Sensor and Simulation Notes

Note 312

4 October 1988

Maximization of Electromagnetic Response at a Distance

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Abstract

There are various factors which affect the electromagnetic response of a system which is spaced away from the electromagnetic illuminator. Resonant waveforms (approximate damped sinusoids) are used for maximum coupling. The various factors are then the source amplitude and pulse width, and the transfer functions for the antenna, propagation, response of the system exterior, and response of the system interior. These are combined to optimize the overall response.

electromagnetic coupling

I. Introduction

In designing an electromagnetic illuminator which is to excite a system at some distance away from the source (compared to wavelength and illuminator antenna dimensions) there are various factors to be considered. As indicated in Figure 1.1 let us break the problem down as the product of a source amplitude V_o' times transfer functions for a radiating antenna (T_a), a propagation path (T_b), the response of the system outer surface (T_o), and the system interior (T_i). This product gives the response V_i (say a voltage) at some interior port of interest.

As shown in [3] maximum response is normally achieved by the use of highly resonant exciting waveforms (such as damped sinusoids). This allows quasi-CW concepts to be used so that the transfer functions can be considered functions of the dominant frequency of the exciting waveforms.

Subsequent sections discuss these factors. They are then recombined to estimate the system response and the conditions for maximizing the system response are estimated.

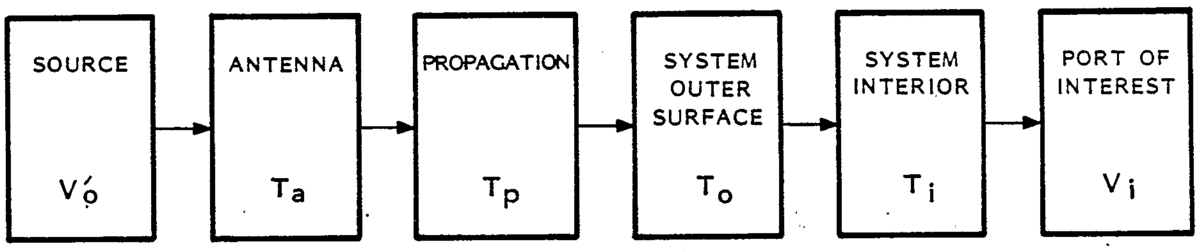

Figure 1.1. Factorization of Transfer Function from Source to System

II. Source

Let us characterize the source by some waveform $f_s(t)$ which is normalized to have approximately a unit peak. In terms of a voltage we might have

$$V_s(t) = V_o\, f_s(t)$$

$$\tilde{V}_s(s) = V_o\, \tilde{f}_s(s) \text{ (Laplace transform)} \tag{2.1}$$

Now not all sources are characterized by a simple voltage. For example, in a non-TEM waveguide mode one might have this term represent some kind of modal coefficient [9]. One can also associate a current as

$$\tilde{I}_s(s) = \tilde{Y}_s(s)\, \tilde{V}(s)$$

$$\tilde{Y}_s(s) \equiv \text{source admittance}$$

$$\tilde{Z}_s(s) \equiv \tilde{Y}_s^{-1}(s) \equiv \text{source impedance} \tag{2.2}$$

where this can also be some kind of modal coefficient if appropriate and the admittance can be a modal admittance. In any event the voltage is related to the electric field, the current to the magnetic field, and the admittance (impedance) to the ratio (in complex-frequency domain, not time domain).

The foregoing discussion is most appropriate at the output of the source since in general in the source something fundamentally nonlinear is usually going on. The source is assumed to be matched to a load (the antenna system) in such a way that the admittance is a meaningful parameter, such as describing a ratio of electric and magnetic parameters in a TEM or waveguide mode. Note also that there may be some dispersion due to a frequency dependence of some modal propagation speed. Assuming an $\tilde{f}_s(s)$ with a concentration in a small frequency band this can be made a small effect.

