Radiation and Conductivity Constraints on the Design of a Dipole Electric Field Sensor

I. Introduction

In designing sensors for close-in nuclear EMP measurements, perhaps the most difficult problems are associated with the nuclear radiation and the conductivity of the air. These are particularly acute in the case of the electric field probe in air. The problem of measuring intense pulsed electric fields in the absence of this peculiar environment is, comparatively speaking, an easy problem. The nuclear radiation displaces charge into and out of the probe and the effects associated with the air conductivity alter the response of the probe to the electric field. It is as important in sensor design to understand these phenomena and their constraints on electric field probe design as it is to understand such things as the effective height and capacitance of the probe.

In this note the nuclear radiation and air conductivity constraints on the design of an electric dipole will be discussed. (The electric field probe utilizing a single element with a ground plane and often called a "monopole" is, strictly speaking, also a dipole.) The same effects must also be considered (although in different forms) in the design of the more "exotic" (and more complex) electric field probes, such as those using the Stark or electro-optic effects. However, all references in this note will be to the electric field dipole. Units for all quantities will be rationalized m.k.s., unless otherwise stated.

II. Time-Varying Air Conductivity

First consider the time-varying air conductivity. During the radiation pulse the conductivity, σ, can be approximately related to the gamma flux, γ, at any given point by

$$\sigma = en\mu(E) = e\,\mu(E) \times \frac{2.1 \times 10^{15}}{\nu}\,\gamma\ \frac{\text{mhos}}{\text{meter}} \tag{1}$$

where e is the electronic charge, ν (about 10^8 sec^{-1}) is the attachment frequency of electrons to neutral oxygen molecules, $\mu(E)$ is the electron mobility as a function of the electric field strength (typically about 1.0 meter^2/volt-sec for electric fields below 10^5 volts/meter), n is the electron density in electrons/meter^3, and γ is the γ-ray flux expressed in roentgens/sec. Thus, during the radiation pulse (assuming small electric fields)

$$\sigma = 3.4 \times 10^{-12}\,\gamma\ \frac{\text{mhos}}{\text{meter}} \tag{2}$$

This conductivity will give a conductance, $G_g(t,E)$, to the electric field dipole as shown in the equivalent circuit in figure 1. As indicated by the variable, t, this conductance will be a function of time. Without this time changing conductance the signal output, V, could be related to the input signal, Eh (where h is the effective height of the dipole), for

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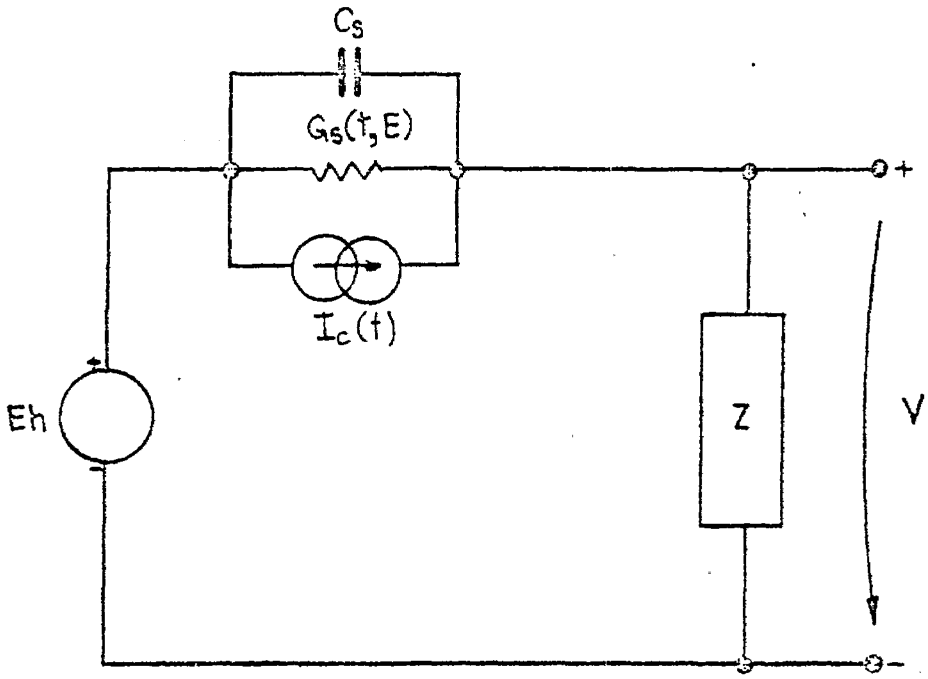


FIG. 1. EQUIVALENT CIRCUIT OF ELECTRIC FIELD DIPOLE.

any Z by either Fourier or Laplace transform techniques. One might even try to continuously measure $G_s(t,E)$ and unfold the output by solving a differential equation. However, the easiest way to get around this problem is to make Z as large an impedance as possible (around 10^8 ohms), for example by use of a cathode follower. Then, if the signal cable capacitance is small compared to the sensor capacitance, C_s, (as discussed in SSN VI) and if the load conductance is small compared to the sensor conductance

$$V = Eh \tag{3}$$

for all frequencies. Thus for the assumed conditions the changing conductance $G_s(t,E)$ has negligible effect on the dipole response.

