 0.81$ mm.

The measured and predicted results are shown on Fig. 10(a) and (b). The distance between the slots is approximately $0.42\lambda_{\text{free-space}}$, so the theoretical radiation pattern for the E plane has a narrow beam. The agreement with the predictions

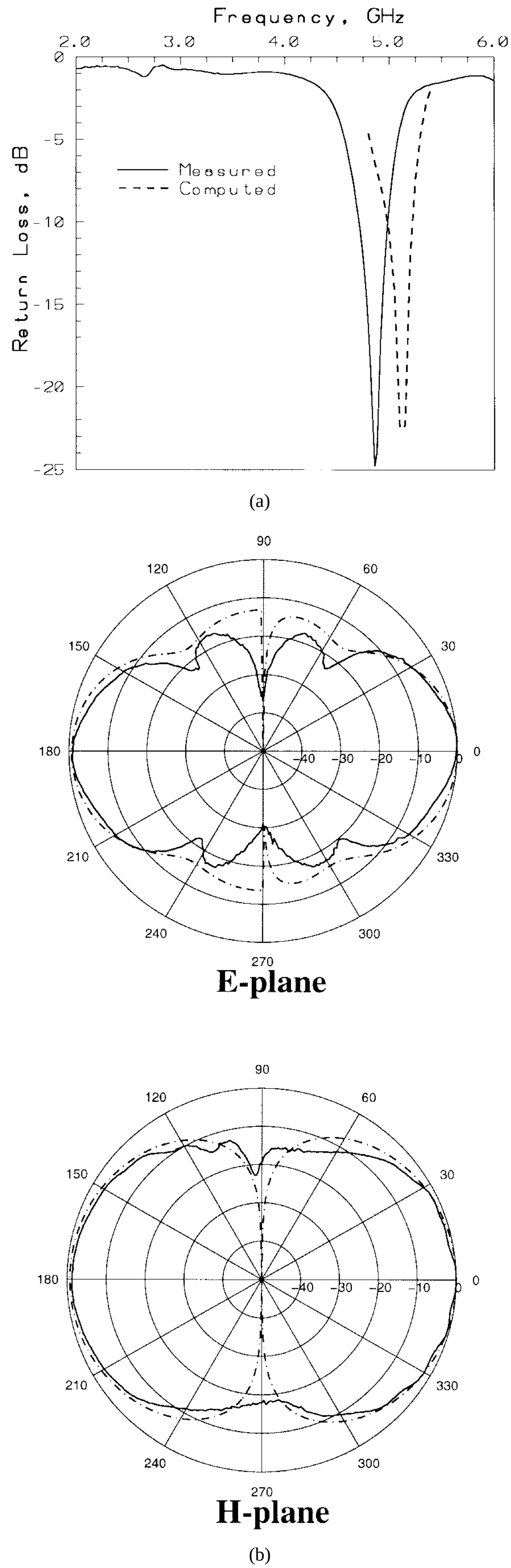
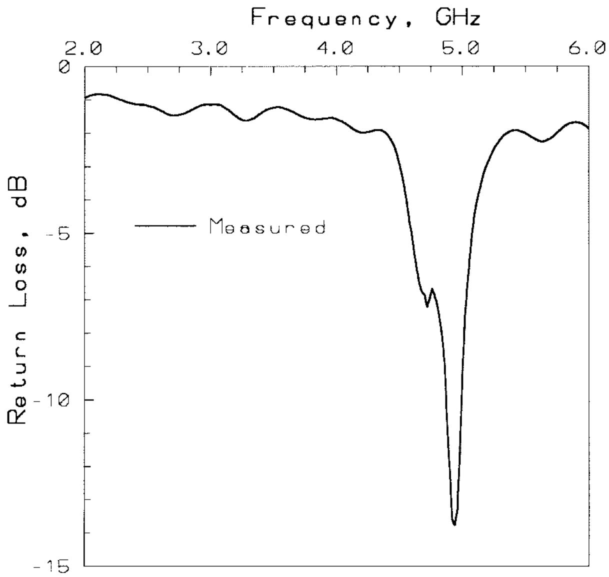
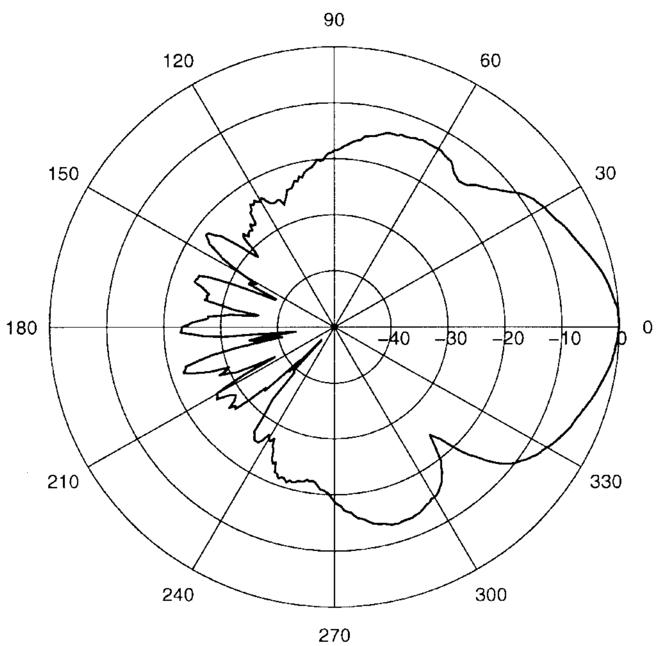


