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EFFECT OF LOCATION ON THE EFFECTIVENESS OF SPATIAL FILTERS.(U)

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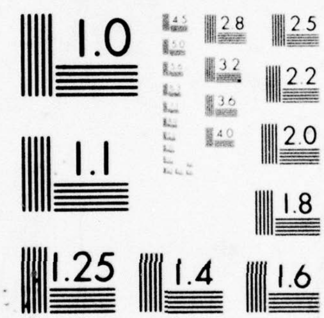

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SEPTEMBER 1976

Effect of Location on the Effectiveness of Spatial Filters

RONALD L. FANTE

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ROME AIR DEVELOPMENT CENTER
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We have demonstrated that the radiation pattern modification caused by the placement of either a multiple-dielectric-layer or a wire-grade spatial filter between an antenna and observer is independent of where the filter is placed, provided the filter is not located too close to the antenna to cause a significant change in antenna current distribution.

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1. INTRODUCTION

It has been demonstrated previously by Mailloux[1] that spatial filters can be useful in eliminating unwanted sidelobes in antenna arrays. These spatial filters may be composed either of multiple plane dielectric layers or wire grids. The analysis performed by Mailloux assumed that the spatial filter was located in the Fraunhofer zone of the source. We will show that the aforementioned results are also valid if the spatial filter is located in the Fresnel zone of the source antenna, provided the spatial filter is not too close to modify the source current distribution.*

Our analysis will be based on decomposing the electromagnetic field of an arbitrary source into an ensemble of propagating and evanescent plane waves, and then seeing how each of these waves is affected by the spatial filter. We will assume that the source field is specified either over the source antenna surface or over a virtual plane just in front of the source.

(Received for publication 29 September 1976)

*This is equivalent to stating that multiple scatter is ignored. That is, we assume that any wave reflected by the spatial filter is not scattered by the antenna and reradiated.

1. Mailloux, R. (1976) Synthesis of spatial filters with Chebyshev characteristics, IEEE Trans. Ant. and Prop. AP-24:174-181.

2. REPRESENTATION OF THE SOURCE

Let us consider a source in the plane $z = -\Delta$, as indicated in Figure 1. The electric field of this source, assuming time dependence $\exp(-i\omega t)$, in the region $z \geq -\Delta$ can be represented quite generally as

$$\underline{E}(x, y, z) = \left(\frac{1}{2\pi}\right)^2 \iint_{-\infty}^{\infty} d\alpha \, d\beta \, \hat{\underline{E}}(\alpha, \beta) \exp[ik(\alpha x + \beta y + hz)] \quad , \tag{1}$$

where $h = (1 - \alpha^2 - \beta^2)^{1/2} = i(\alpha^2 + \beta^2 - 1)^{1/2}$, and $k = 2\pi/\lambda$ where λ is the signal wavelength. It can be shown[2] that $\hat{E}$ can be written in the general form

$$\hat{\underline{E}}(\alpha, \beta) = \hat{x} A_x(\alpha, \beta) + \hat{y} A_y(\alpha, \beta) + \hat{z}\left[\frac{\alpha A_x + \beta A_y}{h}\right] , \tag{2}$$

where $\hat{x}$, $\hat{y}$ and $\hat{z}$ are unit vectors. From Eq. (2) we see that once the x and y components of $\hat{\underline{E}}$ are specified, the z component automatically follows. The complete magnetic field is also automatically specified once A_x and A_y are known.

Now suppose the transverse electric field in the plane $z = -\Delta$ is specified to be the vector function $\underline{f}(x, y) = \hat{x} f_x(x, y) + \hat{y} f_y(x, y)$. Then the vector $\underline{A}(\alpha, \beta) = \hat{x} A_x(\alpha, \beta) + \hat{y} A_y(\alpha, \beta)$ is readily found via

$$\begin{aligned} \underline{A}(\alpha, \beta) &= k^2 e^{ikh\Delta} \iint_{-\infty}^{\infty} dx \, dy \, \underline{f}(x, y) \exp[-ik(\alpha x + \beta y)] \\ &= e^{ikh\Delta} \underline{A}_o(\alpha, \beta) \end{aligned} \tag{3}$$

where $\underline{A}_o(\alpha, \beta)$ is the wavenumber spectrum of the transverse electric field which would be obtained if the source were located at $z = 0$, rather than at $z = -\Delta$. By using Eq. (3) in (2), we easily see that

$$\hat{\underline{E}}(\alpha, \beta) = \exp(ikh\Delta) \, \hat{\underline{E}}_o(\alpha, \beta) \tag{4}$$

where $\hat{\underline{E}}_o(\alpha, \beta)$ is the wavenumber spectrum of the complete electric field in the case when the source is located at $z = 0$, rather than at $z = -\Delta$.