One can also characterize the source in terms of power

$$P_s(t) = V_s(t)\ I_s(t) = V_s(t)\ [Y_s(t) \circ V_s(t)]$$

$$= I_s(t)\ [Z_s(t) \circ I_s(t)]$$

$\circ \equiv$ convolution with respect to time

$$\tilde{p}_s(s) \equiv \tilde{V}_s(s)\ \tilde{I}_s(-s) \tag{2.3}$$

Note that the $\tilde{p}_s(s)$ parameter is not the Laplace transform of $P_s(t)$. They are, however, related by the generalized parseval theorem [5].

$$W_s \equiv \int_{-\infty}^{\infty} P_s(t)\ dt = \frac{1}{2\pi j} \int_{Br} \tilde{p}_s(s)\ ds = \frac{1}{2\pi j} \int_{Br} \tilde{p}_s(-s)\ ds$$

$$= \frac{1}{2\pi} \int_{-\infty}^{\infty} \mathrm{Re}[\tilde{p}_s(j\omega)]\ d\omega = \frac{1}{\pi} \int_{0}^{\infty} \mathrm{Re}[p_s(j\omega)]\ d\omega \tag{2.4}$$

This can also be expressed in terms of energy norms [5] as

$$W_s^{1/2} = ||V_s(t),\ Y_s(t) \circ\ ||_e = ||I_s(t),\ Z_s(t) \circ\ ||_e$$

$$= \frac{1}{\sqrt{2\pi}}\ ||\tilde{V}_s(j\omega),\ \tilde{Y}_s(j\omega)||_e = \frac{1}{\sqrt{2\pi}}\ ||\tilde{I}_s(j\omega),\ \tilde{Z}_s(j\omega)||_e \tag{2.5}$$

If $\tilde{Z}_s(s)$ is a constant, say R_s, this simplifies to

$$W_s^{1/2} = R_s^{-1/2}\ ||V_s(t)||_2 = R_s^{1/2}\ ||I_s(t)||_2$$

$$= \frac{1}{\sqrt{2\pi R_s}}\ ||\tilde{V}_s(j\omega)||_2 = \sqrt{\frac{R_s}{2\pi}}\ ||\tilde{I}_s(j\omega)||_2 \tag{2.6}$$

As discussed in [3] coupling to systems is maximized by oscillatory waveforms tuned to resonances in transfer functions. So let us choose as our canonical source waveform

$$f_s(t) = \frac{1}{2}\left[e^{s_s t + j\nu_s} + e^{s_s^* t - j\nu_s}\right] u(t)$$

$$\tilde{f}_s(s) = \frac{1}{2}\left[\frac{e^{j\nu_s}}{s - s_s} + \frac{e^{-j\nu_s}}{s_s - s_s^*}\right]$$

$$\mathrm{Im}[\nu_s] = 0,\ \Omega_s = \mathrm{Re}[s_s] \leq 0,\ \omega_s = \mathrm{Im}[s_s] > 0 \qquad (2.7)$$

Here ν_s is merely a phase parameter. In another form we have

$$f_s(t) = e^{\Omega_s t} \cos(\omega_s t + \nu_s)\ u(t) \qquad (2.8)$$

so that for $|\Omega_s| << \omega_s$ (a highly resonant waveform) the peak is approximately 1. In norm form we have [4]

$$||f_s(t)||_\infty \simeq 1$$

$$||f_s(t)||_r = 1 \qquad (2.9)$$

From [3] we also have

$$||f_s(t)||_2 = \left\{-\mathrm{Re}\left[\frac{e^{j2\nu_s}}{4s_s}\right] - \frac{1}{4\ \mathrm{Re}[s_s]}\right\}^{\frac{1}{2}} \qquad (2.10)$$

which for a highly resonant waveform is

$$||f_s(t)||_2 \simeq \frac{1}{2}\left\{-\mathrm{Re}[s_s]\right\}^{-\frac{1}{2}} = \frac{1}{2}\left\{-\ \Omega_s\right\}^{-\frac{1}{2}} \qquad (2.11)$$

This result can also be extended to arbitrary p-norms as in [4]. Note that all these norms are extended to $V_s(t)$ merely by multiplication by V_o.