This sensor conductance can be estimated by noting that if non-linearities and boundary layer problems (to be discussed in succeeding sections) in the conductivity are neglected for the moment, then the dipole conductance can be calculated by replacing the permittivity in the sensor capacitance calculation by the conductivity as in equation (2). This is most easily seen in the case of a parallel plate dipole as in figure 2. Where

$$C_s = \varepsilon_o \frac{A}{d} \tag{4}$$

and

$$G_s(t,E) \simeq \sigma \frac{A}{d} \tag{5}$$

where A is the plate area (one side) and d is the plate spacing. If we assume

$$A = 1 \text{ meter}^2 \qquad d = .1 \text{ meter} \tag{6}$$

then

$$C_s = 88.5 \text{ pf} \tag{7}$$

and

$$G_s(t,E) \simeq 3.4 \times 10^{-11} \gamma \text{ mhos} \tag{8}$$

Thus, for an assumed resistive Z of

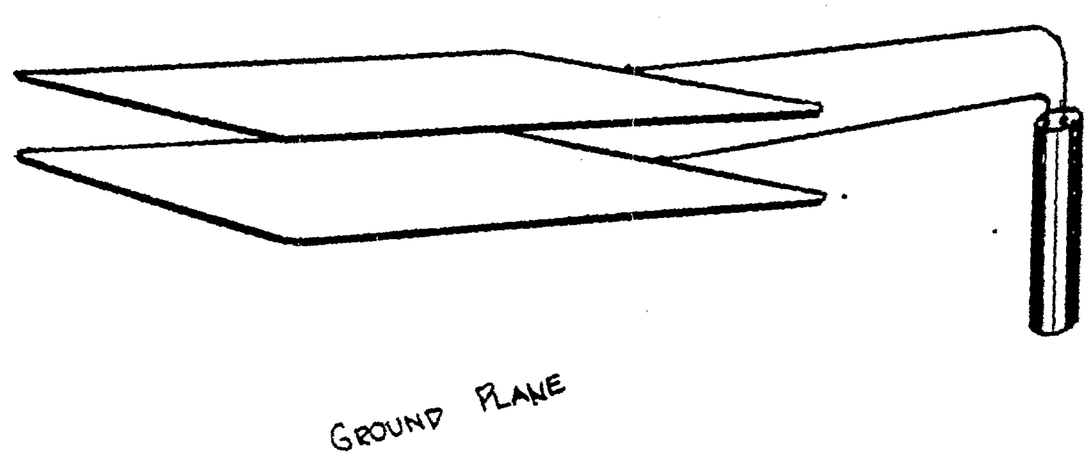
$$Z \simeq 10^8 \text{ ohms} \tag{9}$$

and the assumed conductance of equation (8) then the desired sensor response given by equation (3) will be met for

$$\gamma >> 3 \times 10^2 \ \frac{\text{roentgens}}{\text{sec}} \tag{10}$$

However, there are other conditions under which the characteristics of equation (3) can be achieved. First, one can calculate a time constant from the sensor capacitance and the load resistance as

$$\tau = C_s Z \simeq 10^{-2} \text{ seconds} \tag{11}$$



A = AREA OF ONE SIDE OF ELECTRODE
d = PLATE SPACING
DIPOLE ELEMENTS ARE PARALLEL TO GROUND PLANE

FIG. 2. PARALLEL PLATE ELECTRIC FIELD DIPOLE

Thus, for all frequencies much greater than about 16 cps the sensor capacitance is sufficient (with the assumed load resistance) to make the value of the sensor conductance immaterial. This means that the load resistance can be lowered to some extent without degrading the sensor response. The sensor capacitance must still be much larger than the cable capacitance (as discussed in SSN VI) to optimize the high frequency response.

Thus, a first constraint or design consideration for an electric field dipole is that it should be unloaded. Ideally, it just samples the electric field, drawing a minimum of energy from the field. Then the time changing nature of the air conductivity should have no significant effect on the sensor performance.

III. Non-Linear Electron Mobility

Next consider that the electron mobility is non-linear in the electric field above about 10^3 volts/meter. This is illustrated in figure 3 for the cases of dry air and air with 1% water vapor (molecular fraction). As one can see in the figure the electron mobility can vary about an order of magnitude as the electric field varies between 10^3 and 10^5 volts/meter. If the electric field dipole structure is such as to significantly distort the electric field lines in the immediate vicinity of the sensor then in turn the conductivity will be altered in this immediate vicinity because the conductivity is proportional to the electron mobility (at least during the radiation pulse when the electrons dominate the conductivity) and the electron mobility depends on the electric field. Therefore, for electric fields greater than 10^3 volts/meter any dipole structure which significantly distorts the electric field will cause the air conductivity to be non-uniform (i.e., a function of position) in the vicinity of the sensor.

