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MODELLING OF COAXIAL-FED MICROSTRIP PATCH ANTENNA BY FINITE DIFFERENCE TIME DOMAIN METHOD

Indexing terms: Antennas, Modelling, Microstrip

A direct three-dimensional finite-difference time-domain (FDTD) method is applied to coaxial-fed microstrip antennas. The model is shown to be an efficient and accurate tool for modelling coaxial-fed structures. The reflection coefficient can be determined from the simulated time-domain wave that is reflected down the coaxial line. Excellent agreement over a wide frequency range is shown in two cases between the measured and FDTD derived results.

The coaxial-fed microstrip antenna is a commonly used structure in integrated antenna arrays. The study of coaxial-fed structures is of interest for at least two reasons. The first is to determine the input impedance of the antenna for accurate matching in an array configuration, where the matching includes the effects of the connectors. From the practical point of view, the input impedance of wideband antennas needs to be known over a wide frequency range, as well as in the vicinity of the resonance frequency of the antenna. Secondly, this study can provide detailed information about the bandwidth, resonance frequency, and even higher mode behaviour of coaxial-fed structures.

To date, many numerical techniques have been developed to model this kind of microstrip patch antenna in the spectral domain. One is based on sophisticated attachment models,[1] in which the excitation current spreads over a charge cell. This model was developed to be compatible with the roof-top basis functions. Unfortunately, the matrix is severely ill-conditioned in the vicinity of the resonance frequency. Another popular model, the delta current source model, is based on the use of sinusoidal expansion modes[2] and the assumption that the current on the probe is constant. This technique only gives good results near the resonance frequency. Recently, a more accurate spectral domain model has been developed,[3] in which the fringing field is replaced by a frill of magnetic current and the TEM mode excitation concept is adopted. In this model, the dielectric discontinuity between the coaxial line and patch substrate still cannot be taken into account.

The finite difference time domain (FDTD) method has been widely used to solve electromagnetic problems since 1966. Because the Maxwell equations are discretised directly using central difference in both space and time, the FDTD method is more flexible in modelling complex structures, and can be used where spectral domain techniques fail. In the preceding decade, many investigators used the FDTD method for microstrip problems, but for the coaxial-fed patch antenna the analysis is based on assumptions that deviate from practice. As an example, the discontinuity between the coaxial line and patch region is replaced by an equivalent lump resistance.[4] Furthermore, the characteristic impedance of the coaxial line is not reflected in the model. Obviously, it is very difficult to give an accurate equivalent resistance to incorporate all the details of the discontinuity near the connector.

In this Letter, a judicious treatment of the coaxial line makes this model flexible in use and accurate in a wideband frequency region, in which all parasitic effects are included in the model, as well as fringing fields, dielectric discontinuities and characteristic impedance of the coaxial line.

As shown in Fig. 1, the inner conductor of the coaxial line is attached to the patch antenna going through the dielectric substrate and the outer conductor is connected to the ground plane. In this model, the antenna is divided into two regions

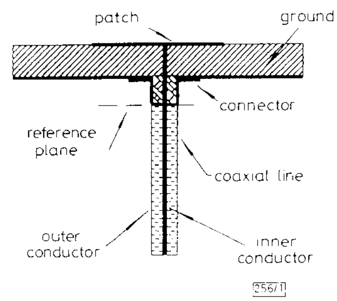

Fig. 1 *Coaxial-fed microstrip antenna*

for purpose of computation. One is the coaxial line region and the other is the patch antenna region. These two regions are carefully merged near the ground interface. The computational effort needed for the coaxial line region is less than 2% of that for the patch antenna region. Although, the boundary of the coaxial line is approximated by the staircasing, the amount of the scattered wave going into the coaxial line is strongly dependent on the characteristic impedance of the coaxial line, i.e. the electrical characteristics, but very weakly on the specific shape, i.e. the physical characteristics. It can be seen that a very good numerical result can be obtained as long as the numerical coaxial line has a 50Ω characteristic impedance (i.e. the same as that of the coaxial line used in the measurement).

The excitation pulse used in this study has been chosen to be Gaussian in shape. From the knowledge of the modes on the coaxial line, a simple field distribution can be specified at the excitation plane in such a way that the field components in the rectangular co-ordinate system take the projected value of the analytic radius field distribution. This is not the exact dominant TEM mode field distribution, although the non-TEM modes excited by this nonphysical excitation will have almost decayed after wave propagation of several lattices. The only mode which can then propagate down the coaxial line is the TEM mode.

The Mur first order absorbing boundary condition[5] was used on the upper fictitious open boundaries. It has been found that for microstrip antenna problems, little improvement can be made by using higher order absorbing boundary conditions.

After the incident wave passes through the reference plane, all the incident wave information is stored in a data array, and the program starts to record the reflected wave coming from the patch antenna. Actually, the Fourier transform is carried out while storing the time sequence information. The reflection coefficient is determined as

$$S_{11}(\omega) = V^r(\omega)/V^i(\omega)$$

where $V^r(\omega)$ and $V^i(\omega)$ is the Fourier transform of reflected voltage and incident voltage, respectively.

