

# Lecture IX

## 2D EM modeling

In this lecture, we start to study 2D and 3D EM modeling. The simplest method is the FDTD method on rectangular grids.

### Part I TE and TM 2D fields

#### 1. Problem statement

In this lecture, we apply the FDTD method developed above for transmission lines to the solution of full-wave Maxwell's equations in free space. There are no significant differences between two approaches.

An inviting formulation is the two-dimensional *transverse magnetic* (TM) to  $z$  field. This field occurs in many problems, including the classic problem for scattering of a metal or dielectric cylinder with the incident field polarized along its axis,  $z$ . The field is characterized by the only one electric field component and two magnetic field components, in the perpendicular plane. Its complement is a TE to  $z$  field, with one H-field component, and two E-field components. In Appendix A, we give a short review of related practical problems where TM to  $z$  mode is *dominant*.

#### 2. Base example of a TM to $z$ field

A wireless link between two antennas through a door in a metal wall schematically shown in Fig. 1 is approximately represented by the TM to  $z$  mode. At a sufficiently high frequency, the field in the  $xy$ -plane is vertically polarized and has only two magnetic field components, both in the  $xy$ -plane.

#### 3. Differential equations

The 2D Maxwell's equations for the present case are written in the form (see Lecture#6)

$$\begin{aligned}\frac{\partial E_z}{\partial t} &= \frac{1}{\epsilon} \frac{\partial H_y}{\partial x} - \frac{1}{\epsilon} \frac{\partial H_x}{\partial y} - \frac{\sigma}{\epsilon} E_z \\ \frac{\partial H_x}{\partial t} &= -\frac{1}{\mu} \frac{\partial E_z}{\partial y} - \frac{\sigma^*}{\mu} H_x \\ \frac{\partial H_y}{\partial t} &= \frac{1}{\mu} \frac{\partial E_z}{\partial x} - \frac{\sigma^*}{\mu} H_y\end{aligned}\tag{1}$$

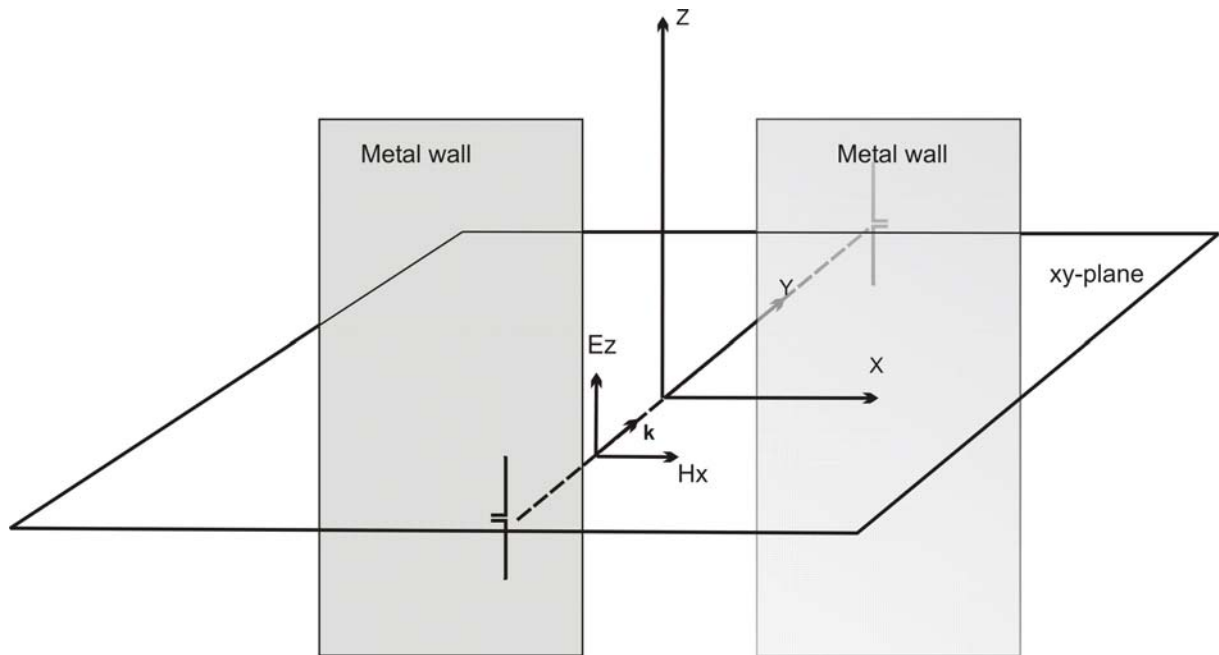


Fig. 1. One possible approximation for antenna radiation - a 2D TM to z solution.

where  $\sigma, \sigma^*$  are the electric and magnetic conductivities (electric and magnetic losses), real or artificial. The staggered grid to be used might be that from Lecture #2, where the electric field now plays the role of the voltage, and the magnetic field is the current. However, we have "two" current components but only "one" voltage component in the present case. Therefore, there is yet no a full similarity with the MTL case considered above in lecture #2.

## Part II Staggered rectangular (structured) grid and the FDTD solution

### 1. Staggered grid

Fortunately, the commonly accepted grid assembly is absolutely identical to that for the 1D MTL - see Lecture #2. Namely, the spatial steps in the general 3D case have the form:

$E_z$  is on the half-grid in  $z$  and on the integer grids in  $x$  and  $y$ ;

$H_x$  is on the half-grid in  $x$  and on the integer grids in  $y$  and  $z$ ;

$H_y$  is on the half-grid in  $y$  and on the integer grids in  $x$  and  $z$ .

In the 2D case, there is no  $z$ -dependence. One such 2D unit is shown in Fig. 2a. Marching on in time will lead to Fig. 2b, that is identical to Fig. 3 of Lecture #2.