This non-uniform conductivity will change the effective height, h, of the dipole. If the air conductivity were constant in the vicinity of the dipole the distribution of the electric field lines would be the same as the case of a constant permittivity, ϵ. This is because the electric field distribution, and thus the effective height, comes from a solution of Laplace's equation which cannot be applied to the case of a non-uniform conductivity. Since the effective height is influenced by both the conductivity and the permittivity the effective height can then be a function of both the electric field and the time derivative of the electric field. This last effect can be seen by assuming for the moment a conductivity which is non-uniform in space but constant in time. Then for sufficiently high frequencies the permittivity (essentially ϵ_o since the collision frequency of electrons in air is much higher than frequencies of interest) will be important, giving one effective height, while for sufficiently low frequencies the conductivity will be the significant factor, giving a different effective height. This would be a rather undesirable sensor characteristic.

Of course one answer to this problem is to avoid significant distortion of the electric field distribution in the vicinity of the dipole. To do this one must first know the direction of the electric field and then position the dipole elements so that they are perpendicular to the field lines. One possible dipole configuration which satisfies these requirements is the parallel plate device shown in figure 2. If the dipole elements are near to and parallel to a conducting ground plane (i.e., conducting compared to the

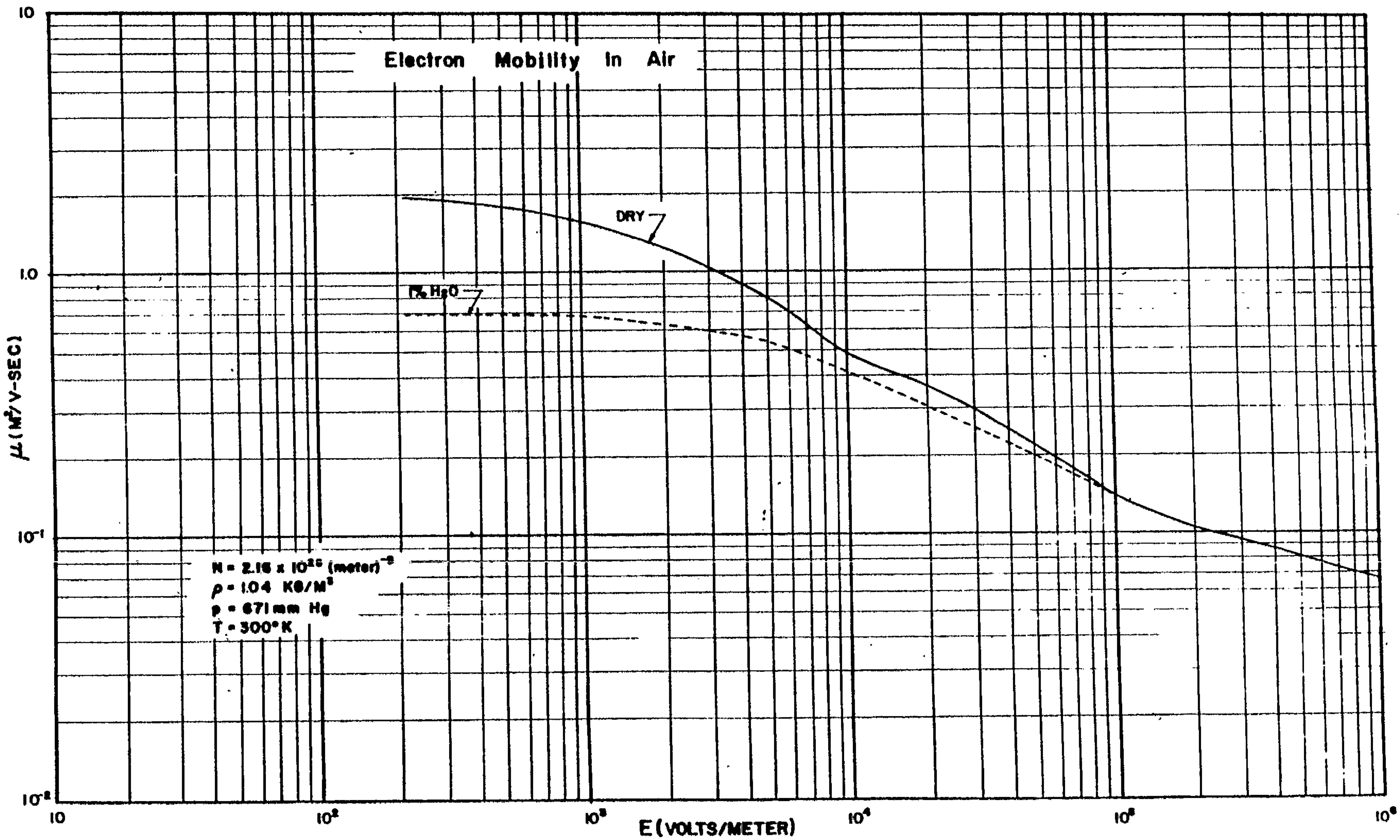

Electron Mobility in Air
DRY
1% H₂O
N = 2.16 x 10²⁵ (meter)⁻³
ρ = 1.04 KG/M³
p = 671 mm Hg
T = 300°K
μ (M²/V-SEC)
10
1.0
10⁻¹
10⁻²
10
10²
10³
10⁴
10⁵
10⁶
E (VOLTS/METER)

Figure 3

air) then the electric field will be essentially perpendicular to this plane and thus also to the dipole electrodes, and the effective height will be just the electrode spacing, d. Such a dipole should not significantly distort the electric field except at the connections to the signal cable. However, by making the dipole elements much larger (electrically) than the signal cable leads (and by insulating the signal cable leads to avoid electrical breakdown) the perturbing effect of these leads can be made negligible.