Two different shapes of numerically defined coaxial feed for a simple microstrip antenna are investigated: one is a staircase shape and another is a square shape. The characteristic impedance for these two coaxial-fed lines is about 50 Ω. The measurements were made with a Hewlett-Packard HP8510B microwave network analyser. The reference plane was located at the dielectric interface between the antenna connector and coaxial-fed line substrates.

Fig. 2 shows the measured and FDTD reflection coefficient results of two different numerical coaxial-fed shapes for a single patch antenna. It is obvious that both the magnitude

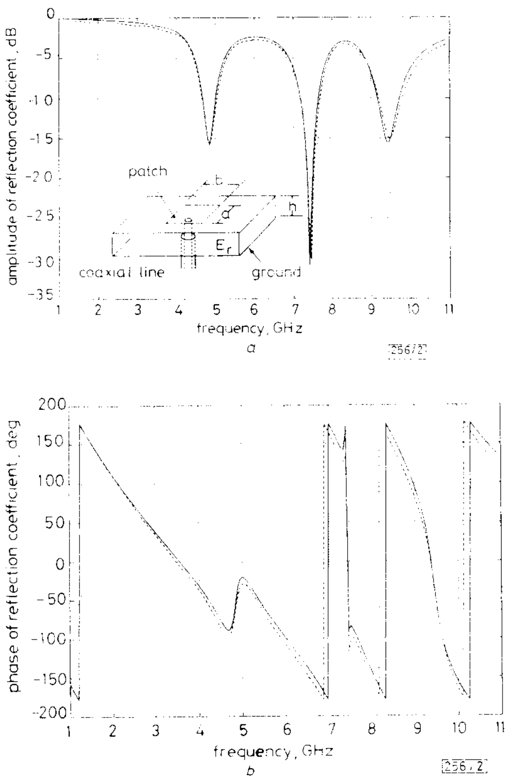

Fig. 2 *Reflection coefficient of single element patch antenna*

$a = 17{\cdot}72$ mm, $b = 25{\cdot}1$ mm, $\varepsilon_r = 2{\cdot}33$, $h = 2{\cdot}35$ mm and feed point at (12·5, 4·0) mm

········ measured results

——— FDTD results with staircase shape coaxial-fed

– – – – FDTD results with square shape coaxial-fed

a Amplitude against frequency

b Phase against frequency

and the phase are in excellent agreement in a wide frequency band including the characteristics of the higher modes.

In conclusion, an accurate FDTD coaxial-fed microstrip antenna model is proposed and is demonstrated and validated using both measured and numerical results. The accuracy is confirmed by comparing the numerical results with experimental results. This model can provide wideband frequency response using only one computation in the time domain. Full wave information is contained in the response. Based on the time domain solution, all the other parameters of interest can be derived, such as input impedance, current distribution on the patch and the radiation pattern. Hence, it is a very useful tool for designing sophisticated coaxial-fed printed antennas.

C. WU
K.-L. WU
Z. BI
J. LITVA

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Communications Research Laboratory
McMaster University
Ontario L8S 4K1, Canada

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1064 nm, 565 Mbit/s PSK TRANSMISSION EXPERIMENT WITH HOMODYNE RECEIVER USING SYNCHRONISATION BITS

Indexing terms: Optical communication, Optical receivers, Phase-locked loops

An optical 565 Mbit/s transmission system at 1064 nm with phase shift keying and homodyne detection using a new carrier recovery technique is presented. The phase error signal in the receiver is obtained by means of synchronisation bits. This method combines the advantages of the Costas loop with the simplicity of the pilot carrier technique.

Introduction: Phase control of the local oscillator in an optical PSK homodyne receiver is usually performed by means of the Costas loop technique.[1–3] In spite of higher complexity and the need of an optical hybrid it is preferred to the simple pilot carrier technique because AC coupling is possible and the PLL is independent of the data signal.[1,2,4,5] We present a phase locked loop design, which combines the advantages of the Costas loop with the simplicity of the pilot carrier receiver. The transmitter multiplexes special bits into the data stream which allows the receiver to obtain the phase error signal. An experimental setup using this carrier recovery method is demonstrated.

Principle of operation: Proceeding from the Costas loop design, the principle of the synchronisation method presented here is shown in Fig. 1. Instead of feeding a fraction α of the received light into the quadrature arm continuously as done in a Costas loop receiver, the inphase arm and quadrature arm are used alternately. It can easily be shown that, if the quadrature arm is used for a portion α of time, the properties of this receiver are the same as compared to a receiver with Costas loop with a power splitting ratio α. The phase error signal is obtained by sampling the quadrature signal and multiplying it with the polarity of the bit sent at this time.

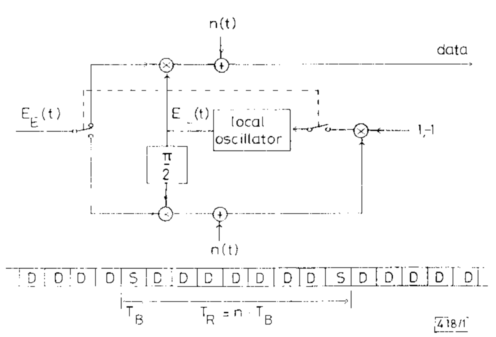

Fig. 1 *Principle of operation derived from Costas loop design*