# ANALYSIS OF PLANAR MILLIMETER WAVE SLOT ANTENNAS USING A SPECTRAL DOMAIN APPROACH

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**Abstract — Self radiating oscillators in the millimeter wave range can be built by combining a slot resonator with an impatt diode. Good performance can be achieved only, if the input impedance of the slot, seen at the terminals of the impatt diode, is in a proper range. In this paper the input impedance, obtained by a rigorous full wave analysis, is presented in dependence of the design parameters slot width and substrate height for silicon and PTFE. It will be shown that the required input resistance (typically 2 ...5 Ω) can be achieved by choosing narrow slots. The comparison with a recently published self radiating oscillator shows the validation of the simulation.**

## I. Introduction

Today, there is growing interest in monolithic integrated transmitters and receivers, where the active devices and the planar antennas are combined on a single chip [1]. In the last years, it has been shown that silicon and gallium arsenide are suitable substrates for monolithic integrated circuits as well as for planar antennas. By the combination of high permittivity and short wavelength, the requirement to occupy only little areas for the radiating element is fulfilled. A very interesting type of a radiating oscillator is a slot antenna fed by a negative impedance amplifier. A radiating oscillator in the X-band using a FET has been reported in [2]. In the millimeter wave region an impatt diode can be used [3]. The layout of the latter one is schematically depicted in Fig. 1. The slot acts simultanously as resonator and as antenna. For optimum performance of the radiating oscillator the radiated power should be as high as possible. This results in a high load of the impatt diode. That can be compensated by increasing the bias current, but due to thermal problems only within a limited range. Therefore, to find the optimum compromise between radiated power and loading of the impatt diode, the precise knowledge of the slot impedance versus various design parameter is necessary.

In this paper we present calculations of the slot impedance versus substrate height $d$ and slot width $w$. The results show good agreement with experimental data of radiating slotline oscillators at 109 GHz [3].

## II. Method of Calculation

A rigorous full wave analysis has been used taking into account radiation losses as well as losses due to surface waves. We have chosen a spectral domain approach, which is based on a Green's function formalism for the magnetic fields. These Green's functions are obtained by a transmission line analogy [4] based on impressed voltage sources as shown in Fig. 2. This results in

$$\begin{bmatrix} \tilde{H}_x \\ \tilde{H}_y \end{bmatrix} = \begin{bmatrix} \tilde{G}^H_{xx} & \tilde{G}^H_{xy} \\ \tilde{G}^H_{yx} & \tilde{G}^H_{yy} \end{bmatrix} \cdot \begin{bmatrix} \tilde{M}_x \\ \tilde{M}_y \end{bmatrix} \quad , \tag{1}$$

where $\tilde{H}$ is the magnetic field and $\tilde{M}$ the magnetic current density in the slot. We use the tilde as a reminder that the spectral representation is used. Since the slot width is far below a wavelength ($w \ll \lambda_0$), the transverse components are neglected. This simplifies (1) to

$$\tilde{H}_x = \tilde{G}^H_{xx} \cdot \tilde{M}_x \quad . \tag{2}$$

The Green's function is given by

$$\tilde{G}^H_{xx} = \frac{k_x^2}{k_t^2}\tilde{G}^h + \frac{k_y^2}{k_t^2}\tilde{G}^e \quad , \tag{3}$$

where

$$k_t = \sqrt{k_x^2 + k_y^2} \quad , \tag{4}$$

$$\tilde{G}^{e,h} = \frac{1}{Z_A^{e,h}} + \frac{1}{Z_B^{e,h}} \quad , \tag{5}$$

with

$$Z_A^{e,h} = Z_0^{e,h} \quad , \tag{6}$$

$$Z_B^{e,h} = Z_1^{e,h} \cdot \frac{Z_0^{e,h} + j \cdot Z_1^{e,h} \tan(k_{z1}d)}{Z_1^{e,h} + j \cdot Z_0^{e,h} \tan(k_{z1}d)} \quad . \tag{7}$$

Applying the Galerkin's method with piecewise sinusoidal basis functions yields a linear equation system