2. Fante, R., and Mayhan, J. (1970) Bounds on the electric field outside a radiating system, IEEE Trans. Ant. and Prop. AP-18:64-68.

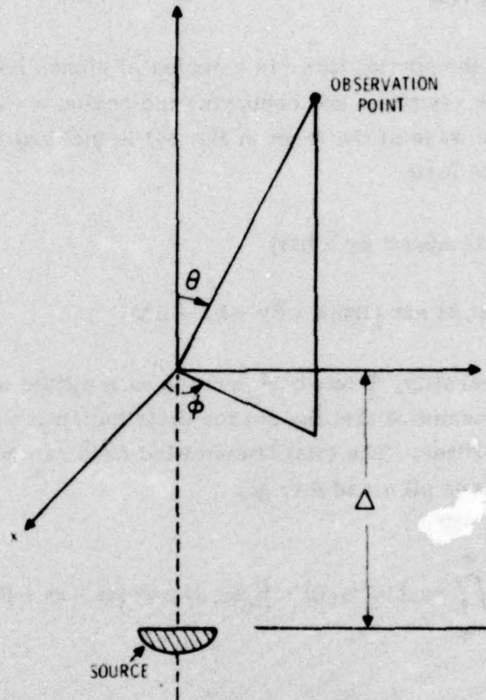

Figure 1. Geometry Assumed

Equation (1) represents the decomposition of the electric field into plane waves of the form

$$\underline{\hat{E}}(\alpha, \beta) \exp\left[ik(\alpha x + \beta y + hz)\right] \quad . \tag{5}$$

When $\alpha^2 + \beta^2 \le 1$ these waves are propagating, whereas when $\alpha^2 + \beta^2 > 1$ they are evanescent. We can now investigate how each of these waves is transmitted by a spatial filter located in the plane $z = 0$. We shall first consider the simple and obvious case of a filter consisting of multiple plane dielectric layers. We will later consider the more complex case when the filter is a wire grid.

3. DIELECTRIC LAYER FILTER

For the case when the spatial filter is a series of plane, homogeneous dielectric* layers lying in the x-y plane and occupying the region $z = 0$ to $z = d$, it is readily shown[3] that if a wave of the form in Eq. (5) is incident upon this layer, the transmitted field has the form

$$\underline{\underline{T}}(\alpha,\beta)\cdot\underline{\hat{E}}(\alpha,\beta)\exp[ik(\alpha x+\beta y+hz)]$$

$$=\underline{\underline{T}}(\alpha,\beta)\cdot\underline{\hat{E}}_o(\alpha,\beta)\exp[ik(\alpha x+\beta y+hz+h\Delta)]\quad, \tag{6}$$

where for complete generality, $\underline{\underline{T}}$ has been written as a dyadic quantity. In deriving Eq. (6) it has been assumed that the source distribution is unaffected by the presence of the spatial filter. The total transmitted field can be obtained from Eq. (6) by integrating over all α and β to get

$$\underline{E}(x,y,z)=\left(\frac{1}{2\pi}\right)^2\iint_{-\infty}^{\infty}d\alpha\, d\beta\, \underline{\underline{T}}(\alpha,\beta)\cdot\underline{\hat{E}}_o(\alpha,\beta)\exp\Big(ik\,[\alpha x+\beta y+h(z+\Delta)]\Big)\,. \tag{7}$$

In the far field ($z \to \infty$) it is easily shown that[2]

$$\underline{E}(x,y,z\to\infty)=\frac{K\cos\theta}{R}\,\underline{\underline{T}}(\alpha_o,\beta_o)\cdot\underline{\hat{E}}_o(\alpha_o,\beta_o)\exp[ik(R+\Delta\cos\theta)] \tag{8}$$

where K is a constant, $\alpha_o = \sin\theta\cos\phi$, $\beta_o = \sin\theta\sin\phi$, $R = (x^2+y^2+z^2)^{1/2}$ and θ and ϕ are shown on Figure 1. If the source were at $z = 0$ instead of at $z = -\Delta$, the far field would be

$$E(x,y,z\to\infty)=\frac{K}{R}\cos\theta\,\underline{\underline{T}}(\alpha_o,\beta_o)\cdot\hat{E}_o(\alpha_o,\beta_o)\exp(ikR)\quad. \tag{9}$$

