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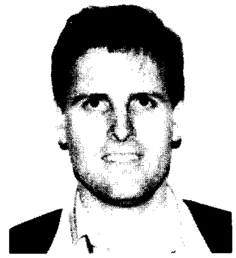

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Tuning Stubs for Microstrip-Patch Antennas

This excellent contribution comes from **Marius du Plessis** and **John Cloete**, of the University of Stellenbosch in South Africa. It relates to a similar article by Dave Pozar in the December, 1987, issue of this column. It was my pleasure to meet John Cloete in Seattle, at our Symposium. Readers wishing to contact the authors can write to the University of Stellenbosch, Department of Electrical and Electronic Engineering, Stellenbosch 7600, South Africa; Tel: 27 21 808 4377; Fax: 27 21 808 4981; jhcloete@firga.sun.ac.za (E-mail).

1. Abstract

A practical method for the simultaneous tuning, of both the resonant frequency and reflection coefficient, of a coaxially fed rectangular-microstrip patch is described. The use of two tuning stubs allows independent adjustment of the effective patch length and the effective position of the feed point. The effective length determines the resonant frequency of the patch, while the effective position of the feed point determines the input impedance. It is demonstrated that the method allows the adjustment of the feed-point reflection coefficient of a patch to less than -60 dB, at a frequency which is within 0.005% of specification.

2. Introduction

The narrow bandwidth of microstrip-patch antennas makes their input impedance very sensitive to manufacturing errors and substrate material tolerances. Thus, to meet precise frequency specifications, the patches may have to be tuned manually by, for example, perturbation of the effective resonant length. Furthermore, in an array of "identical" elements, the macroscopic inhomogeneities in substrate permittivity and thickness may require that the resonance frequency of individual elements be synchronized by tuning.

Pozar [1, 2] showed how the resonant frequency of a patch may be adjusted, by placing a tuning stub on one edge of a patch antenna, with the achievable tuning range dependent on the length and width of the stub. He also pointed out [1] that the practical tuning range is limited by degradation of the input reflection coefficient.

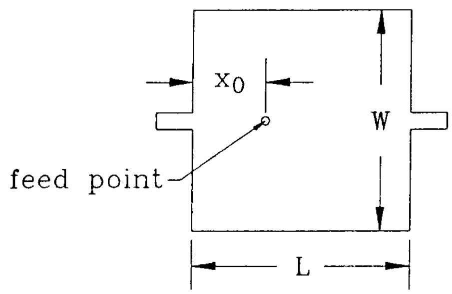

Figure 1. The location of the tuning stubs and feed pin on a rectangular microstrip-patch antenna.

This problem is overcome by the technique introduced in this paper. The idea is to use two tuning stubs, instead of one, which allows simultaneous tuning of the resonance frequency as well as the input impedance. The stubs are positioned on opposite edges of the patch, in line with the coaxial feed point of the patch, as illustrated in Figure 1. This makes both the effective resonant length and the effective position of the feed point adjustable. The shortening of either stub will decrease the effective electromagnetic length of the dominant-mode cavity formed by the patch, thereby increasing its resonant frequency. At the same time, the effective position of the feed point will be shifted towards whichever stub was shortened. For example, if the stub closest to the feed point is shortened, the effective position of the feed is shifted away from the center of the patch.

3. The patch length and feed position

To implement the technique, the patch dimensions must be designed with sufficient accuracy to place the resonant frequency within the tuning range. A variety of accurate computational methods are now available for this purpose [3]. However, the original transmission-line model proposed by Munson, consisting simply of two parallel radiating slots, separated by a transmission-line section [4, 5], allows the patch length to be estimated with accuracy sufficient for tuning. Typically, the accuracy of the patch length given by the transmission-line model is around 2%, which is well within the 10% tuning range reported by Pozar [1] for a. plastic substrate. To allow for the addition of the stubs, the patch dimension must be reduced by a few percent, depending on the stub dimensions.

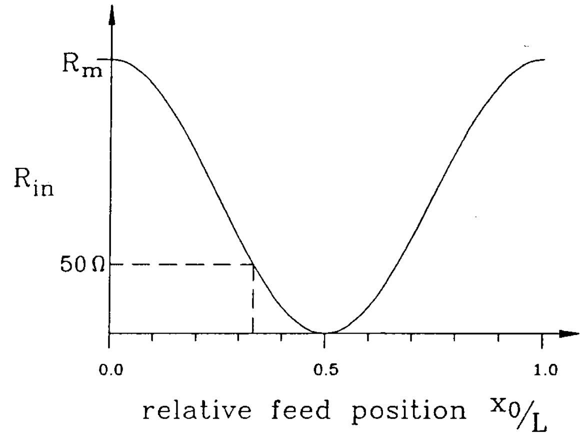

Figure 2. The input-resistance variation as a function of feed position for a microstrip-patch antenna.

