OPTIMIZATION OF A DIODE-LOADED DIPOLE ANTENNA

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Abstract

A circuit model for a diode-loaded dipole electric field measurement probe is presented which is valid for frequencies below 10 GHz. Assumptions are made to allow linear analysis based on the Laplace Transform method. Optimized design criteria for the probe are then developed with the major considerations being large bandwidth, independence of temperature, and maximum output. Experimental measurements on a probe designed for operation between 1 and 1000 MHz and comparison with the theoretical results are presented.

I. Introduction

A dipole antenna loaded at its terminals with a microwave diode has been utilized as a probe to measure the electric field. If the dipole is short, the probe is usable over three decades of frequency and will not perturb the field. Use of semiconductor lines to extract the detected dc voltage further increases the nonperturbing aspect of the probe. Calibration yields information necessary to use the system to measure unknown fields. However, optimization cannot be achieved without reasonable detailed models of the dipole, diode, and extraction system. The resulting nonlinear differential equation can be solved numerically but it is difficult to develop scaling relationships from numerical results. Therefore, assumptions are made to allow linear analysis based on the Laplace Transform method. This analysis is utilized to develop scaling relationships which are then used to obtain optimized design criteria. The major considerations are the design of a probe for large bandwidth, independence of temperature, and maximum output.

The equivalent circuit for the dipole with its load impedance consisting of a diode, low-pass filter, carbon line resistance and voltmeter is shown in Figure 1. The low-pass filter, comprised of $R_{dc}$ and $C_{dc}$, is recommended to reduce the interference associated with removing the dc output voltage from the EM field. The capacitance, $C_{dc}$, is assumed to be large enough so that it can be considered a short-circuit for AC signals in the frequency range of interest. Therefore, the dc output of the low-pass filter, $V_{dc}$, is given by the time average of the voltage across the diode, $V_d$.

It is known that the diode capacitance, $C_d$, is larger when the diode is forward-biased ($V_d > 0$) as opposed to reverse-biased as shown in Figure 2. Also, since the diode resistance, $r_d$, is much smaller when the diode is forward-biased, the charge current available for $C_d$ will be reduced. Therefore, $C_d$ will charge slower when $V_d$ is positive. On the other hand, when the diode is reverse-biased ($V_d < 0$), $C_d$ will charge up quicker because of the reduced value of $C_d$ and the increased value of $r_d$. For these reasons, the maximum positive voltage will always be much less than the maximum negative voltage as shown in Figure 3a. As the frequency increases so will the difference between the positive and negative voltages. If the frequency is high enough, it can be assumed that the diode remains reverse-biased 100% of the time, as shown in Figure 3b. This approximation is valid as long as the diode forward-bias rise time is at least five times the period of the excitation signal, i.e.

$$f = 5/(3 \frac{R_{fb} C_d}{R_{fb} + R_{dc}})$$

where $R_{fb} = \frac{R_{fb} R_{dc}}{R_{fb} + R_{dc}}$ with $r_{db}$ equal to the average forward-bias diode resistance. Assuming $R_{fb} = 50 \text{ k\Omega}$ and $C_d = 0.5 \text{ pF}$, it is estimated that the frequency must be greater than 70 MHz for the 100 % reverse-bias approximation to be valid.

The above discussion indicates that at higher frequencies it is appropriate to assume that the diode is always reverse-biased. Now, if the voltage across the diode can be determined, then the dc output voltage can also be determined, i.e.

$$V_{dc} = V_d = V_{d-p-p}/2 = (2)^{1/2} V_d \text{ rms}$$

II. Dipole and Diode Circuit Models

The simplest model for the dipole is a series circuit consisting of a voltage source, resistance, inductance, and capacitance as shown in Figure 1. The element $V_{oc}$ is the induced open-circuit voltage and is given by:

$$V_{oc} = E_{0} h_{e} \sin \theta$$
where $E_0$ is the electric field magnitude, $\theta$ is the angle of incidence of the field, and $h_e$ is the dipole effective height given by\(^\text{1,5}\)

$$h_e = h \left[ \frac{2}{\tan \frac{\Omega - 1}{2}} \right]$$

with $k$ the wavenumber and $\Omega$ the commonly used dipole shape factor parameter (i.e. $2\ln(2h/a)$ where $h$ is the dipole half height and $a$ is the dipole radius).

