an optimal conversion gain equal to +6 dB, very close to the experimentally obtained value of 5 dB.

5. CONCLUSION

The generation of microwave signal mixing can be realized by FM/IM conversion by using all-optical components such as interferometers. An FP, an RR, and a UMZ have been investigated. The mixing conversion loss which can be obtained is better than the one obtainable with classical solutions of optical-microwave mixing. The best conversion loss is the same with all types of interferometers, but better performances in terms of stability and frequency response are obtained when the transmission response of the interferometer is broadband over the microwave frequency range of interest. We have shown that the UMZ leads to the best compromise between good performance and practical realization. Experimental results obtained with a UMZ integrated on a glass substrate have proved to be in good agreement with simulation results, and have demonstrated the feasibility of this optical method for microwave mixing with a low-cost device of easy fabrication.

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INTERPOLATION TECHNIQUES FOR RAY-TRACING MODELS

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ABSTRACT: In numerous radio-communication applications, the knowledge of the electrical field along a given path is required. In urban environments, at high frequencies, the study of the field values is a time-consuming process. In order to accelerate this task, two interpolation methods for an approximate and fast evaluation of the field are presented. © 2000 John Wiley & Sons, Inc. Microwave Opt Technol Lett 25: 343–346, 2000.

Key words: high frequency; interpolation; module and phase interpolation; ray tracing

1. INTRODUCTION

At high frequencies, in an urban environment, where there are many wave scatters, the electromagnetic field of a radio wave has a relevant spatial variation, the so-called fast or Rayleigh fading. As a consequence, field-strength calculations must be performed at many sampling points along a specified route. At each of these points, it is necessary to determine the passing-by rays (that is the point visibility), as well as their path lengths, amplitudes, delays, and phases, in order to obtain the total field. This process implies a huge calculation load, especially when the reception area in which we want to predict the field is wide. To accelerate this problem, the sampling step is increased, so that the number of points where an exact ray tracing and computation must be performed is reduced. At the intermediate points, an interpolation is made for assessing the field values.

In this paper, for simplicity, besides the direct ray, only single reflected (R) and diffracted rays (D) are considered. Nevertheless, the proposed method can be extended to higher order interactions, i.e., RRR, RDR, RDD, etc., by a recursive application.

The GTD/UTD approach [1] is used to compute the field at high frequency. The reflected field at point P from the reflection point R is computed by [1]

$$E^{r}(P) = E^{i}(R)R^{r}\sqrt{\frac{\rho_{1}^{r}\rho_{2}^{r}}{(s + \rho_{1}^{r})(s + \rho_{2}^{r})}}e^{-jk}$$
(1)

where $E^r(P)$ is the reflected field at the observation point P, s is the distance between the points P and Q along the diffracted ray, $E^i(R)$ is the incident field at R, R^r is the dyadic of the Fresnel reflection coefficients, and $\rho_{1,2}^r$ are the radius of curvature of the wavefront of the reflected field.

The diffracted electric field E^d at the point of observation P is calculated by [1]

$$E^{d}(P) = E^{i}(Q)D\sqrt{\frac{\rho^{d}}{s(s+\rho^{d})}}e^{-jks}$$
 (2)

where $E^d(P)$ is the diffracted field at the observation point P, s is the distance between the points P and Q along the diffracted ray, $E^i(Q)$ is the incident field at Q, D is the dyadic of the GTD diffraction factor, and ρ^d is the radius of curvature of the wavefront of the diffracted field.

Two new interpolation methods, called MPV (module-phase visibility) and MP (module-phase) interpolation are considered hereafter. The MP procedure uses a ray-tracing technique [2] to determine the visibility and path length of each ray, whereas the MPV method uses the information about the homologous ray at the extreme points of the interpolation interval to ascertain the visibility. The term "homologous ray" denotes the ray diffracted on the same wedge or reflected on the same surface. In both methods, the field module and phase are interpolated individually for each ray.

2. METHODOLOGY

We assume that, in complex phase, the electromagnetic field is a vector magnitude with module and phase angle, the time variation being omitted. The longitudinal reception area is split into interpolation intervals of the same length. The module of the field of each ray at an interpolation point will be equal to the module of the field of its module at an endpoint of the interpolation interval. The phase will be the phase of the respective coefficient endpoint plus the ray path phase variation from this extreme to the interpolation point. A double interpolation (module and phase) must be performed, due to the variability of the diffraction and reflection coefficients at the GTD boundaries RSB (reflection shadow boundary) and ISB (incident shadow boundary).

3. MPV APPROXIMATION

In this method, a double interpolation is made affecting both the field and the visibility associated with each ray. To interpolate a diffracted ray, a check is made first, to see if the homologous ray reaches the two endpoints of the interpolation interval. This being true, we must then verify that there is a shadow boundary in the interpolation interval by calculating the module of the difference between the diffraction coefficient phases at both endpoints. In this case, it is necessary to know on which side of the boundary the interpolation point is located (Fig. 1). For diffraction points when there is no shadow boundary or the homologous ray does not exist at one extreme, and for reflected and direct rays, the interpolation interval is halved, and the points are interpolated from the nearest endpoint (Fig. 2).

