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Interaction Notes

Note 505

30 July 1994

SEM Representation of Signals at Internal Ports of Complex Electronic Systems

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Abstract

In understanding the response of a complex electronic system to externally incident electromagnetic waves, the system natural frequencies (complex resonances) are typically of dominant importance. These poles have residues given by a coupling vector which is a function of the angle of incidence. These terms describe the effective height for internal ports (failure ports), showing how large the signals there can be over the various exposure possibilities. These terms can be measured either in reception, or by reciprocity in transmission. For cases where there is an approximate shield around the system, the coupling vectors for the external surface current density can be conveniently applied to the internal ports at these external natural frequencies.

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In understanding the response of a complex electronic system to externally incident electromagnetic waves, the system natural frequencies (complex resonances) are typically of dominant importance. These poles have residues given by a coupling vector which is a function of the angle of incidence. These terms describe the effective height for internal ports (failure ports), showing how large the signals there can be over the various exposure possibilities. These terms can be measured either in reception, or by reciprocity in transmission. For cases where there is an approximate shield around the system, the coupling vectors for the external surface current density can be conveniently applied to the internal ports at these external natural frequencies.

I. Introduction

The interaction of an incident plane electromagnetic wave with a complex electronic system is quite complicated in its details. However, some general characteristics of the system response can be used to summarize the form of the response in a way which lends itself to efficient parameterization for both theoretical and experimental purposes. As discussed in [3, 7, 12, 16] the response is dominated by poles (complex resonances) as described by the singularity expansion method (SEM). In time domain these correspond to damped sinusoids. In frequency domain they correspond to peaks on the $j\omega$ axis of the s (= $\Omega + j\omega$) plane or complex-frequency plane.

As discussed in [3, 12, 16] the range of these natural frequencies extends from some low frequency f_ℓ, based on the external dimensions of the system and lengths of interior cables (perhaps modified by loading impedances), to some high frequency f_h (typically of the general order of a GHz), based on the small resonant dimensions of concern. Below f_ℓ the coupling is typically propertional to f (time derivative for electrically small objects), and above f_h it is proportional to f^{-1} (time integral for coupling to the wires) and can fall off even more rapidly due to propagation losses on the cables. It is this band of frequencies with emphasis on resonances near f_h (due to increasing antenna gain for microwave illuminators) that are of importance for understanding the system response and design of appropriate protection (hardening).

II. Response at Internal Port

Consider a system as in fig. 2.1. Illustrated is some nth ($n = 1, 2, \ldots, N$) internal ports which can also be referred to as failure ports [6, 7]. Each of these is a terminal pair at which there is a voltage $V_n(t)$ which is closely related to a potential failure (upset or permanent damage) of the system. This can be at the pins leading into an equipment box or across some transistor inside.

Besides the voltage at the port there is an impedance $\tilde{Z}_n$ determined by driving the port by a current source $I_n^{(in)}$ and measuring the resulting voltage $\tilde{V}_n^{(in)}$ to give $\tilde{Z}_n$ as the ratio. Note the convention of positive current into the port (+ convention terminal). Now there may be some device directly connected across the terminals of impedance $\tilde{Z}_n^{(L)}$ with some effective source impedance $\tilde{Z}_n^{(s)}$ representing the rest of the system. These combine to give

$$\tilde{Z}_n(s) = \left[\tilde{Z}_n^{(s)^{-1}}(s) + \tilde{Z}_n^{(L)^{-1}}(s)\right]^{-1} = \frac{\tilde{V}_n^{(in)}}{\tilde{I}_n^{(in)}(s)} \tag{2.1}$$

as the important impedance for our analysis. Note that time invariance and linearity, or more precisely linearity to failure in the sense of [6], have been assumed. At the port the voltage $\tilde{V}_n$ is an open-circuit voltage in this convention. The associated short-circuit current is

$$\tilde{I}_n^{(s.c.)}(s) = -\tilde{Z}_n^{-1}(s)\ \tilde{V}_n(s) \tag{2.2}$$

where the minus sign accounts for our above-stated current convention. Note that open-circuit conditions correspond to the system conditions of interest for its response to an incoming plane wave.

An alternate way to view the system response is to redefine the port by breaking the branch (via one of the leads) containing $\tilde{Z}_n^{(L)}$, and inserting the terminal pair at this break. Then one can use the short-circuit current at this port to define $\tilde{I}_n$. This is a dual procedure to the one we are discussing, and it carries through in a similar manner.