Thus, a second constraint or design consideration for an electric field dipole is that it should not significantly distort the electric field. Then the non-linear property of the electron mobility in air should have no significant effect on the sensor performance.

IV. Plasma Sheath Effects

A. Debye Length

A third characteristic of the sensor environment concerns the interaction of the dipole elements with the conducting air which is a conducting gas whose charge carriers consist of electrons and positive and negative ions. This ionized gas will exhibit plasma sheath effects. Since the electrons have much greater mobility than the ions (a factor of about 10^3 to 10^4) then in the absence of an electric field the electrons can drift to the dipole elements, leaving behind a region of positive charge whose thickness is characterized by the Debye length, h_D, given by

$$h_D = \left(\frac{\varepsilon_o KT_e}{2ne}\right)^{1/2} \text{ meters} \tag{12}$$

where KT_e is the electron temperature expressed in electron volts. (An approximate number for this is about 0.1 ev) As before, n is the electron density. Thus,

$$h_D \approx \frac{1.7 \times 10^3}{\sqrt{n}} \text{ meters} \tag{13}$$

From equation (1)

$$n \approx 2.1 \times 10^7 \gamma \frac{\text{electrons}}{\text{meter}^3} \tag{14}$$

(γ in roentgens/sec) and during the radiation pulse

$$h_D \approx \frac{.37}{\sqrt{\gamma}} \text{ meters} \tag{15}$$

and thus the Debye length should be of little concern, being small compared to the dipole dimensions. When the electron density dies away the ions will determine this characteristic distance. The voltages associated with this electron depletion layer will be a small fraction of a volt and thus insignificant.

B. Charge Carrier Depletion

Traditional consideration of a plasma sheath by calculating the Debye length is not the only interaction of the dipole elements with the ionized air. A more important concern is the effect which charge carrier depletion at the dipole electroces has on the spatial distribution of the air conductivity (near the dipole) and the effect of the electric field on the charge depletion at the dipole electrodes.

As a worst case one can assume that no low energy electrons are ejected from the dipole electrodes by the radiation, i.e., that the secondary emission coefficient (for Compton electrons incident on the electrode surface from either side) is zero. Then the currents which can flow between the electrode surface and the gas are conducted by the electrons and negative ions for currents out of the electrode surface, but only by the positive ions for currents into the electrode. Looking at the parallel plate dipole of figure 2 (for which the calculations are easier) one can estimate these effects.

Consider first the distance which an electron drifts under the influence of the electric field before attaching to a neutral oxygen molecule and forming a negative ion. To estimate this distance one needs first the relation between the Compton current density, $\vec{J}_c$*, and the* γ *current,* $\vec{\gamma}$*, which is*

$$\vec{J}_c \approx -2 \times 10^{-8} \vec{\gamma} \ \frac{\text{amps}}{\text{meter}^2} \qquad (16)$$

(where $\vec{\gamma}$ *is expressed in roentgens/sec, as explained in SSN IX). If it is assumed for the purposes of this approximate calculation that the conduction current density,* J*, is of the same magnitude as* $\vec{J}_c$ *and that the gamma flux is approximately the same as the gamma current, i.e.,*

$$J \approx 2 \times 10^{-8} \gamma \ \frac{\text{amps}}{\text{meter}^2} \qquad (17)$$

then relating the conduction current to the electric field as

$$J = \sigma E = en\mu(E)E \ \frac{\text{amps}}{\text{meter}^2} \qquad (18)$$

and noting that the electron drift velocity, v_d*, is given by*

$$v_d = \mu(E)\, E \ \frac{\text{meters}}{\text{sec}} \qquad (19)$$

the drift velocity is approximately

$$v_d = \frac{J}{en} \approx \frac{2X10^{-8}\gamma}{en} \qquad (20)$$

Then referring to equation (1) for the electron density one has

$$v_d \approx \frac{2X10^{-8}\gamma\nu}{e(2.1X10^{15}\gamma)} \approx .6 \times 10^{4} \ \frac{\text{meters}}{\text{sec}} \qquad (21)$$

This can be converted to a drift distance, d_d, given by

$$d_d \approx \frac{v_d}{\nu} \approx .6 \times 10^{-4} \text{ meters} \tag{22}$$

This distance is rather small compared to the dimensions of a realistic dipole and this distance should characterize the thickness of an electron depletion layer next to a dipole electrode.

However, perhaps a more meaningful parameter is the voltage which might develop across this depletion layer. One can estimate this voltage, V_d, calculating the charge, Q, depleted per unit area as

$$Q \approx \int J\, dt \approx 2 \times 10^{-8} \int \gamma dt \ \frac{\text{coulombs}}{\text{meter}^2} \tag{23}$$

The voltage across this depletion layer is then about

$$V_d \approx Q \frac{d_d}{2\varepsilon_0} \approx .07 \int \gamma dt \text{ volts} \tag{24}$$

For close-in measurements this voltage can be a significant error.