We must emphasize that, in this method, the ray visibility is not analyzed, so there can be rays that can be obstructed at the interpolation point and free of obstructions and vice versa. In these cases, obviously, there will be an interpolation error.

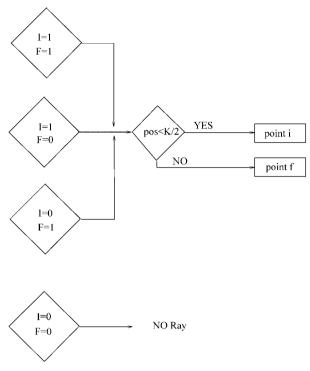


Figure 2 MPV interpolation for diffracted (where no SB is found), reflected, and direct rays. I=1 or F=1 denote the existence of a homologous ray at the initial or final interval endpoints i and f, respectively

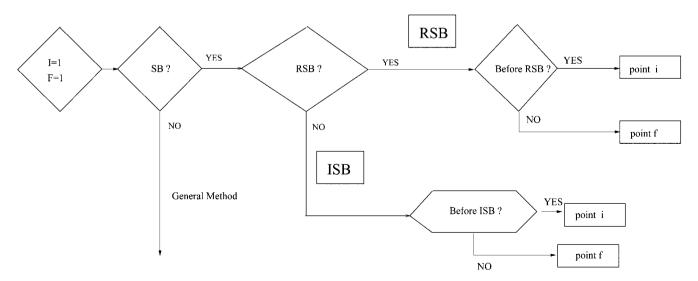


Figure 1 Diffraction interpolation method when an SB is found for MPV and MP approximations. I = 1 or F = 1 denotes the existence of homologous ray at the initial or final interval endpoints i and f, respectively

4. MP APPROXIMATION

In this method, a ray tracing is applied to determine both the visibility of each interpolation point and the distance from the source to that point.

First, for diffraction rays, we check if there is an RSB or ISB in the interpolation interval by the phase difference of the diffraction coefficient at the endpoints. This being the case, the location of the interpolation point with respect to the boundary is determined in the same way as that for the MPV method (Fig. 1). If there are no boundaries, the points are interpolated by the nearest extreme. This process applies when there are homologous rays at both extremes and at the interpolation point. If rays exist at only one extreme and at the interpolation point, this is used for the approximation, and if there are obstructions at both endpoints, the approximation is not made (Fig. 3). This causes an error because, in spite of these obstructions, a ray can exist at the interpolation point.

In this method, the calculation time of the diffraction and reflection coefficients is saved, but more time is devoted to the exact ray-tracing needed, so that, being more exact than the MPV, it takes more computing time.

5. RESULTS

The test area is shown in Figure 4, and corresponds to a typical microcellular environment, so that the transmitter and receiver antennas are placed well below the rooftops. The antennas used for the test were vertical dipoles. The transmitted power was 1 W, and the frequency was 1.8 GHz. Ground reflections were not considered.

The reception area is linear and bounded by P_a and P_b . To evaluate the proposed interpolation methods, first, a simulation with many points (every $\lambda/16$) was carried out. Then, new runs were executed using lower sampling rates, i.e., widening the calculation interval, so that some of the former

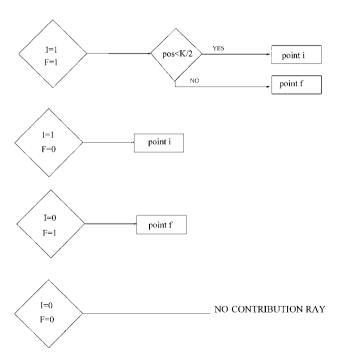


Figure 3 MP interpolation for diffracted (where no SB is found), reflected, and direct rays. I = 1 or F = 1 denote the existence of a homologous ray at the initial or final interval endpoints i and f, respectively

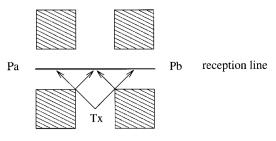


Figure 4 Test environment

calculation points were excluded. Afterwards, we estimated the field in these discarded points using both the MPV and MP methods, and comparisons were made between the original values and the interpolated ones. For assessing the accuracy of the interpolation, an error coefficient was defined, as given by the following expression (5):

$$e(k) = \frac{1}{N} \sum_{i=1}^{N} \frac{P_r(j) - P_{\text{int}}(k)}{P_r(j)}, \quad 1 \le j \le N$$
 (3)

where k is the number of interpolation points per segment, P_r is the exact received power, P_{int} is the interpolated power at the points excluded in the second simulation, and N is the number of points from P_a and P_b sampled with a step of $\lambda/16$.