The following equation, accurate for $kh<2.0$, can be obtained for the dipole resistance\(^\text{2}\)

$$R_a(kh) = 19.31 (kh)^2 [1 + 0.2236 (kh)^2]$$

The dipole impedance can be well approximated by a single capacitance when $kh<<1$ given by\(^\text{2}\)

$$C_a = h/(K_o c)$$

where $c$ is the speed of light and $K_o$ is the dipole characteristic impedance equal to $60(\Omega - 3.386)$. It can be shown that this equation is valid up to and slightly past resonance\(^\text{2}\).

The equivalent lumped-circuit inductance at the unloaded dipole resonance is

$$L_a = h K_f / [c (k_r h)^2]$$

where $k_r$ is the unloaded dipole resonant wavenumber given by\(^\text{2}\):

$$k_r = \arctan \left[ \frac{K_o}{K_1} (kh=k_r h/a) \right]$$

with

$$K_1(k_r h/a)=42.54 - 19.26 h/527.4$$

It can be shown that (7) is also valid below resonance\(^\text{2}\) when $\Omega>>\omega$.

The simplest model for the diode is merely a resistance and a dc voltage source in parallel with a capacitance as shown in Figure 1. More complex models can be conceived but will add no effect until the frequency exceeds approximately 10 GHz\(^\text{2}\).

The low injection, low frequency, ideal diode $V$–$I$ relationship is given by\(^\text{4}\)

$$i = I_s \left[ \exp(V_t/V_s) - 1 \right]$$

where $I_s$ equals the reverse saturation current, $V_t$ the Boltzmann's equivalent voltage ($kT/q$), $q$ the magnitude of the electron charge, $k$ Boltzmann's constant, and $T$ the diode junction temperature. Taking the inverse of the derivative of (8) yields the diode resistance, $r_d$. Assuming the diode is always biased off, $r_d$ has a minimum resistance value given by $V_t/I_s$ which is typically greater than 500 k$\Omega$ for most diodes. The diode resistance can be therefore assumed to be constant and even neglected if the filter resistance, $R_{dc}$, is small enough. Of course this will not be valid for large reverse voltages where reverse breakdown begins to have an affect.

The total diode capacitance, $C_d$, is the parallel combination of the case, transition, and injection capacitances\(^\text{5}\). The case capacitance is a constant independent of frequency or applied voltage. The transition capacitance exists whether a bias is applied or not. The injection capacitance exists only when the diode is forward-biased and will therefore be neglected here. The total diode capacitance, $C_d$, can therefore be approximated by using the Schottky equation for the transition capacitance\(^\text{4}\) plus a constant for the case capacitance, i.e.

$$C_d = K_f/(V_x - V_d)^m + C_c$$

where $m$ is a constant between $1/3$ and $1/2$ depending on the junction gradient, and $K_f$ is a constant dependent on the dielectric constant, junction area, and the impurity doping levels of the P and N-type regions. The offset voltage, $V_x$, is between 0.5 and 0.9 volts for typical temperatures and doping levels. Equation (9) is shown in Figure 2 for $m=1/2$, $V_x=0.75$ volts ($T=250K$), $V_x=0.9$ volts ($T=300K$), $K_f=0.5(10^{-12})$, and $C_c=0.2pF$. If the diode is always reverse-biased then $C_d$ is shown to be essentially a constant independent of voltage or temperature. This independence of temperature is a major reason for operating at frequencies high enough to ensure reverse-bias operation.

### III. Spectral Response

Some of the consequences of assuming the 100% reverse bias condition is that $C_d$ and $r_d$ can be considered constants.