Fig. 10. (a) Return loss and (b) radiation patterns of the uniformly excited three-element slot array (·-·-·-: computed, ——: measured).



(a)



(b)

Fig. 11. (a) Return loss and (b) radiation patterns of the uniformly excited three-element slot array over a metallic reflector.

in the E plane is good, especially in the main beam. In the directions parallel to the substrate, the agreement is better than for the single element (case A) because the array factor is down by 12 dB. Again, the measured resonance frequency is slightly shifted (<4%) compared to the prediction. The bandwidth for a -10-dB return loss is nearly 6%.

In order to obtain unidirectional radiation, a planar reflector was placed under the antenna at $\lambda_{\text{free-space}}/4$ from the printed metallic surface. The drawback of this reflector configuration is that a TEM wave can propagate between the antenna and the reflector metallic plates. The measured performance for this configuration is given in Fig. 11(a) and (b). Following the addition of the reflector, the input resistance changed from 50 to 75 Ω. A practical design should, therefore, include the effect of the reflector in the impedance tuning of the individual slots. There is still a significant backlobe level in the E plane. This can be minimized by increasing the size of the reflector.

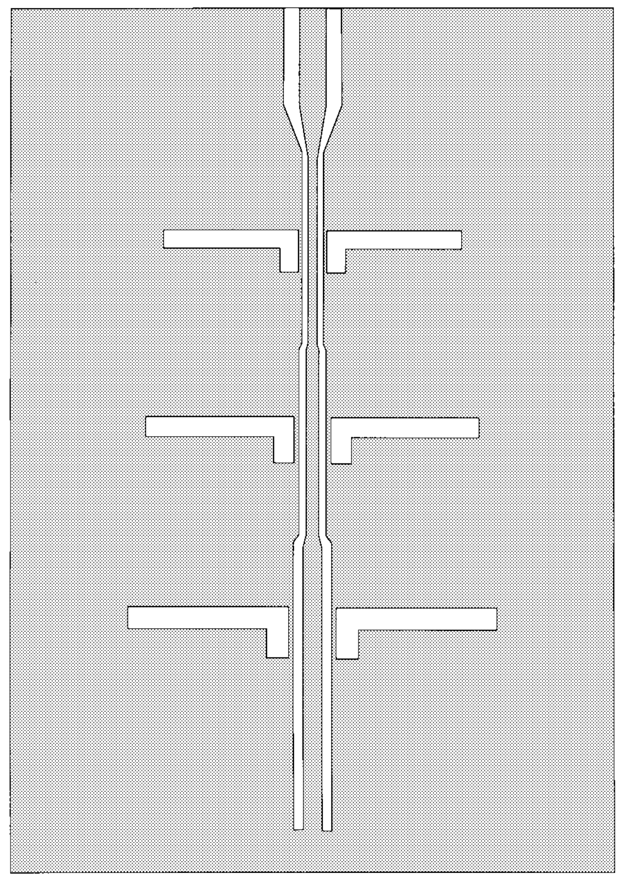
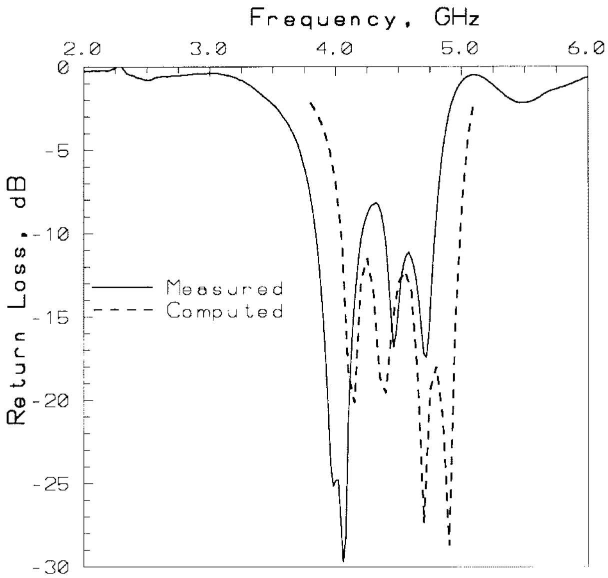


Fig. 12. Layout of the three-element log-periodic slot array (central element: $\epsilon_r = 10.2$, $h = 1.25$ mm, $D = 1.25$ mm, $L = 14.3$ mm, $l/L = 0.27$, $G = 0.22$ mm, $W = 1.15$ mm).