Of course there will be electrons emitted by the dipole electrodes because of the presence of the radiation. Some of the positive ions will also drift to the electrodes, recombining with electrons there allowing some more current to flow into the electrodes. It would seem then that these effects should be enhanced. In particular, the secondary electron emission from the electrodes should be enhanced, if possible (by use of special coatings), so that the secondary electron emission coefficient is greater than one (or even much greater to be sure). Then at least during the radiation pulse (when electronic conductivity is important) the conduction current in the air will not be significantly perturbed by the presence of the sensor. During late times when the ionic conductivity is important (compared to the electronic conductivity) the increase in the secondary emission coefficient, because of the lower radiation level, may not be sufficient to prevent the formation of a charge depletion layer at the electrode surfaces. However, the much lower conductivities at these times may make the voltages associated with such a depletion layer much smaller than those possible because of electron depletion (as in equation (24)).

Due to the complexities of the processes involved with the interaction of the dipole surfaces with the ionized air it may be difficult to pin down the details of the processes quantitatively beyond the very approximate calculations of this section. To be safe one should then try to minimize these effects as much as possible. Unfortunately, the parallel plate electric field dipole of figure 2, because it obstructs the conduction current in the air (which is in the direction of the electric field and thus perpendicular to the plates) tends to maximize these interface problems while from the point of view of the non-linear electron mobility (in Section III) the configuration is quite advantageous.

C. *Wire Mesh "Parallel Plate" Dipole*

Ideally, one might think that a dipole configuration which allowed most of the conduction current to flow around the dipole conductors instead of through them could avoid much of these problems. Such a device is shown in figure 4 in which the parallel plates have been replaced by parellel wire meshes of the same overall area, A. (The actual area of metal surface is of course much less. The different hook-up to the signal cable has no special significance other than to indicate that there are several different techniques for accomplishing such a connection.) If the wire diameter is much less than the wire spacing in the mesh then most of the conduction current should pass between the wires and not through them providing that the wire spacing is much larger than either the electron drift distance in equation (22) or the plasma sheath thickness governed by the Debye length as in equation (13) or any other depletion layer thickness. Now, since the current can flow "through" the electrodes with little perturbation the electrode potential should be that of the air in its immediate vicinity with a difference given by the epithermal potential of the electrons (about .1 ev) which should be negligible. Thus, such a configuration should be capable of avoiding what can be grouped as the plasma sheath effects while at the same time retaining the advantages of parallel plate geometry.

D. *Effect of Using Wire Mesh on Dipole Capacitance*

It might be objected that by going to this type of electrode design the sensor capacitance, C_s, will be appreciably lowered. However, this turns out not to be the case as calculated by Dr. Partridge of LASL and this author in determining the effect of replacing parallel plates by a network of wires in the EMP testing facility now under construction here at Kirtland AFB (see SSN I for the original design considerations for this facility). This can be illustrated by approximately calculating the capacitance between two "plates" of parallel wires as in figure 5 with the assumptions that the wire radius, a, is much less than the wire spacing, b, which is in turn much less than the "plate" spacing, d, i.e.,

$$a \ll b \ll d \qquad (25)$$

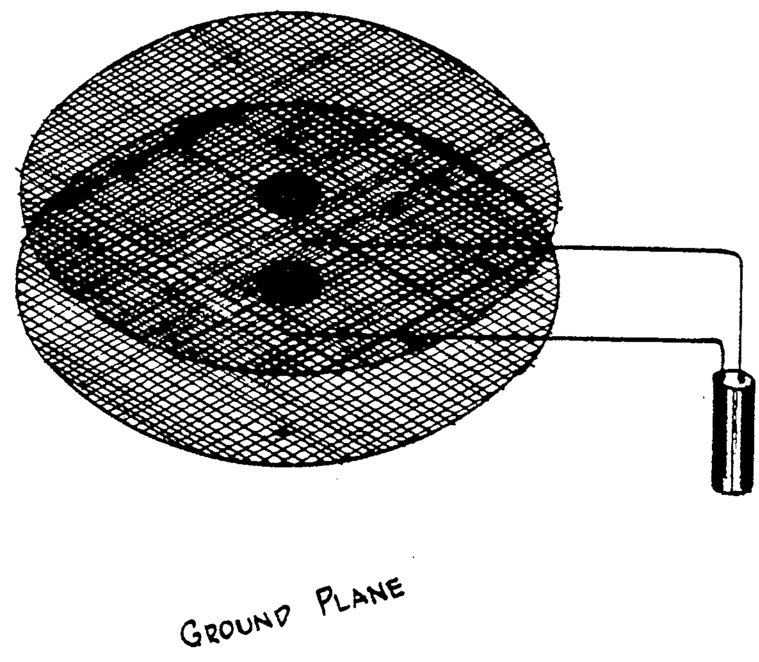
To calculate the capacitance per unit area assume a charge per unit length on each wire, q, (of either sign as appropriate) and calculate the potential between the two arrays of wires. Near one of the wires the electric field, E, will be given by

$$E = \frac{q}{2\pi\epsilon_o r} \quad \frac{\text{volts}}{\text{meter}} \qquad (26)$$

where r is the distance from the center of the wire. Integrating this field out to a distance equal to half the wire spacing one has a voltage contribution, V_1, as

$$V_1 = \frac{q}{2\pi\epsilon_o} \ ln\left(\frac{b}{2a}\right) \text{volts} \qquad (27)$$

At about this distance, b/2, the equipotential lines will begin flattening. The voltage, V_2, between points, b/2, away from each wire mesh will be



A = AREA OF ONE SIDE OF ELECTRODE (IGNORING SPACING BETWEEN WIRES IN EACH MESH)
d = ELECTRODE SPACING

DIPOLE ELEMENTS ARE PARALLEL TO GROUND PLANE.