The circuit can now be simplified to a linear AC circuit with a constant dc bias given by $V_b=0.5V_d$ per (2). The transfer function, $G(s)$, can now be determined from Laplace Transform analysis and is given by

$$G(s) = \frac{s C_a R_{rb}}{[1+sC_a R_{rb}] [1+sC_a R_{ra} + s^2C_a L_a + \frac{s C_a R_{rb}}{1+sC_a R_{rb}}]}$$

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where \( s \) is the Laplace Transform operator, \((\sigma+j\omega)\), and 
\[ R_{rb} = \frac{r_{db} R_{db}}{(r_{db} + R_{db})}, \]
with \( r_{db} \) equal to the diode reverse-bias resistance. The magnitude of the diode voltage, \( V_d(f) \) can now be calculated by 
\[ V_d(f) = G(s=j\omega)V_o(f). \]
The dipole open circuit voltage, \( V_{dc} \) can be related back to the incident field by (3) and the dc output voltage determined using (2) to be
\[ V_{dc} = (2)^{1/2} G(j\omega) E_0 h c \sin\theta \quad (11) \]

It is necessary to simplify (10) to formulate scaling relationships and develop optimized design criteria. This can be accomplished by assuming that the second-order poles are much larger than \((1+sC_dR_{rb})\), i.e., \( s^2C_aL_a<<1 \) and \( sC_aR_a<<1 \), yielding
\[ G(s) = \frac{s C_a R_{rb}}{1 + s R_{rb}(C_a + C_d)} \quad (12a) \]
This is a high-pass filter transfer function form with break frequency given by
\[ f_{3dB} = 1/(2 \pi R_{rb}(C_a + C_d)) \quad (12b) \]
and a high-pass response equal to
\[ G(j\omega) = \frac{C_a}{C_a + C_d} \quad (12c) \]
as shown in Figure 4a. The capacitive divider effect given by (12c) has also been demonstrated by other researchers using a much more complicated technique involving simplifying a nonlinear differential equation solution given in terms of Airy functions of imaginary order and imaginary argument3.

For frequencies much greater than (12b), (10) can be rewritten as a second-order transfer function, namely:
\[ G(s) = \frac{C_a}{C_a + C_d} \left[ 1 + \frac{2\delta}{\omega_n} + \frac{s^2}{\omega_n^2} \right]^{-1} \quad (13a) \]
where the undamped natural frequency, \( \omega_n \), is given by
\[ \omega_n = \left[ C_a + C_d \right]^{1/2} \frac{1}{\left[ C_a \right]^{1/2}} \quad (13b) \]
and the damping factor, \( \delta \), is
\[ \delta = R_d/(2L_a \omega_n) \quad (13c) \]

The steady state transfer function magnitude for high frequencies, \( G(j\omega) \), is shown in Figure 4b. Note that \( f_p = f_{3dB} \) as given by (13b). Combining Figures 4a and 4b gives the overall dipole/diode frequency response.

**IV. Probe Design and Optimization**

Optimum operation is obtained when the probe is used in its frequency-independent region which occurs between a factor of 4 and 25 below the unloaded system resonance frequency2. Making \( h \) as large as possible within this constraint will lead to a probe with higher output voltage. Other benefits of a large \( h \) will become apparent below. The optimum dipole height is therefore given by
\[ h = c/(16 f_{max}) \quad (14) \]

Figure 5 shows the probe loaded resonant frequency as approximated by (13b) and normalized by the thin dipole unloaded resonant frequency. As \( C_d/h \) is decreased this frequency increases. Also, typical values of \( C_d/h \) for probes operating around 100 MHz are about 10 pF/m. Therefore, the resonant frequency is also increased as \( \Omega \) is decreased (fatter dipole). A larger antenna capacitance, \( C_a \) caused by a decreased \( \Omega \) will also reduce the frequency where the output voltage starts dropping at a –6 dB/octave rate as indicated by (12b) or Figure 4a. A smaller \( C_d \) will increase this frequency but this effect is minimal, since \( C_a \) dominates. It is therefore concluded that reducing \( C_d \), increasing \( h \), and decreasing \( \Omega \) will result in a larger frequency-independent region.