Therefore, we will discretize Eqs. (1) in the form of Eqs. (11) of Lecture #2, i.e.

$$\begin{aligned}
E_z(t, z) &\rightarrow E_z(n \cdot \Delta t, k \cdot \Delta x, m \cdot \Delta y) = E_{z,k,m}^n \\
H_x(t, z) &\rightarrow H_x\left(\left(n + \frac{1}{2}\right) \cdot \Delta t, k \cdot \Delta x, \left(m + \frac{1}{2}\right) \cdot \Delta y\right) = H_{x,k,m+1/2}^{n+1/2} \\
H_y(t, z) &\rightarrow H_y\left(\left(n + \frac{1}{2}\right) \cdot \Delta t, \left(k + \frac{1}{2}\right) \cdot \Delta x, m \cdot \Delta y\right) = H_{y,k+1/2,m}^{n+1/2} \\
n &= 1, \dots, N_t + 1 \quad k = 1, \dots, N_x + 1 \quad m = 1, \dots, N_y + 1
\end{aligned} \tag{2}$$

Note that there is no interleaving between  $E_z$  and  $H_x$  with regard to the x-coordinate, and between  $E_z$  and  $H_y$  with regard to the y-coordinate, respectively. Such a property will be important for the difference equations.

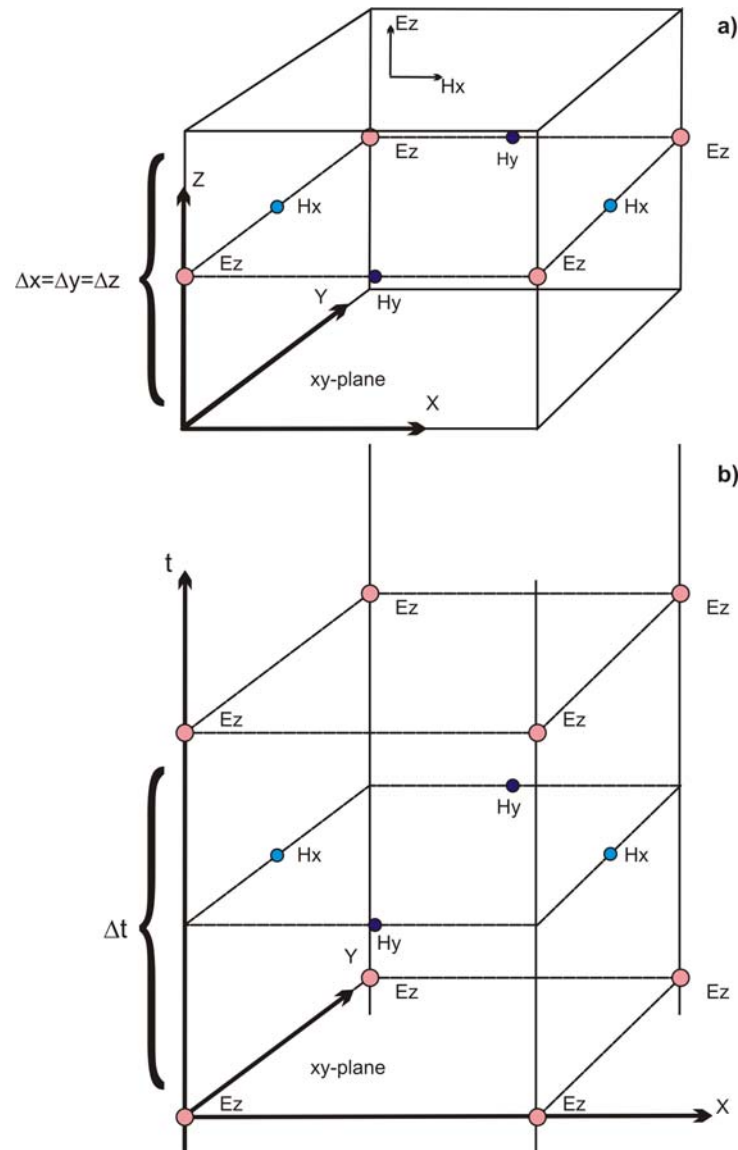


Fig. 2. Staggered grid for the TM to z mode: a) - unit cell; b) - interleaving in space/time.

## 2. FDTD equations

The corresponding difference equations thus have the form (the second-order differences similar to those used in lecture #2):

$$\begin{aligned}
 \frac{E_{z,k,m}^{n+1} - E_{z,k,m}^n}{\Delta t} &= \frac{1}{\epsilon_{k,m}} \frac{H_{y,k+1/2,m}^{n+1/2} - H_{y,k-1/2,m}^{n+1/2}}{\Delta x} - \frac{1}{\epsilon_{k,m}} \frac{H_{x,k,m+1/2}^{n+1/2} - H_{x,k,m-1/2}^{n+1/2}}{\Delta y} - \frac{\sigma}{2\epsilon_{k,m}} (E_{z,k,m}^{n+1} + E_{z,k,m}^n) \\
 \frac{H_{x,k,m+1/2}^{n+3/2} - H_{x,k,m+1/2}^{n+1/2}}{\Delta t} &= -\frac{1}{\mu_{k,m+1/2}} \frac{E_{z,k,m+1}^{n+1} - E_{z,k,m}^{n+1}}{\Delta y} - \frac{\sigma^*}{2\mu_{k,m+1/2}} (H_{x,k,m+1/2}^{n+3/2} + H_{x,k,m+1/2}^{n+1/2}) \\
 \frac{H_{y,k+1/2,m}^{n+3/2} - H_{y,k+1/2,m}^{n+1/2}}{\Delta t} &= +\frac{1}{\mu_{k+1/2,m}} \frac{E_{z,k+1,m}^{n+1} - E_{z,k,m}^{n+1}}{\Delta x} - \frac{\sigma^*}{2\mu_{k+1/2,m}} (H_{y,k+1/2,m}^{n+3/2} + H_{y,k+1/2,m}^{n+1/2})
 \end{aligned} \tag{3a}$$

The medium parameters could depend on the particular position (in an inhomogeneous medium), indeed.