The mean relative error is shown in Figure 5. The dotted and solid lines correspond to the MPV and MP methods, respectively. For sampling rates lower than a wavelength (above), the relative error is less than 0.5, and the two methods have the same performance. If the sampling rate exceeds a wavelength (below), the error increases with a maximum at 8λ because of the fast variability.

For the same example, the relative error in the case of the conventional "linear" interpolation, with no phase information taken into account, is greater than in the methods which use module and phase, as can be seen in Figure 6. Moreover, MPV performs better than MP in running time, so that it should be used for "nonfast" ray-tracing algorithms.

In the example considered here, the obstacles had been taken as perfect conductors. If they were dielectrics, the diffraction coefficient phase would have some discontinuities at different angles of shadow boundaries. Nevertheless, the relative errors for dielectric materials are similar to the ones for conductors.

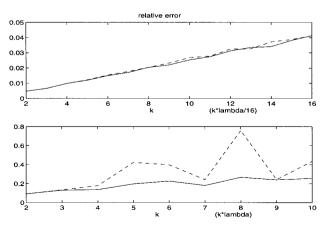


Figure 5 MPV and MP errors

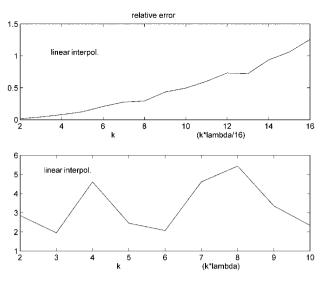


Figure 6 Linear interpolation errors

6. CONCLUSIONS

Module-phase interpolation techniques for ray-tracing models using GTD and Fresnel reflection coefficients have been presented. These interpolations show a better performance than the linear approximations due to the fact that the proposed methods take into account phase information and, for the diffraction rays, the shadow boundaries. These methods can be used for accelerating ray propagation simulations that require high computation times.

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SMALL-SIGNAL ANALYTICAL MOSFET MODEL FOR MICROWAVE FREQUENCY APPLICATIONS

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ABSTRACT: A small-signal analytical MOSFET model suitable for microwave frequency applications is presented. The effect of parasitic elements, the fringing-field effect, and distributed-gate inductance also have been incorporated. The generalized expression for I_d – V_d character-

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istics is given, and the admittance parameters extracted are expressed as a function of the MOSFET parameters, parasitic elements, frequency, and bias conditions. The theoretical predictions are compared with experimental/simulated data, and are in good agreement. © 2000 John Wiley & Sons, Inc. Microwave Opt Technol Lett 25: 346–352, 2000.

Key words: *MOSFET*; *transconductance*; *cutoff frequency*; *admittance parameters*

INTRODUCTION

The geometry and electric field dependence of scaled MOS-FETs have created a great interest in device modeling. A large variety of models have been reported in the literature [1–4] which provide an understanding of the device physics. As the device geometry shrinks, various effects, like the reverse short-channel effect, narrow-width effect, short-channel effect, and fringing-field effect, mix with the circuit complexity and degrade the device performance [5–7].

Several methods of evaluating the small-signal characteristics have been reported [8–10], and the results are applicable only to ideal three-terminal or two-port active devices. The analytical results derived on the basis of the two-port model cannot be applied to many practical cases of high-frequency MOST structures. The need to characterize these double-control devices as four-terminal or three-port active networks has been suggested for both junction gates and metal-oxide gates in the context of their low-frequency behavior. Das [11] reported the essential high-frequency characteristics, in terms of Y-parameters, including the control properties of the metal gate and semiconductor substrate terminals, but the distributed gate effect has been ignored up to 1 GHz. Allam and Manku [12] obtained Y-parameters using a transmission-line equation incorporating only the distributed gate resistance.

This paper presents the essential microwave characteristics of MOSFETs in the form of Y-parameters. The distributed-gate inductance and the fringing-field effect have been included in the present analysis. The fringing-field effect, which is a major constraint in device miniaturization, the narrow-width effect, and the short-channel effect also are suitably incorporated, and the results are compared with simulated/experimental data, and are in good agreement.

DEVICE MODELING

The drain current in a metal-oxide semiconductor field-effect transistor is evaluated using the expression

$$I_d = -\frac{\mu_{\text{eff}}W}{L} \int_{V_{mid}}^{V_{nis}} Q_m \, dV \tag{1}$$

where L and W are the channel length and width, respectively. $\mu_{\rm eff}$ is the effective mobility, and is given as

$$\mu_{\rm eff} = \frac{\mu_0}{1 + \theta(V_g - V_{\rm th} - \delta V_d) + \theta_b V_{sb} + \theta V_d}$$

where θ is called the mobility degradation coefficient, θ_b improves the current-voltage data fit, and is a constant, and $\theta_c = (L \cdot E_c)^{-1}$, where E_c is the critical field. δ is the drain-induced barrier-lowering parameter which affects short-channel devices operated near threshold. V_d , V_g , and V_{sb} are