Assume an incident wave as a plane wave of the form

$$\tilde{\vec{E}}^{(inc)}(\vec{r},s) = E_o\tilde{f}(s)\,\vec{1}_p\,e^{-\gamma\vec{1}_i\cdot\vec{r}}\ ,\ \tilde{\vec{E}}^{(inc)}(\vec{r},t) = E_o\,\vec{1}_p\,f\left(t - \frac{\vec{1}_i\cdot\vec{r}}{c}\right)$$

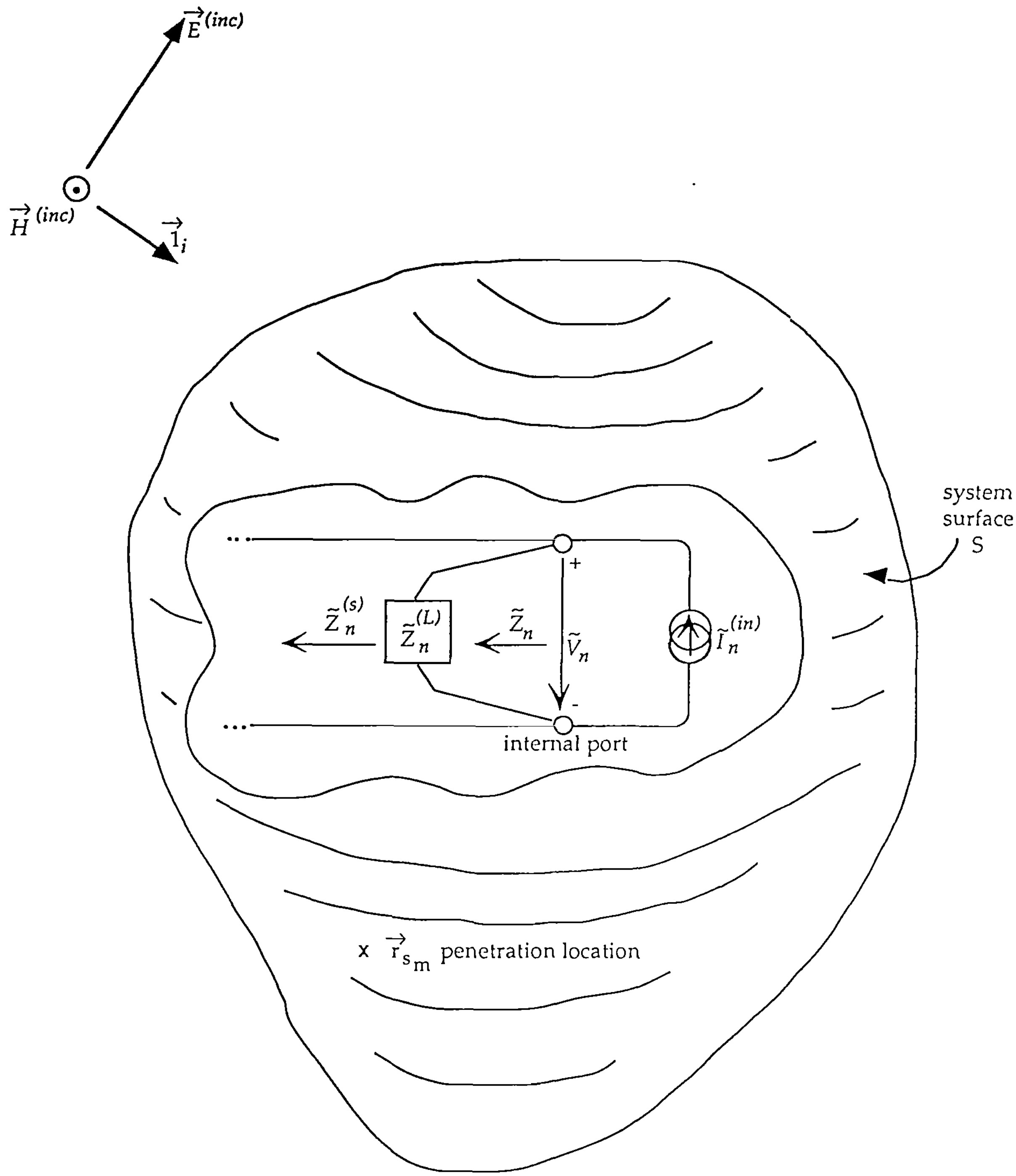

Fig. 2.1 System with Approximate Perfectly Conducting Enclosing Surface, Penetrations, and Internal Ports.