For a highly oscillatory waveform as in (2.7) we can use frequency-domain concepts for approximations in time domain. The peak power is

$$P_o \simeq V_o^2 \, Y_{s_o}$$

$$Y_{s_o} \equiv \mathrm{Re}[\tilde{Y}_s(s_s)] = \mathrm{Re}[\tilde{Y}_s(j\omega_s)]$$

$$s_s \simeq j\omega_s \tag{2.12}$$

One can also think of an average power as a function of time by averaging over a period of the wave; this "average" power has a peak that is $P_o/2$.

Recognizing that the source amplitude involves the admittance, and that in principle one can change this admittance via a transformer, let us define an equivalent peak voltage as

$$V_o' \equiv V_o \sqrt{Y_{s_o} \, Z_o} \; , \; P_o = V_o'^{\,2} \, Z_o^{-1} \tag{2.13}$$

This represents the source peak amplitude for later scaling purposes. Combined with the decay constant $-\Omega_s$ (or pulse width, if you prefer) we have the approximate characterization of the source, including the various norms. One can also use the above to define an equivalent voltage waveform as

$$V_s'(t) = V_o' \, f_s(t) \tag{2.14}$$

This equivalent voltage is defined as though it is driving Z_o, the impedance of free space.

Now varying ω_s and Ω_s one can imagine various possible sources. Beginning with say 60 Hz (50 Hz in Europe) power plants operating continuously at many MW one can go on up in frequency through such things as VLF, MF, HF, etc. transmitters. As one goes higher in frequency so that transmitting antennas can be made highly directional, the fields illuminating a system of interest can be increased by increasing ω_s. For coupling to the system, as will be discussed later, it is desirable to have wavelengths of the order of system dimensions or less (but not too small). This emphasizes the HF band and somewhat higher ω_s.

Recent years have seen considerable work on sources for these high frequencies [6,7,12]. Investigators have obtained a few GW at a few GHz with pulse widths of a few 10s of ns [8]. As one goes even higher in frequency powers achieved fall off. So let us consider V_o' as a function of s_s (or ω_s and Ω_s). In particular in the frequency range of interest V_o' decreases for increasing ω_s, at least in the current state of the art. For reference purposes a few GW gives a V_o' of about a MV.

III. Antenna

Taking the signal from the source we next need an antenna to radiate it in some optimal manner. Assuming an antenna size that is many wavelengths one can radiate the signal in a highly directive manner. This is standardly done by a reflector antenna (such as parabolic and Cassagrain) [13]. Besides the directivity requirement one should also note the high-power requirement for which electric fields may be near breakdown.

For our purposes a simple way to think of a reflector antenna is as an aperture antenna. Maximizing the fields at the system of interest leads to a focused aperture antenna [1]. In this case the fields incident on the system are approximately (assuming linear polarization in some direction $\vec{1}_p$ perpendicular to the direction of propagation $\vec{1}_1$) in the far field

$$\tilde{\vec{E}}^{(inc)}(\vec{r}_o,s) = \vec{1}_p\, E_o\, e^{-\gamma R_o}\, \tilde{f}(s)\, \cos(\theta_o)\, \frac{\gamma A}{2\pi\, R_o}$$

$R_o \equiv$ distance from antenna to system

$$\gamma = \frac{s}{c} \tag{3.1}$$

θ_o = angle away from normal to aperture

$A \equiv$ aperture area

Converting this to retarded time and $\theta_o = 0$ gives

$$\tilde{E}^{(inc)}(R_o,s) = E_o\, \tilde{f}(s)\, \frac{\gamma A}{2\pi\, R_o} \tag{3.2}$$

where now

$E_o\tilde{f}(s) \simeq$ tangential electric field on aperture

$E_o \simeq$ peak tangential electric field on aperture (3.3)

Assuming that

$$\tilde{f}(s) \simeq \tilde{f}_s(s) \tag{3.4}$$

then we have

$$\tilde{E}^{(inc)}(R_o,s) \simeq E_o \frac{s_s A}{2\pi c R_o} \tilde{f}_s(s) \tag{3.5}$$