FIG. 4. WIRE MESH "PARALLEL PLATE" DIPOLE.

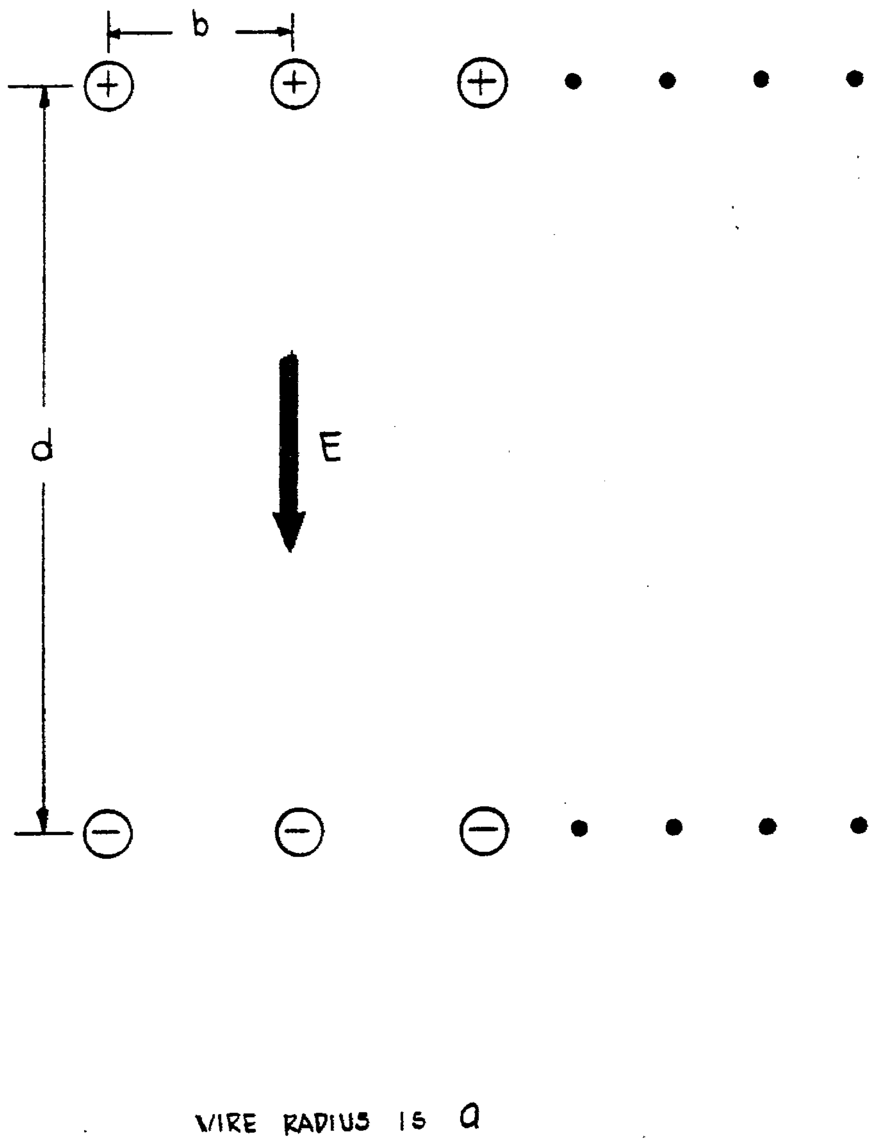


FIG. 5. IDEALIZED WIRE MESH CAPACITANCE CALCULATIONS.

$$V_2 \approx (d-b)\,\frac{q}{b\varepsilon_o} \quad \text{volts} \tag{28}$$

since q/b is the average charge per unit area for each wire mesh. The voltage, V, between the two meshes will then be

$$V \approx 2V_1 + V_2 = \frac{q}{\pi\varepsilon_o}\, ln\left(\frac{b}{2a}\right) + \frac{q}{\varepsilon_o b}\,(d-b) \quad \text{volts} \tag{29}$$

and the capacitance per unit area, C', will be

$$C' = \frac{q}{bV} \approx \varepsilon_o \left[\frac{b}{\pi}\, ln\left(\frac{b}{2a}\right) + d-b\right]^{-1} \quad \text{farads} \tag{30}$$

or

$$C' \approx \frac{\varepsilon_o}{d}\left[1 + \frac{b}{d}\left(\frac{1}{\pi}\, ln\left(\frac{b}{2a}\right) - 1\right)\right]^{-1} \quad \text{farads} \tag{31}$$

while in the true parallel plate case the capacitance per unit area is just ε_o/d. To illustrate with some numbers let

$$\frac{b}{2a} = 100$$

$$\frac{d}{b} = 10 \tag{32}$$

Then equation (31) would imply a correction in C' of about 5%, not a significant effect. Since the wire diameter is contained in the logarithm term one can make this diameter extremely small without appreciably affecting the capacitance or the frequency response (as implied in equation (11)).