$$\tilde{\vec{H}}^{(inc)}(\vec{r},s) = \frac{E_o}{Z_o}\tilde{f}(s)\,\vec{1}_i \times \vec{1}_p\, e^{-\gamma \vec{1}_i \cdot \vec{r}} \;,\; \tilde{\vec{H}}^{(inc)}(\vec{r},t) = \frac{E_o}{Z_o}\vec{1}_i \times \vec{1}_p\, f\left(t - \frac{\vec{1}_i \cdot \vec{r}}{c}\right)$$

$$\begin{aligned}
&f(t) \equiv \text{incident waveform} \\
&c \equiv [\mu_o\, \epsilon_o]^{-\frac{1}{2}} \;,\; Z_o \equiv \left[\frac{\mu_o}{\epsilon_o}\right]^{\frac{1}{2}} \;,\; \gamma \equiv \frac{s}{c} \\
&\vec{1}_i \equiv \text{direction of incidence (real unit vector)} \\
&\vec{1}_p \equiv \text{polarization of incident wave (real unit vector)} \\
&\vec{1}_i \cdot \vec{1}_p \equiv 0 \\
&\sim\; \equiv \text{Laplace transform (2-sided) over time} \\
&s \equiv \Omega + j\omega \equiv \text{Laplace-transform variable or complex frequency}
\end{aligned} \tag{2.3}$$

In reception the voltage is best expressed in terms of the traditional effective height as

$$\tilde{V}_n(s) \equiv \tilde{V}_n^{(o.c.)}(s) \equiv \tilde{\vec{h}}_{V_n}(\vec{1}_i, s) \cdot \vec{1}_p\, E_o\, \tilde{f}(s) \tag{2.4}$$

There are other forms this can take for current and wave variables [4]. Note that this effective height retains phase information and is a linear parameter, unlike the quadratic absorption cross section. Note that the effective height can be defined with the constraint

$$\tilde{\vec{h}}_{V_n}(\vec{1}_i, s) \cdot \vec{1}_i = 0 \tag{2.5}$$

since the incident electric field is always perpendicular to $\vec{1}_i$.

Now if we remove the incident field and drive the port with a current, the system can be thought of as a transmitting antenna with a far field as

$$\begin{aligned}
&\tilde{\vec{E}}_f(\vec{r}, s) = \frac{e^{-\gamma r}}{r}\, \tilde{\vec{F}}_{I_n}(\vec{1}_r, s)\, \tilde{I}_n^{(in)}(s) \\
&\vec{1}_r \equiv \text{direction to observer (in far field)} \\
&\tilde{\vec{F}}_{I_n}(\vec{1}_r, s) \cdot \vec{1}_r = 0 \text{ (transverse far field)}
\end{aligned} \tag{2.6}$$

Assuming that our system is constructed of reciprocal media (symmetric constitutive-parameter matrices) and reciprocal devices (symmetrical impedance/admittance matrices) then the reciprocity theorem [4] can be invoked to give

$$\tilde{F}_{I_n}(\vec{1}_r, s) = -s\frac{\mu_o}{4\pi}\,\tilde{\vec{h}}_{V_n}(-\vec{1}_r, s) \tag{2.7}$$

So one can make measurements (or calculations) in reception or transmission and have the result for both cases via this equation. Note that while the patterns (in frequency domain) are the same (except for a reversal of direction) in reception and transmission, in time domain there is a time derivative (or integral) relating the two cases.

Now write the response in SEM form as

$$\tilde{V}_n(s) = E_o\,\tilde{f}(s)\sum_{\alpha} n_\alpha^{(V_n)}(\vec{1}_i, \vec{1}_p)\left[s - s_\alpha\right]^{-1} e^{-[s-s_\alpha]t_{i_n}} \tag{2.8}$$

As discussed in [10] there is no need for an entire function if the turn-on-time t_{i_n} is chosen appropriately. While this result is based on the surface-current density on a perfectly conducting scatterer, the derivation is not essentially changed if resistive materials are included. However, it is known that losses can be included in scatterers/antennas in such a way as to produce second- or even higher-order poles [5]. For present purposes we assume that only first-order poles are present for the system under consideration. The natural-frequency index α is common for all the ports of the system, although particular resonances can be strong in some ports and weak in other ports (with special cases, e.g. due to symmetry, of zero contribution in some ports).