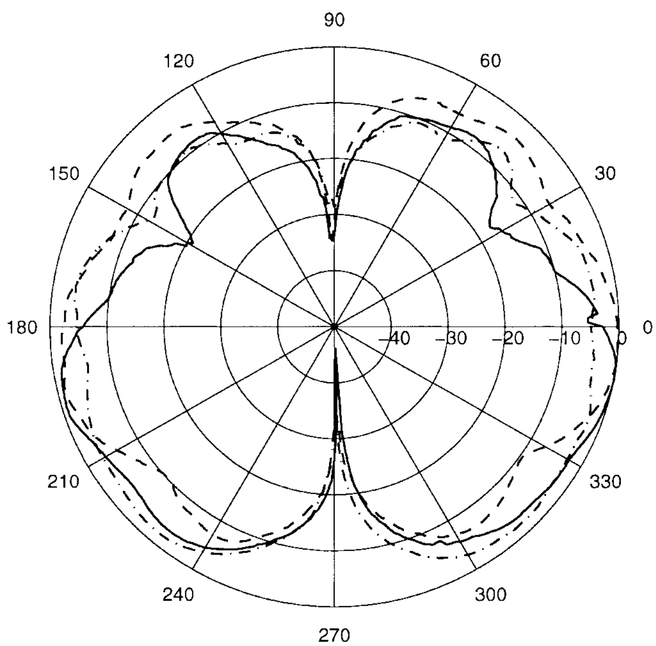
C. Log-Periodic Array—4–5 GHz

A well-known approach to increase the bandwidth of an antenna is to realize a log-periodic (LP) structure (Fig. 12). By applying a scaling factor τ to the first element dimensions we design the other elements. Notice that to be rigorous, this scaling procedure must be applied to the slot and the CPW transmission line in order to maintain the log-periodic frequency characteristics of the CPW-to-slot coupling region. Of course, it was not possible to scale the thickness of the substrate. In our design, we have scaled to coupling region in order to maintain a radiation resistance of 50 Ω in each element, when operated on its own. The scaling factor τ used in the design was 1.1.

At a given frequency, the adjacent active element(s) should be excited as much in phase as possible to maintain a broadside radiation. The best configuration is one in which antenna element n is located at d_n $(d_n = (2n-1)\lambda_{n\text{-cpw}}/2)$ from the CPW short circuit at the end of the feed line. $\lambda_{n\text{-cpw}}$ is the wavelength in the CPW line at the resonance frequency of element n and $n = 1$ corresponds to the low-frequency



(a)



(b)

Fig. 13. (a) Return loss and (b) radiation patterns of the three-element log-periodic array (·-·-·-: 4 GHz, - - - -: 4.48 GHz, ——: 4.74 GHz).

element. The electrical characteristics of a three-element LP antenna are presented in Fig. 13.

A bandwidth of about 20% $(\approx\tau - 1/\tau)$ was achieved due to the large scaling factor. This large factor also applies to the resonance frequencies of the adjacent narrow-band individual elements. This leads to a poor phase match between the phase of adjacent elements and in practice; there is nearly only one active element at each frequency. Consequently, the radiation characteristics of the array are similar to those of a single element and also suffer from ripples caused by the finite dimensions of the substrate. Using a scaling factor closer to 1 would lead to simultaneous excitation of more than one element and give better radiation patterns, of course at the expense of a greater number of elements or a narrower bandwidth.

V. CONCLUSION

A new feeding approach for slot antennas coupled to CPW transmission lines was investigated and validated using numerical modeling and experiments. An inductive coupling whose characteristics depend on the geometrical parameters of the structure excites the slots. The CPW-to-slot distance G, the normalized length of the coupling region l/L, and the width of the slot W appear as critical parameters allowing flexibility in the tuning of the antenna's characteristics. This was demonstrated in the paper by the design of four antenna prototypes. The characteristics of these prototypes were predicted with a good level of accuracy using moment-method solutions of a mixed-potentials integral equation. Experimental results indicate that improvement of the modeling could be achieved by taking into account the finite dimensions of the dielectric substrate and ground plane.

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