Upon comparing Eqs. (8) and (9) we see that, in the Fraunhofer zone, the only effect of the distance Δ between the source and the spatial filter is the introduction of the phase factor $\exp(ik\Delta\cos\theta)$, which is of no consequence. The reason that this result is so simple is because the evanescent modes in the source play no role in the far field. We shall see in the next section that this is no longer the case for a wire-grid filter.

*It is assumed that each layer is homogeneous and infinite in the x-y direction. The filter is, of course, not necessarily homogeneous in the z-direction.

3. Stratton, J. (1941) Electromagnetic Theory, Chapter 9, McGraw-Hill, New York.

4. WIRE-GRID FILTER

A rectangular grid of metal wires can also be used as a spatial filter. Consider the wire grid shown in Figure 2, and let us suppose that this lies in the $z = 0$ plane, with the source again located in the plane $z = -\Delta$. We will also assume that the wire separations, a and b, are much smaller than the signal wavelength λ. It can readily be shown[4] that, if a wave of the form given in Eq. (5) is incident upon this grid, the transmitted field in the region $z > 0$ has the general form

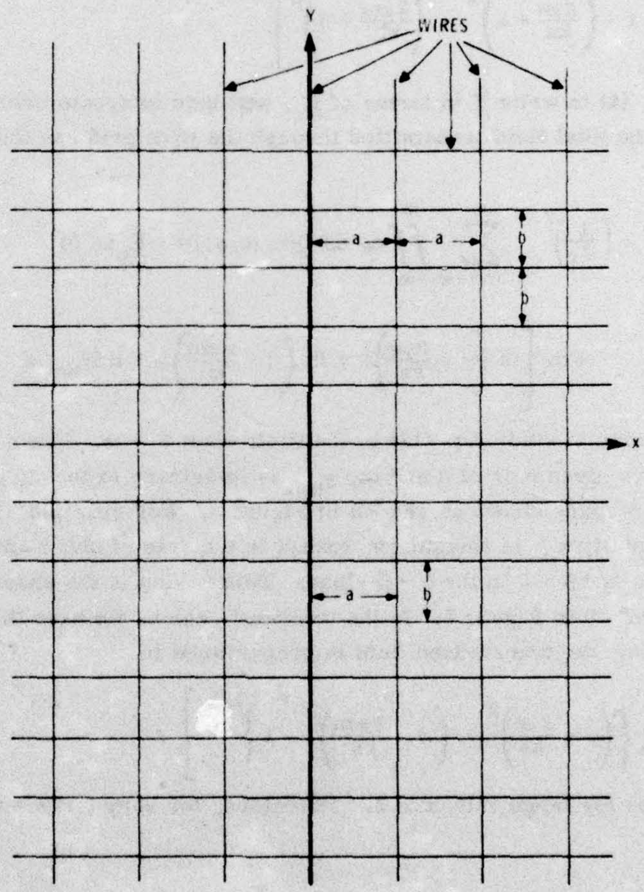

Figure 2. Wire-Grid Orientation

4. Hill, D., and Wait, J. (1976) Electromagnetic scattering of an arbitrary plane wave by a wire mesh with bonded junctions, Canadian J. Phys. 54:353-361.

$$\sum_{n,\,m=-\infty}^{\infty} \underset{\approx}{Q}(n, m, \alpha, \beta) \cdot \hat{\underline{E}}(\alpha, \beta) \exp\left\{ ik\left[\left(\alpha + \frac{2\pi n}{ka}\right)x + \left(\beta + \frac{2\pi m}{kb}\right)y + \gamma_{nm}\, z\right]\right\} \tag{11}$$

where $\underset{\approx}{Q}$ is a general dyadic quantity which depends on n, m, a b, α, β, and k, but not on Δ, and

$$\gamma_{nm} = \left[1 - \left(\frac{2\pi n}{ka} + \alpha\right)^2 - \left(\frac{2\pi m}{kb} + \beta\right)^2\right]^{1/2} . \tag{12}$$