Using (2.7), assuming a highly resonant waveform so that frequencies of importance are concentrated near s_s and s_s^*, we have

$$s_s \tilde{f}_s(s) \simeq \frac{1}{2}\left[\frac{s_s e^{j\nu_s}}{s-s_s} + \frac{s_s^* e^{-j\nu_s}}{s-s_s^*}\right]$$

$$\simeq \frac{1}{2}\left[\frac{j\omega_s e^{j\nu_s}}{s-s_s} - \frac{j\omega_s e^{-j\nu_s}}{s-s_s}\right] \tag{3.6}$$

Defining

$$\nu_f \equiv \frac{\pi}{2} + \nu_s \tag{3.7}$$

we then have a waveform in the far field

$$\tilde{f}_f(s) = \frac{1}{2}\left[\frac{e^{j\nu_f}}{s-s_s} + \frac{e^{-j\nu_f}}{s-s_s^*}\right]$$

$$f_f(t) = \frac{1}{2}\left[e^{s_s t+j\nu_f} + e^{s_s^* t-j\nu_f}\right] \tag{3.8}$$

so that (in retarded time)

$$E^{(inc)}(R_o,t) = E_o \frac{\omega_s A}{2\pi c R_o} f_f(t) \tag{3.9}$$

Here f_f has peak value approximately 1, and

$$E_f = E_o \frac{\omega_s A}{2\pi c R_o} = E_o \frac{f_s A}{c R_o} = E_o \frac{A}{\lambda_s R_o} \qquad (3.10)$$

represents the peak far field in agreement with the usual aperture-antenna theory. Actually this is accurate for focused apertures as long as R_o is large compared to both the aperture radius and λ_s (wavelength) [1]. Note the phase shift of $\pi/2$ in (3.7) introduced by the antenna.

Note that it is important to restrict frequencies to near ω_s (or f_s) because a reflector such as this is a differentiator as in [1] which can give a narrow early-time spike in the far field. We concentrate on the frequencies near s_s for optimum coupling.

Not only are there limitations on say $V_o'(\omega_s, \Omega_s)$ at the source, the antenna also has limitations related to electrical breakdown around the antenna. In sea-level air electric fields are limited to a few MV/m. For very narrow pulse widths this can be increased slightly, but this is not in general advantageous since we need $|\Omega_s|$ small enough that the response at ω_s can "ring up", the time for this being related to the width of the resonance in the system of interest.

The peak power radiated by the antenna is

$$P_a \simeq E_o^2 \frac{A}{Z_o} \qquad (3.11)$$

If E_o is set a little below breakdown field then P_a is of the order $10^{10} A$ (in Watts). So large peak powers imply large antennas. This is especially important for the feeds where special insulation will be useful. So there are limitations in power implied by both source and antenna.

Assuming negligible losses we may equate the power from the source to that radiated by the antenna via (2.13) and (3.11) as

$$V_o' = E_o A^{1/2} \qquad (3.12)$$

The equivalent voltage of the source is then related to the electric field on the antenna times a length, i.e., $A^{1/2}$.

The antenna transfer function to the system is then

$$T_a \equiv \frac{E_f}{V_o'} = \frac{E_o A}{V_o' \lambda_s R_o} = \frac{A^{1/2}}{\lambda_s R_o} = \frac{f_s A^{1/2}}{c R_o} = \frac{\omega_s A^{1/2}}{2\pi c R_o} \tag{3.13}$$

with dimensions $(\text{meter})^{-1}$. This is what the antenna really does.

The basic lesson of (3.13) is that large ω_s is good, particularly if A and E_o are fixed. There are limits to this, related to mechanical tolerances in antenna dimensions as one goes down to mm wavelengths. One must trade off source power (decreasing as ω_s increases) versus antenna gain (increasing as ω_s increases). There is also the limitation that A must be large enough (including feed dimensions and insulation to allow the power out of the antenna region).