The tuning technique also allows adjustment of the effective feed position. Accurate initial positioning will maximize the achievable tuning range, while poor initial placement can severely constrain the simultaneous tuning of the frequency and the input impedance. Figure 2 shows the approximate dependence of the dominant-mode resonant input resistance on the relative feed-pin position, x_0 / L, where x_0 is the distance from the edge of the patch, and L is the patch length. The approximation is given by the function [5, Equation (49)]

$$R_{in} = R_m \cos^2\left(\frac{\pi x_0}{L}\right). \tag{1}$$

The maximum resistance in Equation (1), R_m, can be approximated by

$$R_m = 60\lambda_0 / W, \tag{2}$$

where W is the patch width, and $\lambda_0$is the free-space wavelength [4], [5, Equation (7)]. More-accurate approximations are now available [6], and the maximum input resistance can also be obtained from the measurement of the reflection coefficient at the edge of a patch fed by a microstrip line. These formulas allow for simple estimation of the feed-pin position.

4. The patch-tuning procedure

With the feed pin at a suitable position, the patch can be tuned in an iterative manner. As discussed above, the patch is designed to resonate at a frequency which is a few percent below the specified operating frequency. This will enable the resonance frequency of the patch to be increased by systematic trimming of either of the stubs. The choice of which stub to trim depends on the input resistance of the patch, measured at the feed pin. In the following example, a 50 Ω coaxial-feed line is assumed for the patch.

If the measured input resistance at resonance (below the design frequency) is less than 50 Ω, the effective feed position is too close to the center of the patch, as can he deduced from Figure 2. The effective feed position must therefore be shifted towards the edge of the patch, by shortening the stub closest to the feed pin.

If the measured input resistance at resonance is too high (greater than 50 Ω), the feed pin is too close to the edge of the patch. The stub furthest from the feed pin must then be shortened, in order to shift the effective position of the pin towards the center of the patch, thereby lowering the input resistance.

If the initial feed position is sufficiently close to its correct position, i.e. 50 Ω is achieved immediately, the resonant frequency will still be slightly below the specified frequency. The patch can then be tuned to the correct frequency, by shortening each stub alternately.

The use of thin stubs will allow for very sensitive tuning, but over a limited range, while wide stubs will increase the range, but with less sensitivity.

5. Experimental results

As an example, results are presented for a patch which was required to be matched to a 50 Ω coaxial-feed line at precisely 3 GHz. For a square patch, with substrate thickness of 0.824 mm and relative dielectric constant of $\varepsilon_r = 2.53$, the transmission-line model, described by Carver and Mink [5], predicts a patch length of 30.50 mm for resonance at 3 GHz. The addition of tuning stubs will increase the effective length and, therefore, the patch must be fabricated a few percent shorter, to facilitate tuning. In this example, a patch length of 30 mm was selected, i.e. 1.6% less than the above value.

A 2.25 mm-wide microstrip line, with characteristic impedance approximately 50 Ω, was used to first measure the input resistance at the edge of the patch. Using Equation (1) and the measured maximum input resistance of 270 Ω, the ideal feed-pin location was found to be about 10.7 mm from the edge of the patch, as indicated in Figure 3. If Equation (2) is used, the maximum input resistance is predicted to be 200 Ω, and the corresponding difference in feed-pin position is 0.7 mm. The microstrip-feed line used for measuring R_m was then cut away to leave only a 5 mm stub, identical to the one on the other edge of the patch. With the feed pin in position, the tuning was then done in the manner described above.

Before tuning, the patch resonated at 2.9320 GHz, with a reflection coefficient (return loss) of about -30 dB at resonance. The stubs were then first tuned for better impedance matching, and at 2.9500 GHz, the reflection coefficient was better than -45 dB. Next, the resonant length of the patch was reduced, by trimming both stubs while continuously measuring the input resistance. At 3.0000 GHz, both the tuning stubs were approximately half their original length of 5 mm. Figure 3 shows the reflection coefficient of the patch tuned to the specified frequency. Note that the reflection coefficient of -63 dB at 3.0000 GHz indicates that the patch is matched to the 50 Ω coaxial line to within 0.1 Ω.

By using an identical patch, it was found that the reflection coefficient of the patch could very easily be tuned to less than -45 dB, at any frequency over the range 2.95 to 3.05 GHz, i.e. a tuning range of 3.33%, without degradation of the match.