Again, though  $\Delta x, \Delta y$  and  $\Delta t$  might appear to correspond to first-order differences, we remember that the corresponding derivative is always calculated at the center point of the  $\Delta z$  or  $\Delta t$  intervals, so that all approximations in Eq. (3a) are in fact second-order. A symbolic explicit form of marching on in time Eqs. (3a) is often used. That may be given by (only the time index is kept here)

$$\begin{aligned}
 E_z^{n+1} &= e_{z1} E_z^n + e_{z2} \Delta_x H_y^{n+1/2} - e_{z3} \Delta_y H_x^{n+1/2} \\
 H_x^{n+3/2} &= h_{x1} H_x^{n+1/2} - h_{x2} \Delta_y E_z^{n+1} \\
 H_y^{n+3/2} &= h_{y1} H_y^{n+1/2} + h_{y2} \Delta_x E_z^{n+1}
 \end{aligned} \tag{3b}$$

Here,

$$\begin{aligned}
 E_z^{n+1} &= E_{z,k,m}^{n+1} \\
 H_x^{n+1/2} &= H_{x,k,m+1/2}^{n+1/2} \\
 H_y^{n+1/2} &= H_{y,k+1/2,m}^{n+1/2}
 \end{aligned} \tag{3c}$$

and the symbol  $\Delta$  denotes the finite spatial differences in Eqs. (3a). Eqs. (3b) are very useful for programming purposes. The numerical constants in Eqs. (3b) are expressed in the form

$$\begin{aligned}
e_{z1} &= \frac{1-\tau/2}{1+\tau/2}, & e_{z2} = e_{z3} &= \frac{\Delta t}{\varepsilon_{k,m} \Delta d} \frac{1}{1+\tau/2}, & \tau &= \frac{\Delta t \sigma}{\varepsilon_{k,m}} \\
h_{x1} &= \frac{1-\tau^*/2}{1+\tau^*/2}, & h_{x2} &= \frac{\Delta t}{\mu_{k,m+1/2} \Delta d} \frac{1}{1+\tau^*/2}, & \tau^* &= \frac{\Delta t \sigma^*}{\mu_{k,m+1/2}} \\
h_{y1} &= \frac{1-\tau^*/2}{1+\tau^*/2}, & h_{y2} &= \frac{\Delta t}{\mu_{k+1/2,m} \Delta d} \frac{1}{1+\tau^*/2}, & \tau^* &= \frac{\Delta t \sigma^*}{\mu_{k+1/2,m}}
\end{aligned} \tag{3d}$$

Those numerical constants depend on material properties and on the FDTD grid parameters - spatial steps  $\Delta x = \Delta y = \Delta d$  and  $\Delta t$ . Also note that not only the electric permittivity and magnetic permeability, but also the losses,  $\sigma, \sigma^*$  may be functions of spatial coordinates. According to Eqs. (3a), all material parameters in the difference equations should be evaluated at the same nodes where the corresponding field quantities are defined – see Fig. 3 that follows.

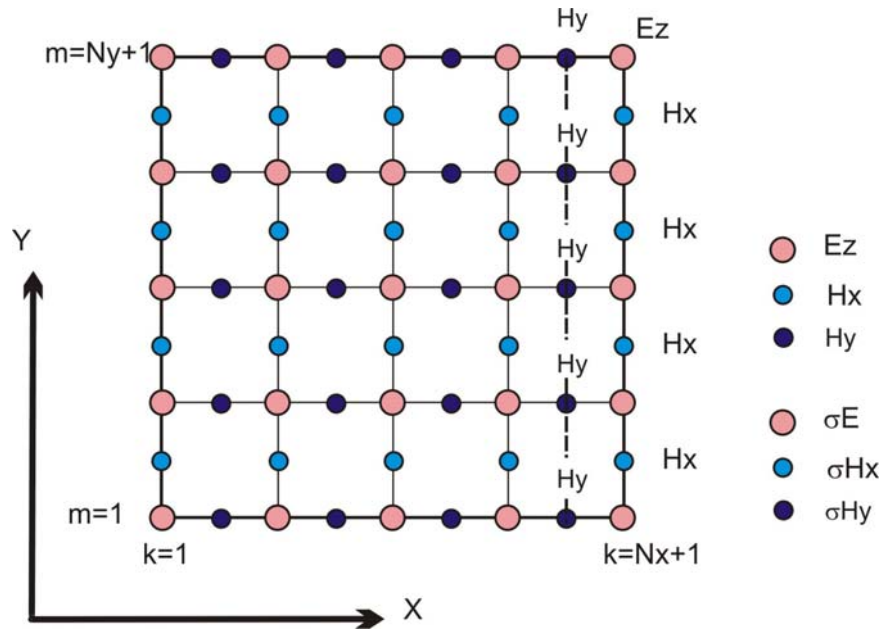


Fig. 3. 2D Staggered grid for the TM to z mode.

Last but not least we note that the losses are always divided by the permittivity and permeability; in other words, only ratios  $\sigma/\varepsilon, \sigma^*/\mu$  are important to us.

### 3. Initial conditions

In the present case, the solution could start with zero initial conditions, which imply that

$$\begin{aligned}
E_{zk,m}^1 &= 0 \\
H_{xk,m+1/2}^{3/2} &= 0 \\
H_{yk+1/2,m}^{3/2} &= 0 \\
n &= 1, 2, \dots, N_t + 1 \quad k = 1, 2, \dots, N_x + 1 \quad m = 1, 2, \dots, N_y + 1
\end{aligned} \tag{4}$$

Note again that these initial conditions may be not exact. Strictly speaking, both voltage and current should be given at the *same* initial time moment. However, such an approximation typically has little, if any, influence on the solution accuracy.