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$$\begin{bmatrix} J_1 \\ \vdots \\ J_N \end{bmatrix} = \begin{bmatrix} \int\int B_1 \tilde{G}^{xx} B_1 \, dk_x dk_y & \cdots & \int\int B_1 \tilde{G}^{xx} B_N \, dk_x dk_y \\ \vdots & \ddots & \vdots \\ \int\int B_N \tilde{G}^{xx} B_1 \, dk_x dk_y & \cdots & \int\int B_N \tilde{G}^{xx} B_N \, dk_x dk_y \end{bmatrix} \cdot \begin{bmatrix} A_1 \\ \vdots \\ A_N \end{bmatrix} , \quad (8)$$

where the excitation vector $\vec{J} = \vec{n} \times \vec{H}$ represents the impressed currents. $B_i$ are the basis functions and $A_i$ the corresponding weighting coefficients. In order to minimize the numerical effort we implemented the recently reported method of integrating along the axes of the Cartesian instead of the polar coordinate system [5]. Furthermore a path of integration in the complex plane was chosen to circumvent the tedious task of extracting surface wave poles in the case of lossless structures. These poles result from the zeros of (7). The impatt diode was modelled by an impressed line current located at the center of the slot. Thus, the excitation vector contains one non zero element

$$\vec{J} = [0, \cdots, 0, 1, 0, \cdots, 0]^T \quad (9)$$

Since the excitation current is normalized to unity, the input impedance $Z_i$ is given by the coefficient of the center basis function.

## III. Results

For two different substrate materials (silicon: $\epsilon_r = 11.6$, PTFE: $\epsilon_r = 2.55$) the input impedance versus various parameters has been calculated. Fig. 3 depicts the input impedance $Z_i$ versus $L/\lambda_0$, where $\lambda_0$ represents the free space wavelength and $L$ the length of the slot. The slot width was $w = L/14$ and the substrate thickness was $d = L/11.2$. As can be seen the half-wave and full-wave resonance represents a parallel-type and series-type resonance, respectively. Since the input resistance at the parallel-type resonance is very high, a full-wave slot has to be used for impatt diode oscillators. The resistance of the resonator seen by the impatt diode terminals should be typically in the range of $2 \ldots 5\,\Omega$.

In the next two figures the behaviour of the input resistance at the series-type resonance is shown versus the substrate height $d/\lambda_0$ (Fig. 4) and the slot width $w/\lambda_0$ (Fig. 5), respectively. A minimum of the resistance is obtained for $d/\lambda_0 = 0.037$ for silicon and $d/\lambda_0 = 0.07$ for PTFE. That means, there exists an optimum substrate thickness where the quality factor of the resonator becomes maximum. Since there is only a small variation of the resistance around the minimum, the accuracy of the substrate thickness is not very critical. This is important for reproducible design of slot antennas.

As can be seen in Fig. 5, decreasing the width of the slot yields an almost linear decrease of the resonance resistance. Thus, a desired resonance resistance can easily be adjusted by choosing the proper slot width.

To validate our results we have calculated the resonance resistance of the slotline of the impatt oscillator published in [3]. For the frequency, where the reactance of the diode is matched, an input resistance of $6.1\,\Omega$ was obtained. For the given bias current of 300 mA the resistance of the impatt diode is somewhat below $-6\,\Omega$. Thus, the oscillation starts only for bias currents larger 300 mA as reported in [3].

## IV. Conclusion

The characteristics of a slot antenna on a substrate with finite thickness has been calculated using a full wave analysis. The feasibilty of CW operation of a radiating oscillator consisting of slot antenna and impatt diode has been shown. The variation of the slot width allows an optimal matching of the resonator to the diode for any desired bias current.

Planar slot antennas are very promising devices for millimeter wave applications, because they provide larger bandwidth than microstrip patches and dipoles. The presented results give necessary design parameters for planar slot antennas.


## V. Acknowledgement

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## References


[1] Russer, P., "Silicon Monolithic Millimeterwave Integrated Circuits," *Proc. 21th Microwave Conf.*, pp. 55–71, Stuttgart, Germany, 1991.