The tuning method was also successfully used on the two-patch sub-arrays shown in Figure 4. In these, a high-impedance microstrip transmission line joins the edge of the driven patch with the edge of the second patch. The trimming stubs are located on the two outside edges, and allowed the array of eight two-patch sub-

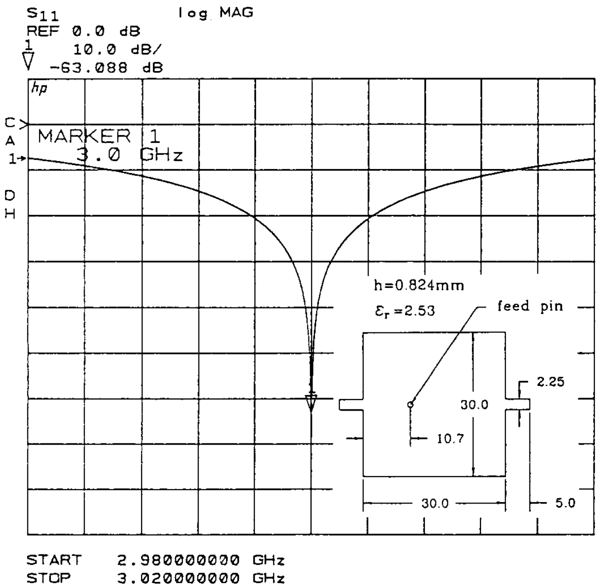

Figure 3. The measured reflection coefficient as a function of frequency for an experimental patch antenna.

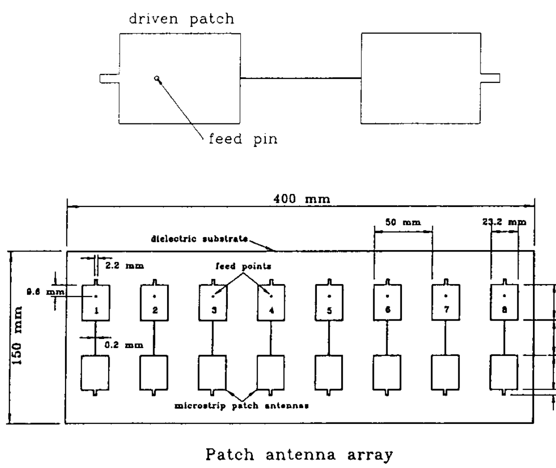

Figure 4. An array of eight double-patch sub-arrays, with each sub-array individually tunable.

arrays to be tuned to comply with precise reflection coefficient and resonant-frequency specifications. Results similar to those given for the single patch were achieved [7].

All the measurements described above were performed using an HP8510B network analyzer, in a temperature-controlled anechoic environment.

6. Cross-polarization due to the stubs

Pozar [1] argued that if a trimming stub is mounted on the patch center line, the resulting symmetry of the structure will tend to suppress the generation of cross-polarized spurious radiation. He also recommended that for negligible radiation-pattern perturbation, the stub lengths should be kept shorter than half the patch length.

This point was numerically investigated by van Tonder [8], using a moment-method code for a microstrip patch almost identical to Pozar's experimental patch [1, Figure 4]. The parameters were width = 25.44 mm, length = 18.50 mm, ε_r = 2.2, and substrate thickness d = 1.6 mm. The feed point was located on the center line, perpendicular to the wide edges, at x_0 = 5.78 mm (see Figure 1). The effect on the radiation pattern of center-line tuning stubs was studied, for a fixed stub width of 2.31 mm and different lengths of $n \times 2.31$ mm, with $n = 0,1,2,3,4$. Thus, the maximum stub length was 9.24 mm, or almost half the patch length. For this extreme case, the relative level of cross-polarized radiation at the peak of the beam (direction normal to the patch surface) had risen to a mere -50 dB. In the angular range between 40° and 80° from the peak of the beam, the E-plane cross-polarization level rose by a maximum of about 5 dB. These findings thus support Pozar's statement.

Cylindrical near-field measurements [7] of the array of Figure 4, mounted on a cylinder, also confirmed that the stubs had a minimal effect on both co-polarized and cross-polarized radiation patterns.

7. Conclusions

Microstrip patches can be designed, fabricated, and tuned to very strict frequency and reflection-coefficient specifications, in a few iterations, by the use of two stubs. In the example shown, a single patch was tuned to have a reflection coefficient of -63 dB, at a frequency which was within 50 kHz of the specified frequency of 3 GHz.

An array of eight two-element patches was also designed, built, and tuned to similar specifications.