IV. Propagation

Another factor is the possible attenuation of the signal between antenna and system. For example, if the propagation path goes towards higher altitudes where electric-field breakdown is smaller one may wish to introduce a separate attenuation, above the geometrical one in Section 2, to account for this.

Yet another possible propagation factor is dispersion, such as introduced by the ionosphere. In this case there is a frequency-dependent delay. To the extent that the frequencies in the pulse are narrowly concentrated in (3.8) (by a highly resonant waveform), this effect is reduced.

One can assign a transfer function to this propagation allowing for any attenuation (beyond the antenna factor in Section 3) and dispersion, say T_p. Strictly, this is a function of complex frequency. However, for the highly resonant pulse in (3.8) we can consider this a function of ω_s, provided there is not too much dispersion (such as one might experience at certain frequencies in the ionosphere). For present purposes we take

$$T_p \simeq 1 \tag{4.1}$$

It should also be noted that breakdown electric field strength can vary along the propagation path. This could be associated, for example, with variations in air pressure. So there may also be limitations on the signal imposed by nonlinear characteristics of the propagation path.

V. Interaction with System Exterior

In analyzing the interaction of an electromagnetic wave with a complex system, it is often possible to identify some outer surface S_o of the system which serves as a crude conducting shield. The first-order way to look at this concerns the general subject of electromagnetic topology [2,11]. The first step in this approach is to find the exterior response of the system. Most important, this involves finding the short-circuit surface current and charge densities on S_o, these being the parameters that drive small antennas and apertures. There are cases in which there are large appendages (such as power or communication lines), for which the analysis can be generalized to include appropriate equivalent circuits of these from the external interaction problem.

In analyzing the external interaction there are three basic frequency ranges of concern [11]. For wavelengths large compared to the largest characteristic dimension (say ℓ) of the scatterer we have the quasistatic regime. In this regime the surface current density is proportional to the incident magnetic field, and the surface charge density is proportional to the incident electric field. These relationships are frequency independent, but depend on geometry, position of observation, and polarization of the incident electric and magnetic fields. The surface fields can be some factor times the incident fields which can easily be like an order of magnitude (10 times) higher or lower, with even more extreme variation possible.

In the resonance regime the basic concept for analysis is the singularity expansion method (SEM) [10]. The natural frequencies (in the left half s plane) of the exterior represent damped sinusoids in time domain or poles in complex frequency plane. This frequency region extends from a low frequency where ℓ is like a half wavelength to somewhat higher frequencies, both in the sense of multiple half-wavelength resonances and in the sense of the half or quarter wavelength corresponding to various protrusions on (or depressions in) S_o.

In the high-frequency (or "optical") regime the basic concept for analysis is the geometrical theory of diffraction (GTD) and its variants. For our purposes we can use the first term, often referred to as "physical optics", at least on the illuminated side of the system [11]. For our purposes the surface fields are just twice the relevant components of the incident fields. Near edges (and exterior points or corners) these can be even larger. On the shadow side a more detailed analysis can estimate the fields, but in general the fields are smaller and less important there.

Summarizing, the exterior transfer function has

$$T_o \simeq 1 \text{ (order of magnitude sense)} \tag{5.1}$$

except in the resonance region where it can have significant peaks (which could be matched to ω_s in (3.8) to increase response). For low frequencies there is a variation of this over the body. For high frequencies this applies primarily to the illuminated side of the body.

VI. Interaction with System Interior

Going the next step in the topological decomposition of the system response let us consider the internal interaction. Here we lump this as a transfer function from the surface current and charge densities on the exterior of S_o to some interior port of interest producing voltage and current waveforms there. Let V_i designate some voltage there.

At low frequencies apertures and small antennas are usually differentiators [11]. Consider a small aperture with a wire behind it. This is usually modeled as a transmission line with a series voltage source proportional to the exterior surface B-dot and a transverse current source proportional to the exterior surface D-dot. Provided the wire is terminated in finite, non-zero load impedances (at low frequencies) then the transfer function T_i in this low-frequency range is proportional to ω. Similar comments apply to electrically small antennas on S_o; the low-frequency behavior is usually proportional to ω (or an even higher power of ω if special designs are used to further reduce low-frequency response).