[2] Kawasaki, S., Itoh, T. "A Layered Negative Resistance Amplifier and Oscillator Using a FET and a Slot Antenna," *1991 IEEE MTT-S Int. Microwave Symp. Digest* vol. 3, pp. 1261–1264, Boston, MA., June 1991.

[3] Büchler J., Strohm K.M., Luy J.F., Göller T., Sattler S., Russer P., "Coplanar Monolithic Silicon IMPATT Transmitter," *Proc. 21th Microwave Conf.*, pp. 352–357, Stuttgart, Germany, 1991.

[4] Itoh, T., "Spectral Domain Immitance Approach for Dispersion Characteristics of Generalized Printed Transmission Lines," *IEEE Trans. Microwave Theory Tech.*, vol. MTT–28, no. 7, pp. 733–736, 1980.

[5] Yang H.Y., Nakatani A., Castaneda J.A., "Efficient Evaluation of Spectral Integrals in the Moment Method Solution of Microstrip Antennas and Circuits," *IEEE Trans. Antennas Propagat.*, vol. 38, No. 7, pp. 1127–1129, 1990.

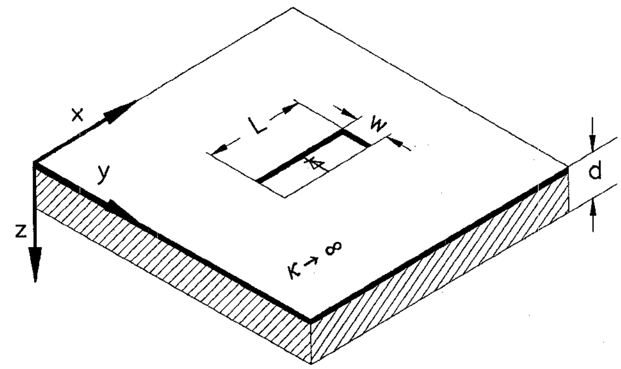
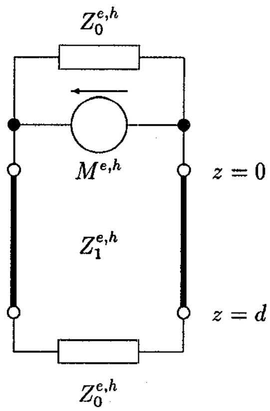



Fig. 1. Schematic structure of radiating oscillator.




$$Z_i^e = \frac{k_{zi}}{\omega \epsilon_i}$$

$$Z_i^h = \frac{\omega \mu_i}{k_{zi}}$$

$$k_{zi} = \sqrt{k_i^2 - k_x^2 - k_y^2}$$

$$i = 0, 1$$

Fig. 2. Equivalent transmission line for the slot antenna structure.

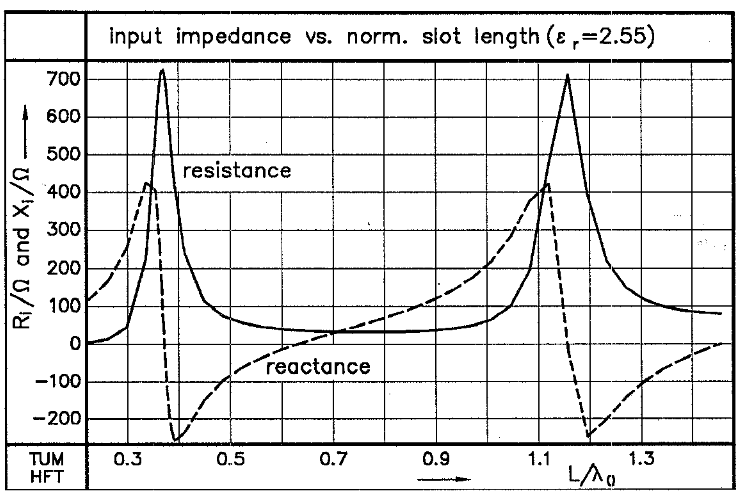



Fig. 3a. Slot impedance $Z_i = R_i + jX_i$ for PTFE ($w = L/14$, $d = L/11.2$).