A disadvantage of this method is that tuning is only possible from a lower to a higher frequency, due to the "destructive-trimming" technique. Another disadvantage is that it may not be practical to tune microstrip elements when a radome is to be used. Lastly, the technique can not compensate for variations in the resonant frequency due to temperature dependence of the substrate-dielectric constant.

8. Acknowledgments

The original version of this article appeared in the *1993 IEEE Antennas and Propagation Society International Symposium Digest, Volume 2*, pp. 964-967. It is reproduced here with minor changes.

Johann van Tonder is thanked for his moment-method analysis of the co-polarized and cross-polarized radiation patterns of a microstrip patch, with different tuning-stub lengths.

9. References

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More on Broadband Antenna Gain

The article by **Dave Pozar** on “Directivity of Omnidirectional Antennas,” in the October, 1993, issue, has brought comments from two of our readers.

1. Rajeev Bansal, of the University of Connecticut, Department of Electrical and Systems Engineering, wrote to say that a related article was published in the October, 1985 *Microwave Systems News and Communications Technology* (pp. 83-85). This article, by **K. M. Keen**, is entitled, “Method Permits Gain Estimation for Very Wide Beam Satellite Terminal Antennas.” It deals with antennas such as quadrifilar and conical-helix antennas, which have coverages in excess of hemispherical. Those antennas have a very broad front lobe, symmetrical about their axis, and a weak back lobe. Their coverage does not have a null on-axis, as do the antennas in Pozar’s article. Keen presents a graph similar to Pozar’s, relating directivity (or gain) to the angular coverage. For example, an antenna with 200-degree coverage (±100 degrees from the axis), and a -10 dB backlobe, has a gain of about 1.75 dBi, which includes an assumed dissipative loss of 0.25 dB.

An approximate expression for the relationship between the angular coverage ($\pm\theta_0$) and the directivity (D_0) is given by

$$\cos\theta_0 = \left[\frac{1+k-2/D_0}{1-k}\right],$$

where k is the backlobe level, relative to unity for the front lobe.

Many thanks to Professor Bonsal for calling our attention to this related article, and for his kind expressions of interest in our column.

2. Noel McDonald, of the Royal Melbourne Institute of Technology (Australia), Department of Communication and Electronic Engineering, wrote to say he published a communication, similar to Pozar’s article, in the March, 1978, issue of the *Transactions on Antennas and Propagation* (pp. 340-341). It was entitled, “Approximate Relationship Between Directivity and Beamwidth for Broadside Collinear Arrays.” It solves the same problem as Pozar did, but uses a technique which avoids the need for curve fitting. The resulting graph agrees with Pozar’s to within about 0.2 dB, for half-power (3 dB) beamwidths between 3 and 130 degrees. McDonald’s relationship between directivity (D) and half-power beamwidth (BW) is given by

$$D\,(\mathrm{dBi}) = 10\log\left[\frac{101}{\left(BW - 0.0027BW^2\right)}\right].$$

Many sincere thanks to Professor McDonald for his interest in our column, and for sending this interesting related information.

More on Radiation Resistance and Directivity of Circular Loop Antennas

The article by **John D. Mahony**, in the August, 1994, issue, brought the following comments by **A. David Wunsch** of the University of Massachusetts Lowell, Department of Electrical Engineering. We are most grateful to Dave Wunsch for these comments, and present them here, along with John Mahony’s reply.

The August, 1994, contribution to the “Antenna Designer’s Notebook,” by John D. Mahony, should carry with it a caveat that the results presented must be used with caution, or serious errors will result. Dr. Mahony has found a nifty way to compute the radiation resistance and directivity of small- and intermediate-size loop antennas. Not mentioned in the article is that these results are valid only to the extent that the current along the circumference of the loop is uniform in both magnitude and phase. For electrically small loops, the assumption is certainly valid. However, for the intermediate loops considered, it is generally not a valid assumption, unless the antenna has multiple feed points along its circumference, or has lumped impedances strategically placed at various locations in the wire to insure uniformity. Both arrangements are potentially awkward, and generally not used.

Several researchers have made contributions [1, 2, 3] that permit us to see the actual current distribution on a uniform-circular loop, fed at a single point. In reference [3], Figure 9.3.5(a), we find graphical results based on Storer’s integral-equation solution for the current on a loop antenna. With $ka = 0.1$ (a small loop–we use Mahony’s notation), the current found by Storer is very nearly uniform. However, for $ka = 0.4$, the current magnitude exhibits a variation the ratio of peak to minimum value of which is about 5:1. Figure 9.3.8 of the same reference shows some experimental