If we use Eq. (4) to write $\hat{\underline{E}}$ in terms of $\hat{\underline{E}}_o$, and then integrate over all α and β, we find that the total field transmitted through the wire grid has the form

$$\underline{E}(x, y, z) = \left(\frac{1}{2\pi}\right)^2 \sum_{n,\,m=-\infty}^{\infty} \iint_{-\infty}^{\infty} d\alpha\, d\beta\, \underset{\approx}{Q}(n, m, \alpha, \beta) \cdot \hat{\underline{E}}_o(\alpha, \beta)$$

$$\cdot \exp\left[ik\left(\alpha + \frac{2\pi n}{ka}\right)x + ik\left(\beta + \frac{2\pi m}{kb}\right)y + ik(\gamma_{nm}\, z + h\Delta)\right] . \tag{12}$$

We now wish to study Eq. (12) in the limit when $z \to \infty$. From Eq. (11) it is clear that for a given pair of n and m, γ_{nm} is imaginary except in the appropriate region in α - β space shown as shaded in Figure 3. For example, for n = 1 and m = 0 the quantity γ_{oo} is imaginary, except in a circle of unit radius centered about $\alpha = -\lambda/a$ and $\beta = 0$ in the α - β plane. This region is the shaded circle denoted by n = 1 and m = 0 in Figure 3. In the unshaded regions we have that, for a given pair of n and m, the transmitted field is proportional to

$$\exp\left[-kz\left\{\left(\alpha + \frac{2\pi n}{ka}\right)^2 + \left(\beta + \frac{2\pi m}{kb}\right)^2 - 1\right\}^{1/2}\right] \tag{13}$$

which vanishes for large values of z. Therefore, for very large z we may rewrite Eq. (12) as

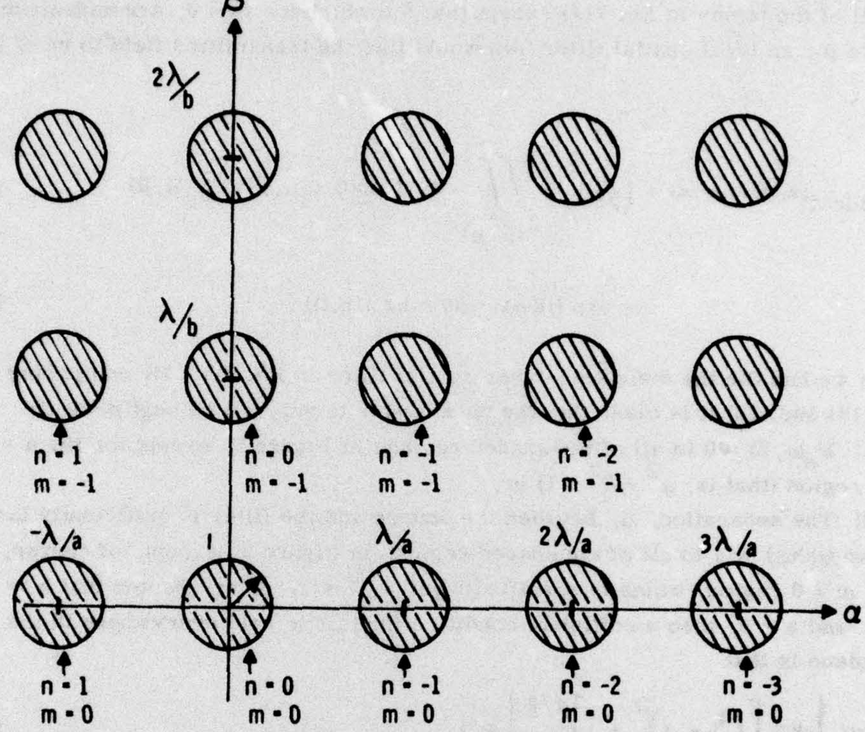

Figure 3. Alpha-Beta Space

$$\underline{E}(x, y, z \to \infty) = \left(\frac{1}{2\pi}\right)^2 \sum_{n, m=-\infty}^{\infty} \iint_{\substack{(n, m)\\ \text{shaded}\\ \text{region}}} d\alpha \; d\beta \; \underset{\approx}{Q}(n, m, \alpha, b) \cdot \underline{\hat{E}}_o(\alpha, \beta)$$

$$\cdot \exp\left[ik\left(\alpha + \frac{2\pi n}{ka}\right)x + ik\left(\beta + \frac{2\pi m}{kb}\right)y + ik(\gamma_{nm} z + h\Delta)\right] , \tag{14}$$

where each integral is only over the appropriate shaded region denoted by (n, m) in Figure 3. That is, when n = 1 and m = 0 the integration is only over the n = 1, m = 0 circle shown as shaded in Figure 3.