#### 4. Simple boundary conditions

In the case of a surrounding metal box, those conditions are quite simple (the tangential E-field component, i.e.  $E_z$ ) must be zero at the boundaries, i.e.

$$\begin{aligned}
E_{zk=1,m}^n &= E_{zk=N_x+1,m}^n = 0, \quad m = 1, 2, \dots, N_y + 1 \\
E_{zk,m=1}^n &= E_{zk,m=N_y+1}^n = 0, \quad k = 1, 2, \dots, N_x + 1
\end{aligned} \tag{5}$$

#### 5. Simple source excitation

At a certain node  $k = k_e, m = m_e$  in space, we might prescribe the source excitation in the form:

$$E_{zk=k_e,m=m_e}^n = E_0 \sin(\omega t_n) \tag{6}$$

This excitation approximately corresponds to a (long) dipole antenna located at the same point in space. In FDTD theory, it is also called the "hard source". The excitation may be terminated after certain time - see below in this Lecture.

## Part III Realization

### 1. Arrays of unknown fields

Marching on in time also implies that the solution at all time moments does not have to be stored. For example, six arrays - the *past* fields on the grid, and the *next* fields on the grid - are quite sufficient. Often, only three recursive arrays are used. Similar to Lecture #2, the arrays with integer and non-integer indexes in Eqs. (2) and (3) are replaced by the following MATLAB arrays:

```

HyN = zeros(Nx+0, Ny+1); % sixth layer in the series of six (n+3/2)
HxN = zeros(Nx+1, Ny+0); % fifth layer in the series of six (n+3/2)
EzN = zeros(Nx+1, Ny+1); % fourth layer in the series of six (n+1)

HyP = zeros(Nx+0, Ny+1); % third layer in the series of six (n+1/2)
HxP = zeros(Nx+1, Ny+0); % second layer in the series of six (n+1/2)
EzP = zeros(Nx+1, Ny+1); % first layer in the series of six (n+0)

```

Eqs. (2), (3) say that the marching on in time operates in the following way:

1. We find all electric field values  $E_{zN}$  in space at time step  $n+1$ ,  $t = (n+1)\Delta t$  using electric values  $E_{zP}$  at the previous time step  $n$ ,  $t = n\Delta t$  and magnetic field values  $H_{xP}$ ,  $H_{yP}$  at the intermediate time step  $n+1/2$ ,  $t = (n+1/2)\Delta t$ .
2. We find all magnetic field values  $H_{xN}$ ,  $H_{yN}$  in space at the intermediate time step  $n+3/2$ ,  $t = (n+3/2)\Delta t$  using magnetic field values  $H_{xP}$ ,  $H_{yP}$  at the previous intermediate time step  $n+1/2$ ,  $t = (n+1/2)\Delta t$  and electric field values at time step  $n+1$ ,  $t = (n+1)\Delta t$ .
3. We find all electric field values ..., etc.

The method thus remain "bootstrapping" - see Lecture #2.

## 2. Vector operations - batch mode

Eqs. (2) and (3) say that the FDTD method is equivalent to the addition/subtraction of vectors without index shift, or with one shifted index. The argument values from Eqs. (3) corresponding to those six arrays are then programmed in a relatively simple form:

```
% x- and y-coordinates for the E-field nodes
Lx = 3.00;          Ly = 3.00;          % in meters;
Nx = 100;           Ny = Nx;
dx = Lx/Nx;         dy = Ly/Ny;
% x- and y-coordinates for the E-field nodes
x = [0:dx:Lx];      y = [0:dy:Ly];      % Nx+1 total
```

The batch (fast) mode in MATLAB is used to do marching on in time, i.e.

```
indv = 2:Nx;          % batch process in space (E-nodes)
indi = 1:Nx;          % batch process in space (H-nodes)

n = 1;
EzN(k_e, m_e) = E0(n); % excitation ( a hard source)

while n < NT-1
    EzN(indv, indv) = ez1*EzP(indv, indv) + ...
                    ez2*(HyP(indv, indv) - HyP(indv-1, indv)) - ...
                    ez3*(HxP(indv, indv) - HxP(indv, indv-1));

    EzN(k_e, m_e) = E0(n+1);

    HxN(indv, indi) = hx1*HxP(indv, indi) - ...
                    hx2*(EzN(indv, indi+1) - EzN(indv, indi));
    HyN(indi, indv) = hy1*HyP(indi, indv) + ...
                    hy2*(EzN(indi+1, indv) - EzN(indi, indv));

    EzP = EzN; HxP = HxN; HyP = HyN;

    n = n + 1;
end
```

The FDTD method is done! The above equations automatically satisfy the boundary conditions of a big metal box (do not update the boundary nodes). The corresponding MATLAB realization is given in the scripts `fddt_2d_01.m` and `fddt_2d_02.m` (`fddt_2d_02.m`) to this Lecture.

### 3. A magic time step

The same dispersion analysis may be applied as in Lecture #2. It leads to the following expression for the "magic" (or dispersion-preserving) time step

$$\Delta t = \frac{1}{c_p \sqrt{1/\Delta x^2 + 1/\Delta y^2}} \quad (7)$$

which replaces Eq. (21) of Lecture #2.

### 4. Experimentation

With the script `fddt_2d_01.m`:

1. Simple propagation
2. How much time does it take to run the script without the figure (to blocks need to be commented)?
3. Effect of a metal wall
4. Effect of absorption (time permitting)

With the script `fddt_2d_02.m` (`fddt_2d_03.m` - Faraday Cage):

1. Fast graphics
2. MATLAB function `diff`
3. Diagonal propagation
4. Faraday cage

## Part IV About a 3D solution

### 1. Staggered grid

The 3D staggered grid for a unit cube is shown in Fig. 4. The same idea of bootstrapping as before applies, this time - in space.