Suppressing the turn-on time by appropriate choice of $t = 0$, then (2.4) gives

$$\tilde{\vec{h}}_{V_n}(\vec{1}_i, s)\cdot\vec{1}_p = \sum_{\alpha} n_\alpha^{(V_n)}(\vec{1}_i, \vec{1}_p)\left[s - s_\alpha\right]^{-1} \tag{2.9}$$

Choosing two orthogonal choices for $\vec{1}_p$ as $\vec{1}_2$ and $\vec{1}_3$ (both perpendicular to $\vec{1}_i$) we have

$$\tilde{\vec{h}}_{V_n}(\vec{1}_i, s) = \sum_{\alpha} \vec{C}_\alpha^{(V_n)}(\vec{1}_i)\left[s - s_\alpha\right]^{-1}$$

$$\vec{C}_\alpha^{(V_n)} = \vec{1}_2\, n_\alpha^{(V_n)}(\vec{1}_i, \vec{1}_2) + \vec{1}_3\, n_\alpha^{(V_n)}(\vec{1}_i, \vec{1}_3)$$

$$\vec{1}_2 \vec{1}_2 + \vec{1}_3 \vec{1}_3 = \overleftrightarrow{1} - \vec{1}_i \vec{1}_i \equiv \overleftrightarrow{1}_i \equiv \text{identity dyad tranverse to } \vec{1}_i \tag{2.10}$$

Applying reciprocity we also have

$$\tilde{\vec{F}}_{I_n}(\vec{1}_r, s) = -s \frac{\mu_o}{4\pi} \sum_\alpha \vec{C}_\alpha^{(V_n)}(\vec{1}_i)\left[s - s_\alpha\right]^{-1} \tag{2.11}$$

giving the response in transmission to a current source into the port. Note the commonality of the natural frequencies in transmission and reception.

Now we can see that the vectors $\vec{C}_\alpha^{(V_n)}(\vec{1}_i)$ are a convenient way to describe the response at the poles which in turn approximate the peaks in frequency domain (along the $j\omega$ axis). Since the spatial distribution of a natural mode is independent of the incident field, then the relative response of each of the N internal ports must be independent of $\vec{1}_i$ and $\vec{1}_p$, giving

$$\vec{C}_\alpha^{(V_n)}(\vec{1}_i) = \upsilon_{n,n'}^{(\alpha)} \vec{C}_\alpha^{(V_{n'})}(\vec{1}_i)$$

$$n' = \text{conveniently choses port with non-zero response to } \alpha\text{th mode} \tag{2.12}$$

So one can display the dependence of the αth response as a vector pattern common to all the ports time scaling constants for each port.

As discussed in [8, 11, 13, 15] there are cases for which there can be more than one natural mode for a given natural frequency. This is the case of modal degeneracy, typically associated with certain symmetries in the system (normal degeneracy). In such a case any linear combination of such natural modes is also a natural mode. The number of such linearly independent modes is the degree of the degeneracy, and the scaling in (2.12) needs to be applied to an appropriate set of such linearly independent modes.

III. Division into External and Internal Resonances

In a topological decomposition of a system there is a good-shielding approximation in which the penetration of signals through a closed boundary surface (shield) is sufficiently small that one can separately consider the volumes on each side of this boundary and treat the coupling through the boundary as a perturbation [9]. For present purposes, let us consider the outer system surface S in fig. 2.1 as such a shield.

Approximating S as a perfectly conducting surface we have an SEM representation of the surface current density [10] as

$$\begin{aligned}\tilde{\vec{J}}_s(\vec{r}_s, s) &= E_0\, \tilde{f}(s) \sum_{\alpha_{ex}} \eta_{\alpha_{ex}}^{(J_s)}(\vec{1}_i, \vec{1}_p)\, \vec{J}_{s_{\alpha_{ex}}}(\vec{r}_s)\left[s - s_{\alpha_{ex}}\right]^{-1} \\ &= E_0\, \tilde{f}(s)\, \vec{1}_p \cdot \sum_{\alpha_{ex}} \vec{C}_{\alpha_{ex}}^{(J_s)}(\vec{1}_i, \vec{1}_p)\, \vec{j}_{s_{\alpha_{ex}}}(\vec{r}_s)\left[s - s_{\alpha_{ex}}\right]^{-1} \qquad (3.1)\end{aligned}$$

$\vec{r}_s \in S$

$\alpha_{ex} \equiv$ index for external natural frequencies, modes, etc.

There are various ways to compute the natural frequencies, modes, coupling vectors (or coefficients), etc. which are considered elsewhere and are not repeated here. It is the form of the result that is important here. The various terms may even be found by experiment.