As frequency is increased interior resonances appear. These can be associated with internal wires which may be as long as ℓ (the largest characteristic dimension of the object) or perhaps somewhat longer. The first resonance may then occur at ω_ℓ where this is half-wave or even quarter-wave resonant. Various lumped-elements on these wires can even lower the first resonance. As one goes up in frequency shorter wires become important. Other features such as cavities and apertures become important [11]. Of some interest are the smallest dimensions important for resonances. These might include things like small apertures (such as windows) in S_o, or even the dimensions of boxes. In any event this establishes some high resonant frequency ω_h of interest. Between ω_ℓ and ω_h the transfer function T_i exhibits resonant behavior with order-of-magnitude variations as a function of frequency.

Above ω_h we can estimate the transfer function a different way. Considering the various wires carrying electrical signals between the various boxes, let us note that for wavelengths less than the distance of these wires from a ground plane (provided by local structure) the signals induced are proportional to ω^{-1}, i.e., the wires act as integrators [11]. At these high frequencies apertures in S_o let in fields in a roughly frequency-independent manner, so the transfer function T_i of exterior fields to wire signals (voltages and currents) should go roughly like ω^{-1}. This neglects things like special filters at the box inputs. One might estimate ω_h based on the smallest typical resonant dimensions as of the rough order of a GHz.

So now we can define a canonical system response as illustrated in Figure 6.1. Here we combine T_o from Section 5 with T_i to form a composite transfer function from incident fields to response at box inputs. This is a canonical transfer function in that it applies to typical systems, but notes there can be exceptions for special system features.

Divide the frequency spectrum as:

Band 1: $f_s \lesssim f_\ell$ (aperture and small antenna coupling region)

Band 2: $f_\ell \lesssim f_s \lesssim f_h$ (resonance region, external and internal)

Band 3: $f_h \lesssim f_s$ (integration region) (6.1)

In these three bands we have

Band 1: $T_o T_i$ = constant x ω_s

Band 2: $T_o T_i$ = oscillatory function of ω_s (peaks and valleys)

Band 3: $T_o T_i$ = constant x ω_s^{-1} (6.2)