The electric field is given at the edge centers. In particular,  $E_x$  is placed at the centers of the edges *parallel* to the x-axis,  $E_y$  is placed at the centers of the edges *parallel* to the y-axis, etc. The magnetic field is given at the face centers. In particular,  $H_x$  is placed at the

centers of the faces *perpendicular* to the x-axis,  $H_y$  is placed at the centers of the faces *perpendicular* to the y-axis, etc.

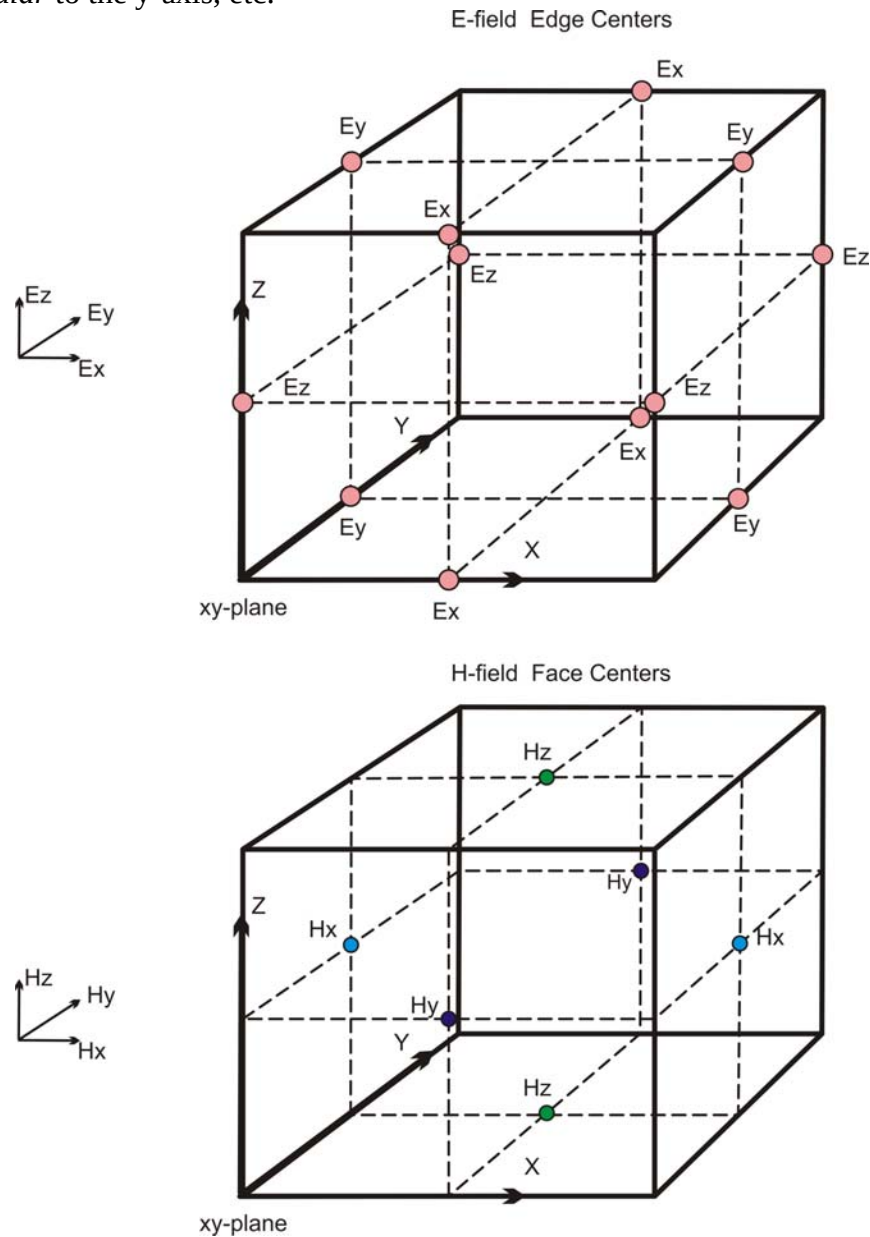


Fig. 4. Staggered grid in 3D.

The rest of the solution remains ideologically the same.

## HW#9 exercises

### 1. Energy calculations (is the total energy preserved?)

The total field energy is calculated for a two-dimensional box  $D$  according to

$$E_{\text{total}}(t) = \frac{1}{2} \iint_D \epsilon E_z^2 dx dy + \frac{1}{2} \iint_D \mu (H_x^2 + H_y^2) dx dy \quad (8)$$

To find  $E$ , the corresponding MATLAB code should find the average energy value over every unit cell first, and then find the sum of all such values.

Using MATLAB script `fdtd_2d_02.m` please,

- A. Program the above formula for  $E_{\text{total}}(t)$ ;
- B. Plot the energy behavior to scale over the time interval from 0 to 50 ns (insert the grid into the plot).

### 2. 3D FDTD

Using the textbook:

1. Retype the MATLAB code on pages 71 to 72.
2. Explain what exactly is this code doing?
3. How much CPU time per step does this code require (use  $N_t=5$ ) to run a 3D problem with
  - $N_x=100, N_y=100, N_z=100$ ?
  - $N_x=300, N_y=300, N_z=300$ ?
4. Given that there are 10 nodes per wavelength, a cubic volume of what size could we model for  $N_x=300, N_y=300, N_z=300$  at 1 GHz using the amp server?