To first order the sensor conductance should be changed with about the same correction factor, neglecting the plasma sheath effects around the wires. Including these effects the dipole conductance should be decreased. However, if the sensor is not loaded (as is desirable per Section II) this perturbation in the sensor conductance should have little effect.

E. Conclusions

There are at least two general methods to get around what can be classed as plasma sheath effects:

(1) Utilize the gamma radiation to place low energy secondary electrons back into the air to replace those which are pulled away from the electrodes by the electric field. To do this the secondary electron emission coefficient must be maximized, perhaps by the use of a thin layer (a few atoms or molecules thick) of some material such as might be used in photomultiplier tubes.

(2) Allow the conduction current in the air to flow around the electrode material, not through it. This can be accomplished by the use of wire mesh electrodes without sacrificing other important electrical properties. This procedure is one of evading the problem and for this reason perhaps more successful.

However, since these two techniques are not mutually exclusive, it may be wise to use both.

V. Compton Current Effects on Dipole

Turning now to the direct radiation effects on the dipole, the most important of these mechanisms is Compton scattering, in which the γ rays scatter high energy electrons (of about half the energy of the γ ray) from the air into the dipole structure and from the dipole structure back into the air. Essentially then the electric field dipole acts as a γ radiation detector called the Compton diode, because a certain net current, I_c, indicated in the equivalent circuit in figure 1, is caused to appear as a noise signal. To estimate this current one can look at the relationship in air (and low atomic number materials) between the Compton current density and the gamma current given by equation (16). A certain fraction of these Compton electrons impinging on the dipole will be collected and to first order the net current deposited in any part of the dipole structure will be related to the Compton electron flux at that point times the γ-ray attenuation factor of that part of the dipole structure.

There are certain techniques which one can use to minimize this net Compton current, I_c, in the dipole circuit (in relation to the signal from the electric field). These techniques will now be discussed.

A. Symmetrical Dipole

A first technique which should come to mind is that of making the dipole symmetrical, as is shown in both figures 2 and 4. The dipole electrodes should be essentially identical. Then the net Compton current, I_c, would only be the difference between the Compton currents deposited in each of these electrodes, a reduction of perhaps one to two orders of magnitude from the current deposited in either electrode. This technique excludes, for example, a vertical whip antenna with a ground plane or any other such unsymmetrical sensor. This requires that there be signal leads into the environment but the radiation currents which these collect can be made small compared to the radiation currents in the dipole. It is not the total Compton current in the dipole which is important but the current per unit electrode area because an increase in size not only increases I_c, but also increases the dipole capacitance, C_s, and conductance, $G_s(t,E)$ (to be discussed later), which determine the voltage produced by I_c. In addition, of course, the signal leads can themselves be made symmetrical by the use of twinax (balanced shielded pair) to similarly reduce the Compton current signal introduced at this point.

B. Minimization of Electrode Mass

Another technique for reducing I_c is to reduce the electrode mass while keeping the effective electrical surface area a constant. Essentially only the metal surface of the electrodes (to a few skin depths) contributes to the electrical properties of the dipole. Anything else is useful only for structural support and should be removed to as great an extent as possible, minimizing the γ-ray attenuation of the dipole. Since the dipole will have to be unloaded (as discussed in Section II), then for pulse widths longer than the electrical round trip transit time on the dipole and associated signal cable, essentially no current will be drawn from the dipole. Thus, the electric structure need be no more than a few skin depths thick at the frequency corresponding to this transit time.

To get a rough idea of how thick a conductor this implies consider a sensor-cable system with a round trip transit time,

$$t_r \approx 10 \text{ ns} \tag{33}$$

Equating t_r to the reciprocal of the half period at radian frequency, ω, in the skin depth formula one has for aluminum approximately

$$\delta = \sqrt{\frac{2}{\omega\mu\sigma}} \approx \sqrt{\frac{2\ t_r}{\pi\mu\sigma}} \approx .12\sqrt{t_r} \tag{34}$$

or

$$\delta \approx .12 \times 10^{-4} \text{ meters} \tag{35}$$

or

$$\delta \approx .47 \times 10^{-3} \text{ inches} \tag{36}$$

As one can see this dimension is extremely small allowing one to make the sensor electrodes extremely thin and thus for parallel plate electrodes (as in figure 2) attenuate the γ rays normally incident on the electrodes by much less than one percent.

One can also compare this skin depth (of equation (35)) with the range of a .5 Mev Compton electron in Aluminum which is about

$$Re \approx .67 \times 10^{-3} \text{ meters} \tag{37}$$

or

$$Re \approx 2.6 \times 10^{-2} \text{ inches} \tag{38}$$

Thus, for very thin electrodes it is possible to even make the electrodes somewhat "transparent" to the Compton electrons. Once this point is reached then differences in the ratio of the Compton electron range to the γ-ray mean free path between air and the electrode material, as well as various geometric factors should be the parameters determining the net Compton current in the dipole circuit. The net Compton current into the dipole structure should then be small compared to the conduction currents in the air in the vicinity of the dipole.