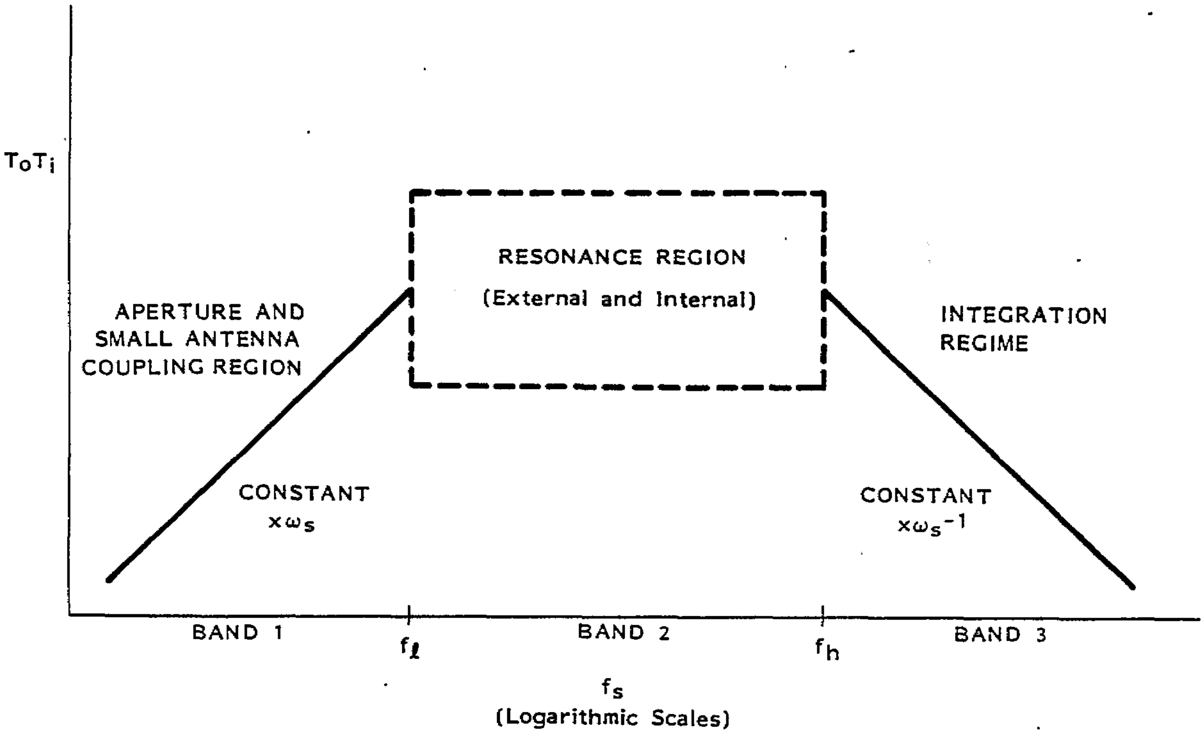

Figure 6.1 Canonical System Response as a Function of Frequency

VII. Overall Response

Combining the canonical system response in Figure 6.1 with the propagation response transfer function (4.1), the antenna transfer function (3.13), and the source V_o' in (2.14) gives an overall response as

$$V_i = V_o' \, T_a \, T_p \, T_o \, T_i \tag{7.1}$$

where V_i is the response at a box input. Neglecting any variation in T_p it is the combination of $V_o' T_a$ from (3.12) and (3.13) which gives

$$E_f = V_o' \, T_a = V_o' \frac{\omega_s \, A^{1/2}}{2\pi c \, R_o} = V_o' \frac{f_s \, A^{1/2}}{c \, R_o} \tag{7.2}$$

Together with (7.1) this is

$$V_i \simeq V_o' \frac{\omega_s \, A^{1/2}}{2\pi c \, R_o} T_o \, T_i = V_o' \frac{f_s \, A^{1/2}}{c \, R_o} T_o \, T_i \tag{7.3}$$

which is appropriate if the basic limitation is the equivalent source voltage V_o'. If the basic limitation is the electric field E_o on the antenna we have

$$V_i \simeq E_o \frac{\omega_s \, A}{2\pi c \, R_o} T_o \, T_i = E_o \frac{f_s \, A}{c \, R_o} T_o \, T_i \tag{7.4}$$

For the case of antenna limitation as in (7.4) clearly larger V_i is produced by larger f_s, except that from Figure 6.1 nothing is gained by making f_s larger than f_h. For the case of source limitation as in (7.3) we need to estimate how rapidly V_o' falls off as a function of f_s. If it falls off like f_s^{-1} then from Figure 6.1 an optimal choice of f_s lies between f_ℓ and f_h. In any event f_h is about the highest frequency of interest and this is expected to be around a GHz. Looking at both (7.3) and (7.4) we can see that V_i is also increased by increasing A.

It is not only the peak amplitude V_i of the voltage that is significant. The width of the resonant pulse is also significant. First of all $|\Omega_s|$ must be smaller than the width of the system resonance to which ω_s is tuned, ie., some particular resonance, say with f_s near f_h. This maximizes the peak V_i, or in norm terms maximizes $||V_i(t)||_\infty$. The ∞-norm or peak may not be the only norm of interest. As in [4] the p-norm of a resonant waveform is proportional to the peak times $|\Omega_s|^{-\frac{1}{p}}$ where S_s is assumed to characterize the response (provided $|\Omega_s|$ is small enough). The 2-norm is often used if energy into the port is important. Then pulse width is also important, but of course this requires more energy from the source. In any event the pulse width should be large enough for the signal to ring up ($|\Omega_s|$ smaller than width of system resonance).

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