## Appendix A Examples of 2D TM/TE fields

### A1. Diffraction. Effect of ground plane. Antenna backlobe

A question of significant practical importance relates to understanding the impact of the finite ground plane on the antenna radiation. The geometry of a dipole antenna above the finite ground plane is shown schematically in Fig. A1 - I. A first naive guess about the field distribution of the antenna is that of pure *geometrical optics* (GO)– see Fig. A1 -I. The geometrical optics predicts no radiation in the shadow zone, below shadow boundary denoted by SB.

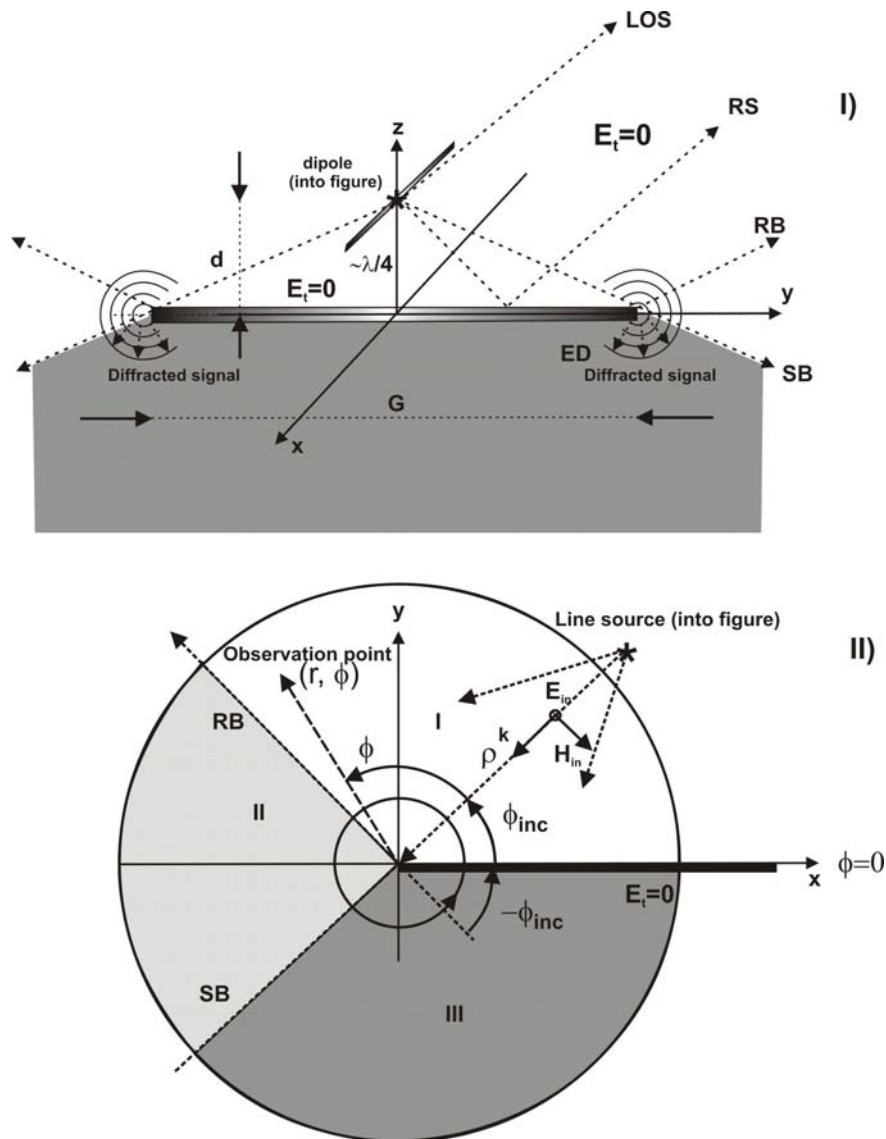


Fig. A1. Top - geometrical optics approximation for a dipole above a finite ground plane; bottom problem formalization in terms of the TM to  $z$  field.

In terms of geometrical optics, the total field is considered to be a combination of rays emanating from the dipole and then reflected from the metal surface according to the Snell's law. According to geometrical optics the field everywhere within the reflection boundary (RB) in Fig. A1 is a combination of the incident (line-of-sight or LOS signal) and the reflected signal (RS). The field everywhere outside the reflection boundary but still inside the shadow boundary (SB) is the LOS signal from the dipole. The field below the shadow boundary is zero.

However, the practice shows that any finite ground plane cannot block the signal in the shadow zone – the field in the shadow zone is nonzero, which leads to antenna *backlobe*. The backlobe creates an extra noise signal and is especially important for high-fidelity antennas, e.g. for GPS antennas. The backward radiation indeed decreases, when the ground plane size increases, but not monotonically. The antenna backlobe and the diffraction effect can be studied and quantified the Geometrical Theory of Diffraction (GTD) that mostly related to 2D problems.

In the first half of the last century there was relatively little work on high frequency diffraction. However all that changed in 1953 when Keller introduced the Geometrical Theory of Diffraction (GTD) as an extension of geometrical optics to include diffracted rays. This theory introduces diffracted rays in addition to the usual geometric optical rays. Unlike geometric optics, the diffracted rays can enter the shadow regions. This was a very important development, because it made it possible to calculate the high frequency radiation from antennas and scatterers of a quite general shape and to understand the various radiation mechanisms involved. Also, the GTD motivated the asymptotic treatment of the numerous canonical problems to determine the fields diffracted from edges, vertices and smooth curved surfaces.

On the other hand, the same problem can be *quantitatively* analyzed numerically, including the relative backlobe level, by virtue of the 2D TM to z-field shown in Fig. A1 –II. We will use the numerical approach as being more simple and versatile.

## A2. Horn antenna

Fig. A2 shows the *E*-plane and *H*-plane cuts of a rectangular or pyramidal horn. When studying the radiation pattern the throat region with the feed is usually not considered. Instead, one replaces the feed region by a point source of a diverging wave. That is usually a  $TE_{10}$  mode of the rectangular waveguide –with the spherical wave front (or *eiconal*). For a 2D case or when the horn dimension in the *E*-plane is much smaller than in the *H*-plane, the diverging  $TE_{10}$  mode is reduced to a cylindrical diverging wave emanating from a line source at origin as shown in Fig. A2.