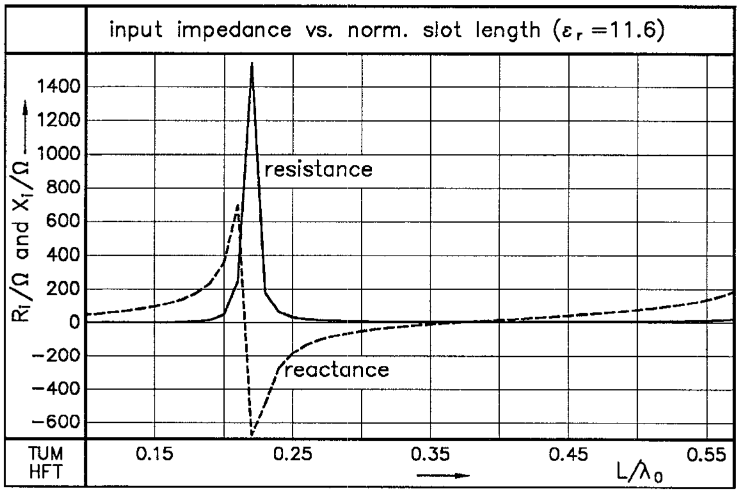



Fig. 3b. Slot impedance $Z_i = R_i + jX_i$ for silicon ($w = L/14$, $d = L/11.2$).

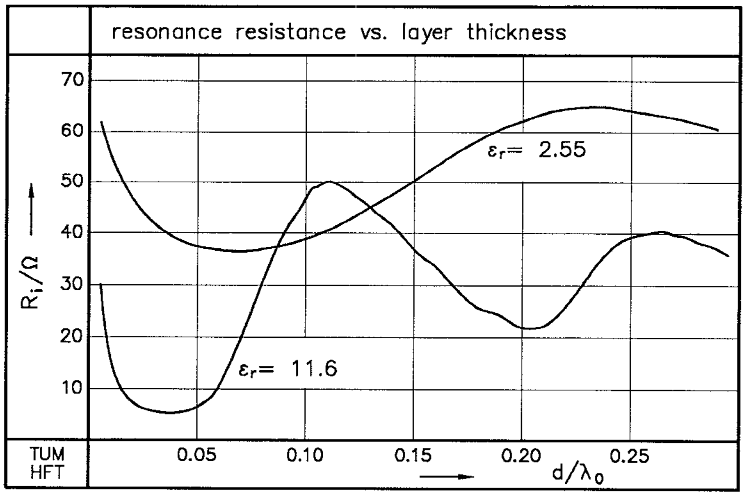



Fig. 4. Resonance resistance $R_i$ for PTFE and silicon ($w = L/14$).

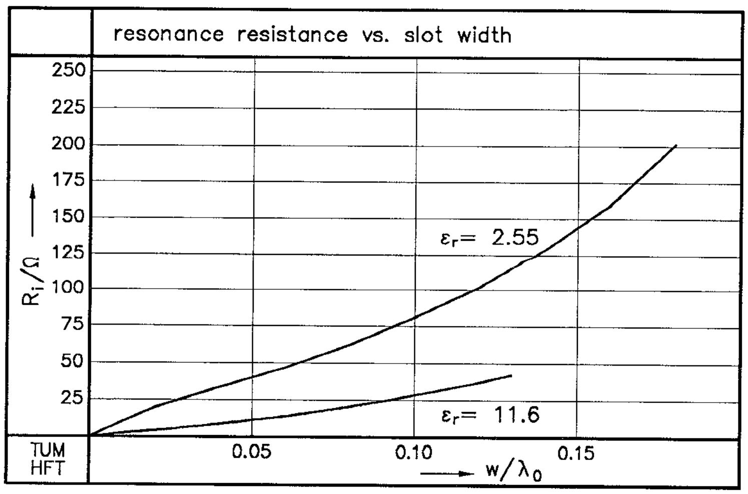



Fig. 5. Resonance resistance $R_i$ for PTFE and silicon ($d = L/11.2$).