In the same form as discussed in [1, 2] let us consider that there are a set of M penetrations through S. Each penetration is labelled according to a particular component of the surface current density (or surface magnetic field) or surface charge density (normal electric field). So $\vec{r}_{s_m}$ refers to the location of a particular penetration and associated field component. More than one value of m may be associated with the same position on S. Let there be a transfer function $\tilde{T}_{n,m}(s)$ from each such $\vec{r}_{s_m}$ to the port voltage $\tilde{V}_n(s)$. For penetration proportional to the surface current density we have

$$\tilde{V}_n(s) = \tilde{T}_{n,m}(s)\, \vec{1}_m \cdot \tilde{\vec{J}}_s(\vec{r}_{s_m}, s)$$

$\vec{1}_m \equiv$ unit vector tangential to S specifying a particular component of the surface current density (3.2)

and for penetration proportional to the surface charge density we can write

$$\tilde{V}_n(s) = \tilde{T}_{n,m}(s)\ \tilde{\rho}_s(\vec{r}_{s_m}, s) = \tilde{T}_{n,m}(s)\left[-\frac{1}{s}\nabla_s \cdot \vec{\tilde{J}}_s(\vec{r}_s, s)\right]_{\vec{r}_s = \vec{r}_{s_m}} \tag{3.3}$$

Consider now some external natural frequency $s_{\alpha_{ex}}$ and let the $\tilde{T}_{n,m}$ have no poles near $s_{\alpha_{ex}}$. Then $\tilde{V}_n$ has a pole at $s_{\alpha_{ex}}$ and sampling the surface current density as in (3.2) and (3.3) is the same as sampling the natural mode $\vec{j}_{s_{\alpha_{ex}}}$ in (3.1). The dependence of $\tilde{V}_n$ on the direction of incidence and polarization is then described by $\vec{1}_p \cdot \vec{C}^{(J_s)}_{\alpha_{ex}}$. Interpreted in terms of the effective height as in (2.10) the angular dependence is the same as for the exterior modes and (2.12) can be written as

$$\vec{C}^{(V_s)}_{\alpha_{ex}}(\vec{1}_i) = v^{(\alpha_{ex})}_{n,o}\ \vec{C}^{(J_s)}_{\alpha_{ex}}(\vec{1}_i) \tag{3.4}$$

this result applying to all the $\tilde{V}_n$ or $\vec{\tilde{h}}_{V_n}$. So by measuring the coupling vectors for the surface current density modes we have, to a constant scaling factor, the coupling vectors for this set of resonances in all the *internal* responses.

Now the transfer functions $\tilde{T}_{n,m}(s)$ may also have poles with labels α_{in}. The various transfer functions may in general have different $s_{\alpha_{in}}$, but some may be common to various of the $\tilde{T}_{n,m}$, say due to a common resonant structure (e.g., a cavity) through which the signal passes to more than one port, and/or from more than one penetration. The $s_{\alpha_{in}}$ being, by hypothesis, separate from the $s_{\alpha_{ex}}$, then $\vec{\tilde{J}}_s(\vec{r}_s, s_{\alpha_{in}})$ is well behaved.

Note that $\vec{J}_s$ can be derived from an integral equation of the form [10]

$$\left\langle \tilde{\overleftrightarrow{Z}}_t(\vec{r}_s, \vec{r}\,'_s; s);\ \vec{\tilde{J}}_s(\vec{r}\,'_s, s)\right\rangle = \vec{\tilde{E}}^{(inc)}(\vec{r}_s, s) \tag{3.5}$$

with the subscript t referring to tangential components. Here a symmetric kernel appropriate to the impedance (or E-field) integral equation is used for convenience. The formal solution is

$$\vec{\tilde{J}}_s(\vec{r}_s, s) = \left\langle \tilde{\overleftrightarrow{Z}}_t^{-1}(\vec{r}_s, \vec{r}'_s; s);\ \vec{\tilde{E}}^{(inc)}(\vec{r}'_s, s) \right\rangle$$

$$= E_0 \tilde{f}(s) \vec{1}_p \cdot \left\langle \tilde{\overleftrightarrow{Z}}_t^{-1}(\vec{r}_s, \vec{r}'_s; s), e^{-\gamma \vec{1}_i \cdot \vec{r}'} \right\rangle \tag{3.6}$$