Thus, the 3D pattern behavior of a real pyramidal horn including the relative sidelobe level is very well described by two 2D models:

- E-plane pattern obtained using the 2D TE to z solution in Fig. A2 - I
- H-plane pattern obtained using the 2D TM to z solution in Fig. A2 -II

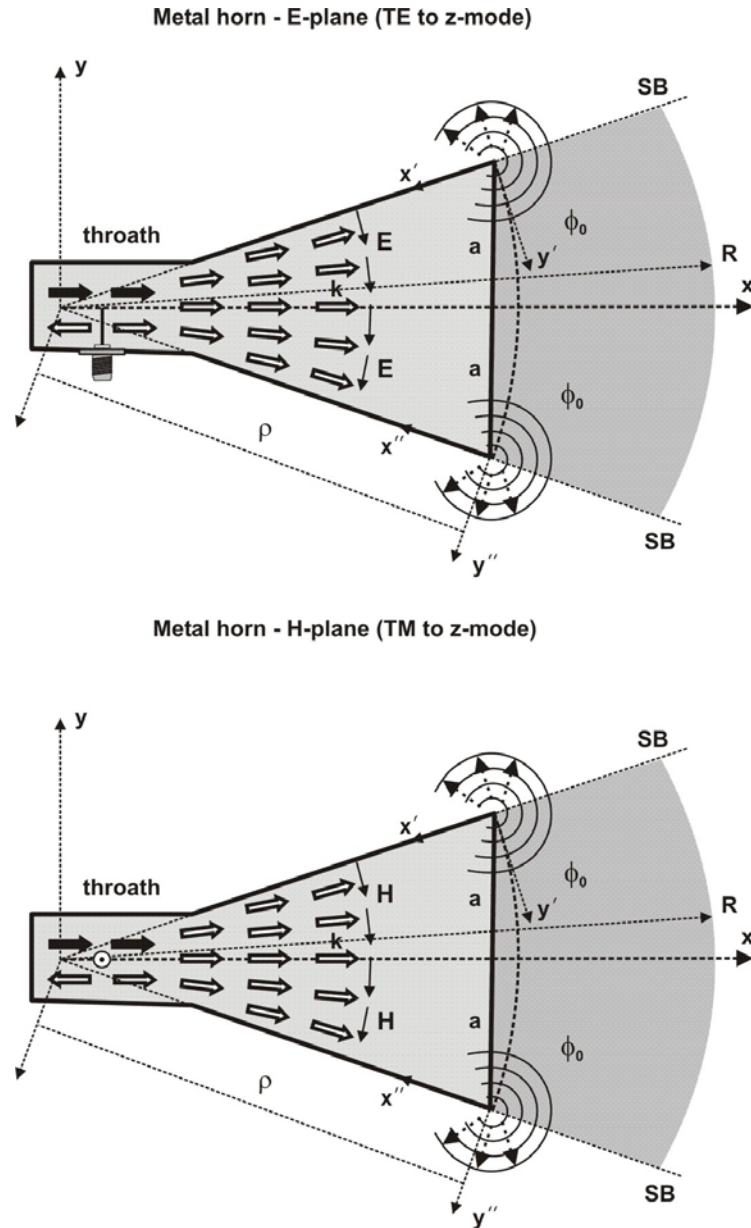


Fig. A2. Separation of the horn antenna radiation problem into TE to z and TM to z modes.

This approach cannot be used to accurately predict the impedance of a probe-fed horn antenna, indeed. However, it is very well suited for pattern optimization including

- the horn geometry;

- the rolled edge effect;
- the corrugation effect, etc.

### A3. Parabolic reflector

Fig. A3 shows a cross-section of the cylindrical parabolic reflector infinite in the  $z$ -direction – we again consider a 2D case.

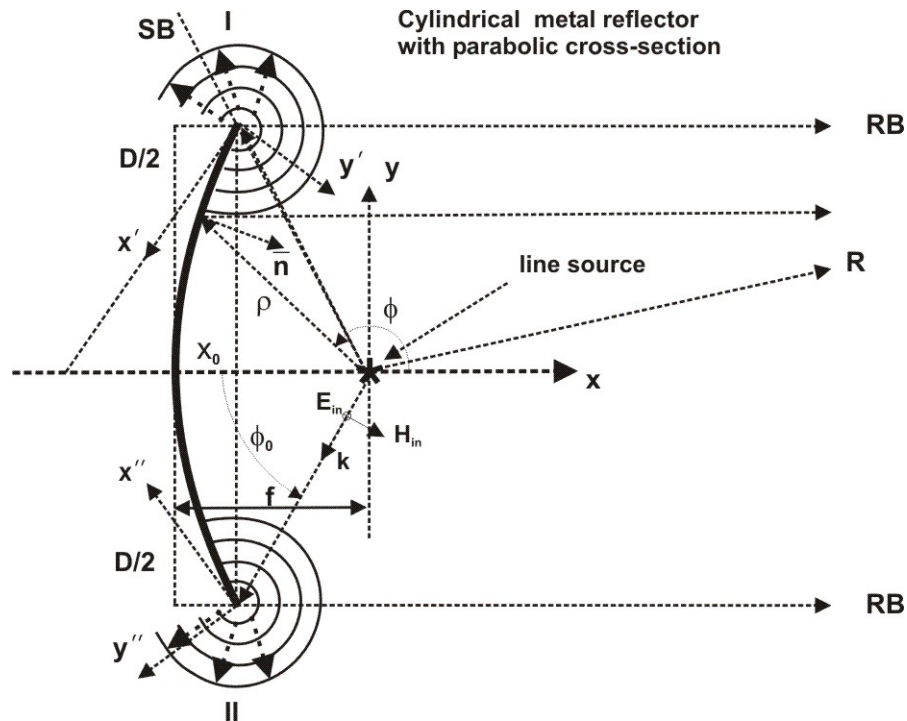


Fig. A3. The TM to  $z$  field for a parabolic reflector.