There is yet another technique for reducing the Compton current desposited in the electrodes. In considering the plasma sheath effects (in Section IV) it was found that the parallel plate structure of the dipole could be advantageously modified by replacing the solid plates by a wire mesh such that the physical electrode area is much smaller than the total geometrical area encompassed by the electrode. This same design has a significant advantage in the reduction of the Compton current signal because this signal is proportional to the physical electrode area (not the total geometrical area or effective electrical area, both of which remain essentially constant). Thus, by using wire mesh electrodes I_c can be reduced another two orders of magnitude.

A final way to reduce the mass is to simply reduce the density of the electrode material. This would seem to have limited application beyond noting that conducting materials having lower densities have a built-in advantage from this viewpoint. Thus, for example, aluminum would be preferred over copper.

C. Utilization of the Sensor Capacitance and Conductance

Combining the techniques of symmetry in the dipole structure and the two techniques of minimizing the electrode mass (making the electrodes thin and then going a step further to thin wires to reduce the conductor area) one can perhaps obtain a Compton current signal (per unit area of one electrode) which is down about 6 orders of magnitude (about 2 orders of magnitude for each of the first three techniques) from the Compton current density in the air. Thus, perhaps

$$I_c \approx 2 \times 10^{-14} \gamma A \quad \text{amps} \tag{39}$$

Then during the radiation pulse one can compute a Compton noise signal, V_c, by use of the sensor conductance and electrode area from equations (6) and (8). Thus,

$$V_c = \frac{I_c}{G_s(t,E)} \approx .6 \times 10^{-3} \text{ volts} \tag{40}$$

which for EMP purposes is quite negligible. The sensor conductance is actually somewhat smaller than indicated by equation (8). By using the wire mesh dipole electrodes there is a correction to the conductance as well as to the capacitance of the dipole, but this type of correction is small. More important the conductance will be lowered by the plasma sheath effects at the wires (as discussed in Section IV). However, as indicated by equation (40) there is a safety margin of a few orders of magnitude. After the radiation pulse, the Compton signal should be much smaller because the conductivity will become dominated by the ions and hold up for some time.

Thus, the air conductivity alone is sufficient to keep the Compton noise signal down, far below the signal level if the various minimization techniques (and others as they are discovered) as outlined in this section are employed. The sensor capacitance also tends to keep the Compton noise signal down, but only for short transients.

VI. Other Radiation Effects

In addition to the Compton current effects, noise signals can also be generated by other specie of radiation, most importantly neutrons and X rays. Neutrons can be important from at least two standpoints:

(1) The neutrons can produce γ rays by inelastic scattering in some material associated with the dipole. These γ rays in turn can produce Compton noise signals in the dipole. The design criterion in this case should of course be to use materials with as small inelastic neutron cross sections as possible.

(2) The neutrons can produce proton recoil currents in hydrogenous materials by elastic scattering with the hydrogen nuclei. This would imply that such materials should be avoided, particularly where they might introduce these currents into the electrical structure of the dipole.

X-rays are important because of the photoelectrons which they can eject from the surface of the dipole conductors. Generally, to minimize this effect one should use low atomic numbe[illegible]terials. Ideally, one might think of trying to match the average [illegible] for the [illegible] -tors then aluminum m[illegible] choice. [illegible] [illegible] photoelectrons in air w[illegible] much less than that of the Compton [illegible] -rons ma[illegible] th[illegible] second order effect.

[illegible]e Compton electrons associated [illegible] the γ rays present the most significant radiation noise signal problem because by proper choice of materials these other effects should be minimal.

[illegible]II. Summary

Briefly then it does not appear impossible to build an electric field dipole which is capable of measuring the electric field associated with the close-in EMP. However, there are certain constraints placed on such a sensor by the nuclear radiation and the conducting air. The adverse effects of the environment can, at least under present theory, be averted by

(1) operating the dipole unloaded. This avoids the problem of the time changing nature of the air conductivity.

(2) minimizing distortion of the electric field lines by placing the dipole electrodes along two different equipotential surfaces. This avoids the problem of the non-linear electron mobility. However, it also requires that the direction of the electric field be known, as, for example, near a conducting surface.

(3) allowing the conduction currents in the air to flow around the dipole elements instead of into and out of the conducting surfaces, thus, avoiding the interface problems between the air and the dipole elements. This can be accomplished while retaining the advantages of parallel plate geometry by the use of wire mesh electrodes. In addition, special coatings on the electrode surfaces can help in reducing this problem by increasing the secondary electron yield.

(4) minimizing the mass of the dipole electrodes and building the dipole symmetrically. The combination of these procedures can reduce the Compton noise signal to acceptable levels. A symmetrical dipole with wire mesh electrodes using a low density conductor like aluminum goes a long way toward minimizing the Compton signal.

(5) avoiding materials with large (n, γ) cross sections or materials with hydrogen if these latter materials must be in contact with the dipole electrodes.

(6) avoiding high atomic number materials to minimize X-ray effects.

It seems, however, that since so many factors come into play in this electric field dipole design that it would be wise to ultimately subject such a dipole to some realistic simulation of these effects.

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11 February 1965