If we have a transfer function with an SEM expansion as

$$\tilde{T}_{n,m}(s) = \sum_{\alpha_{in}} T_{n,m}^{(\alpha_{in})} \left[s - s_{\alpha_{in}} \right]^{-1} \tag{3.7}$$

then we can write for the port voltage or effective height

$$\vec{C}_{\alpha_{in}}^{(V_n)}(\vec{1}_i) = T_{n,m}^{(\alpha_{in})} \vec{1}_m \cdot \left\langle \overleftrightarrow{Z}_t^{-1}(\vec{r}_{s_m}, \vec{r}'_s; s_{\alpha_{in}}), e^{-\gamma_{\alpha_{in}} \vec{1}_i \cdot \vec{r}'} \right\rangle \tag{3.8}$$

for surface-current-density penetrations as in (3.2). For surface-charge-density penetrations as in (3.3) we have

$$\vec{C}_{\alpha_{in}}^{(V_n)}(\vec{1}_i) = T_{n,m}^{(\alpha_{in})} \left[-\frac{1}{s_{\alpha_{in}}} \nabla_s \cdot \left\langle \tilde{\overleftrightarrow{Z}}_t^{-1}(\vec{r}_s, \vec{r}'_s; s_{\alpha_{in}}); e^{-\gamma \vec{1}_i \cdot \vec{r}'_s} \right\rangle \right]_{\vec{r}_s = \vec{r}_{s_m}} \tag{3.9}$$

From an experimental point of view for a given $\vec{1}_i$ one can choose two orthogonal polarizations as $\vec{1}_2$ and $\vec{1}_3$ (used in (2.10)) and measure the surface current density with both incident polarizations to form

$$\vec{C}_{\alpha_{in}}^{(V_n)}(\vec{1}_i) = T_{n,m}^{(\alpha_{in})} \frac{1}{E_0 \tilde{f}(s)} \left[\vec{1}_2(\vec{1}_i) \vec{\tilde{J}}_s(\vec{r}_{s_m}, s_{\alpha_{in}}, \vec{1}_2) + \vec{1}_3(\vec{1}_i) \vec{\tilde{J}}_s(\vec{r}_{s_{in}}, s_{\alpha_{in}}, \vec{1}_3) \right] \cdot \vec{1}_m \tag{3.10}$$

for surface-current density penetrations. Similarly for surface-charge-density penetrations we have

$$\vec{C}^{(V_n)}_{\alpha_{in}}(\vec{1}_i) = -T^{(\alpha_{in})}_{n,m} \frac{1}{E_o\tilde{f}(s)} \frac{1}{s_{\alpha_{in}}} \Big[\vec{1}_2(\vec{1}_i)\nabla_s \cdot \vec{\tilde{J}}_s(\vec{r}_s, s_{\alpha_{in}}, \vec{1}_2) + \vec{1}_3(\vec{1}_i)\nabla_s \cdot \vec{\tilde{J}}_s(\vec{r}_s, s_{\alpha_{in}}, \vec{1}_3) \Big]_{\vec{r}_s = \vec{r}_{s_m}} \tag{3.11}$$

By measurements of the surface current and charge densities at $s_{\alpha_{in}}$ and the pole residue for the appropriate transfer function then one has $\vec{C}^{(V_n)}_{\alpha_{in}}(\vec{1}_i)$. However, unless this angular dependence from a given $\vec{r}_{s_m}$ applies to more than one $\tilde{V}_n$, it may be more convenient to measure this density at the mth port as discussed in Section II.

Now, if more than one $\tilde{T}_{n,m}$ for a given n has the same $s_{\alpha_{in}}$, then (3.8) through (3.11) are readily generalized by a linear combination, summing over the applicable m. This requires measurement for the various positions and modes of penetration designated by $\vec{r}_{s_m}$. Again it may be more convenient to measure the coupling vector at the nth port in reception or transmission.

IV. Concluding Remarks

An efficient way to display the important interaction of incoming plane waves with complex electronic systems is then to display the coupling vectors (two components perpendicular to $\vec{1}_i$) as a function of $\vec{1}_i$ (in general over 4π steradians, or $0 \leq \theta \leq \pi$ and $0 \leq \phi \leq 2\pi$ in the usual spherical coordinates). For cases that allow an approximate division of the modes into exterior and interior ones, measurement (or calculation) of the coupling vectors for the external surface current density can be used to within a scaling constant. For the interior modes the situation is in general not as simple.

The response is conveniently considered in terms of an effective height which dot multiplies the incident field. One can measure this by exposure to an incident field, or using reciprocity [14], one can drive internal ports of interest with a known current source and measure the far field (or measure the near field and process it to obtain the far field). The coupling vectors for the natural frequencies apply equally well to both descriptions.

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