Figure 6 shows that the effect of diode capacitance variation due to either temperature or voltage on the probe output voltage. This effect can be minimized by decreasing \( C_d \) or \( \Omega \) and increasing \( h \). This also reduces the effects of differences between one diode and another which will make it easier to match multiple probes.

In some cases it may be desirable to maximize the probe output. The normalized probe output voltage as calculated from (11) is shown in Figure 7. For a given \( C_d/h \) there exists an optimum \( \Omega \) that will maximize the probe output given by
\[ \Omega_{opt} = 1 + [21.475/(C_d/h) - 0.9218]^{1/2} \quad (15) \]
Typical values of \( C_d/h \) may lead to a \( \Omega_{opt} \) less than 4. However, the dipole model used here is not valid below a shape factor of 4. Figure 7 also illustrates that a larger probe output will be obtained by reducing \( C_d \) and increasing \( h \).

Figure 4a and (12b) shows that the break frequency is proportional to the inverse of the load resistance, which dictates that the diode resistance, \( r_d \) should be made as large.
as possible. An additional benefit of making $r_d$ large is to reduce temperature effects on probe output. If $r_d$ is large with respect to $R_{dc}$, then $R_{dc}$ will dominate and the temperature variations of $r_d$ will not affect the probe output. This involves a tradeoff, since reducing $R_{dc}$ may also decrease the size of the frequency independent region. To achieve a large $r_d$ requires choosing a diode with a very small saturation current, $I_s$, and a large reverse breakdown voltage, $V_{lb}$.

V. Summary

Experimental measurements were obtained on a dipole of height 0.0826 meters and radius 0.001195 meters loaded with a 1N832A microwave diode. Figure 8 shows the comparison between the experimental measurements and the analytical results obtained from the transfer function technique. Areas of disagreement include a compression effect at high field strengths and a $-2 \text{ dB/ octave}$ slope region below the frequency-independent region which was not predicted by the theoretical approach. The compression effect is most obvious between 100 MHz and 500 MHz and begins at approximately 80 V/m. This is caused by the small resistance of the 1N832A diode at large reverse-bias voltages which is due to the small reverse breakdown voltage of this diode. The approximation of constant or negligible diode resistance is therefore invalid at high field strengths. The $-2\text{ dB/ octave}$ slope below 100 MHz occurs because the period of oscillation has become long enough to allow the diode capacitance to charge positive. This indicates a forward-bias voltage for part of each cycle. The approximation that the diode is reverse-biased 100% of the time therefore becomes invalid below 100 MHz.

Even with these discrepancies, the transfer function technique does provide a good quantitative fit to the measured data. Therefore, the scaling relationships developed in this paper are valid and can be utilized to develop the optimized design criteria listed below:

a. Choose a diode with a small total capacitance, a small reverse saturation current, and a reverse breakdown voltage in excess of 10 volts. The latter two items will lead to an increased diode resistance.

b. Choose the dipole half-height as given by (14).

c. Determine the optimum dipole shape factor, given by (15), for maximum output voltage. If maximum output voltage is not a concern, then choose $\Omega = 4$ as the optimum value. Note that the two values may be equal depending on the diode used.

d. Use teflon or polystyrene for the support structure, limit the semiconductor line length to less than 20 meter length, and utilize a high impedance voltmeter.

REFERENCES


Figure 3: Output voltage of the dipole/diode system

Figure 4: Dipole/diode transfer functions

Figure 5: Probe resonant frequency

Figure 6: Effect of diode capacitance variations on output voltage

Figure 7: Dipole/diode normalized output voltage

Figure 8: Comparison of measured data and the transfer function technique