All of the terms in Eq. (14) except that for which n = m = 0, are undesirable because for an ideal spatial filter, we would like the transmitted field to be of the form*

$$E_{ideal}(x, y, z \to \infty) = \left(\frac{1}{2\pi}\right)^2 \iint\limits_{\alpha^2+\beta^2 \leq 1} d\alpha\, d\beta\, \underset{=}{Q}(0, 0, \alpha, \beta) \cdot \hat{\underline{E}}_o(\alpha, \beta)$$

$$\cdot \exp\left[ik(\alpha x + \beta y + hz + h\Delta)\right] \quad , \tag{15}$$

just as we had for the dielectric-layer spatial filter in Eq. (7). By comparing Eqs. (14) and (15) it is clear that the undesirable terms will be negligible if:

(1) $\hat{E}_o(\alpha, \beta) = 0$ in all of the shaded regions in Figure 3, except for the n = 0, m = 0 region (that is, $\alpha^2 + \beta^2 \leq 1$) or,

(2) The separation, Δ, between the source and the filter is sufficiently large that $\exp(ikh\Delta) \ll 1$ in all of the shaded regions in Figure 3, except, of course, the n = 0, m = 0 region, which is specified by $\alpha^2 + \beta^2 \leq 1$. If we assume that $\lambda \gg a$, $\lambda \gg b$, and $a \geq b$, then a sufficient condition for this to hold everywhere in the (α, β) plane is that

$$\exp\left\{-k\Delta\left[\left(\frac{\lambda}{a} - 1\right)^2 - 1\right]^{1/2}\right\} \ll 1 \tag{16}$$

which leads to the requirement that

$$\Delta \gg \frac{a}{2\pi}\left[1 - \frac{2a}{\lambda}\right]^{-1/2} \quad . \tag{17}$$

When Eq. (17) holds we can be assured, except for pathological cases, that the unwanted terms are unimportant, even when $\hat{\underline{E}}_o(\alpha, \beta)$ is non-zero in some of the shaded regions in $\alpha - \beta$ space in Figure 3. As an example of the practical meaning of Eq. (17), let us suppose that $a = 0.1\lambda$. Then Eq. (17) requires that $\Delta \gg 0.018\lambda$, a condition which is generally quite easy to satisfy. Therefore, if we are interested only in the far fields, it doesn't matter where we place the wire-grid spatial filter relative to the antenna, so long as the separation is such that Eq. (17) is satisfied.

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*Note that $\gamma_{oo} = h$

5. DISCUSSION

We have demonstrated that, if one assumes the original source distribution is unaltered by the presence of the spatial filter, the filtering properties of either infinite, planar dielectric layers or infinite, planar wire grids are the same whether we place the filter near to the source antenna or far from it. That is, the far field of an antenna with a dielectric-layer filter (which is infinite and homogeneous in the x-y direction) or a wire-grid filter in front of it, is independent* of the separation between the antenna and the filter. Of course, in practice, we cannot place the filter arbitrarily close to the source, because then our assumption that the source distribution is unaffected by the presence of the spatial filter, is no longer valid. It is probably necessary to have at least a $\lambda/4$ separation between the antenna and filter in order to avoid any significant distortion of the original current distribution on the source.

*Provided Eq. (17) holds, for the case of the wire grid.

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1. Mailloux, R. (1976) Synthesis of spatial filters with Chebyshev characteristics, IEEE Trans. Ant. and Prop. AP-24:174-181.

2. Fante, R., and Mayhan, J. (1970) Bounds on the electric field outside a radiating system, IEEE Trans. Ant. and Prop. AP-18:64-68.

3. Stratton, J. (1941) Electromagnetic Theory, Chapter 9, McGraw-Hill, New York.

4. Hill, D., and Wait, J. (1976) Electromagnetic scattering of an arbitrary plane wave by a wire mesh with bonded junctions, Canadian J. Phys. 54:353-361.