The resulting field is again the TM to  $z$  field. In polar coordinates shown in Fig. A3, the equation of the parabolic surface is

$$\rho = \frac{f}{\sin^2(\varphi/2)}, \quad x_0 = \frac{D^2}{16f}, \quad D = 4f \tan(\varphi_0/2)$$

and the unit normal to the surface has the form

$$\bar{n} = \bar{x} \sin \frac{\varphi}{2} - \bar{y} \cos \frac{\varphi}{2}$$

where  $\varphi_0$  is the half *subtended* angle of the reflector.

The most important task in the design of a parabolic reflector antenna is predicting the main lobe and several sidelobes. Geometrical optics cannot accomplish this task - the parabolic surface will create a beam of plane waves that has formally an infinite gain at

infinity (no spherical or cylindrical divergence). Hence, the GTD and UTD (Uniform Theory of Diffraction) needs to be developed for curved surfaces – not only for the edges or wedges – which is a difficult problem in 3D. The problem here is that the GTD predicts a non-physical solution at a ray intersection – *caustic* – and such intersections are quite common for curved concave surfaces. The solutions are available but is rather cumbersome. A “competitor” to UTD is a *physical optics approximation* that is usually used to study the main beam. However, the UTD formulation is still very useful in predicting the backlobe radiation from a reflector where the physical optics does not give accurate results. Therefore, it the only analytical method to estimate the backlobes of a large reflector.

On the other hand, the same problem can be analyzed numerically, including the relative backlobe level and pattern dynamics, by virtue of the 2D TM to z-field shown in Fig. A3. We will again use the numerical approach as being more simple and versatile.

#### A4. Frequency selective surface (FSS) and artificial magnetic conductors

Fig. A4 shows the oldest (and perhaps simplest) frequency selective surface on the base of the quarter wave metal corrugation.

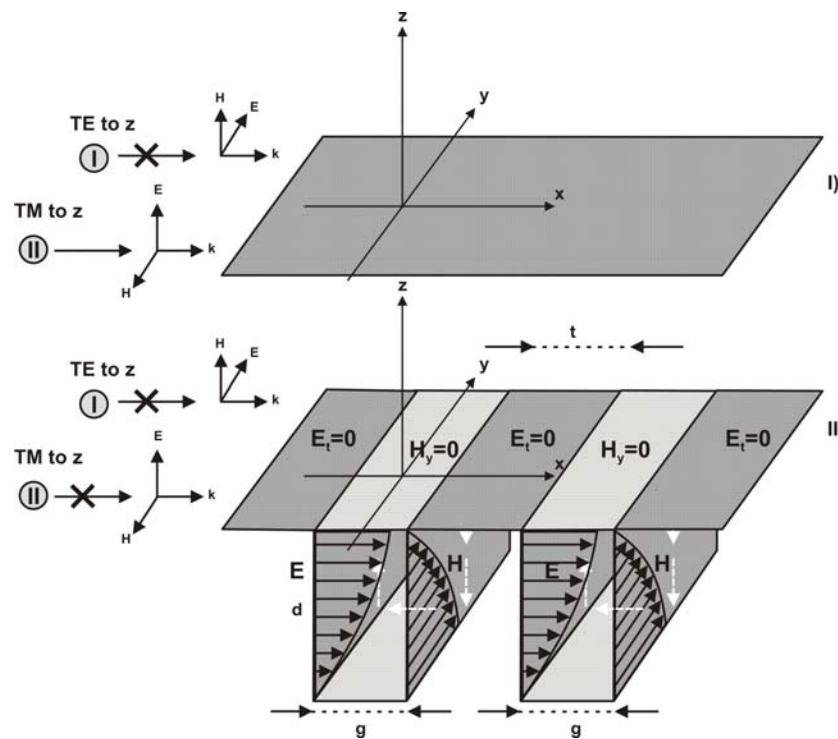


Fig. A4. A corrugated metal surface on the base of quarter wave resonators (bottom) versus the solid metal ground plane – top.

We shall briefly compare the performance of the corrugated surface on the base of quarter wavelength resonators with that of the solid metal surface – see Fig. A4- I.

We first consider two plane wave types labeled as I and II in Fig. A4, which potentially might propagate along both surfaces. Oblique plane wave of type I is often designated as a TE to  $z$  or, shortly,  $TE_z$  plane wave since it does not have a  $E$ -field component in the  $z$ -direction. Similarly, the oblique plane wave of type II is the TM to  $z$  or  $TM_z$  plane wave. The metal surface shorts out signal I since  $E_t = 0$  but still allows for signal II to propagate. Similarly, for the corrugated surface, the metal teeth require  $E_t = 0$  at top and will still short out the signal I very efficiently. At the same time, the waveguide openings require  $H_y = 0$  at their top and will short out the signal II, for which it must be  $H_y \neq 0$ . Thus, neither mode I nor mode II can propagate along the resonant corrugation in contrast to the solid metal. Such a property of corrugated surfaces is important and allows us to use it as a special finite ground plane, in particular for circularly-polarized antennas.

The idea of the corrugated surface as a ground plane for an antenna is perhaps best explained on the base of Fig. A5. Consider a monopole-like antenna located in the middle of a metal (Fig. A5 -left) or a corrugated (Fig. A5-right) finite ground plane. The metal ground plane supports the plane wave of type II that travels all the way toward the surface edge and then is being radiated as if the edge tip were an essentially point source.

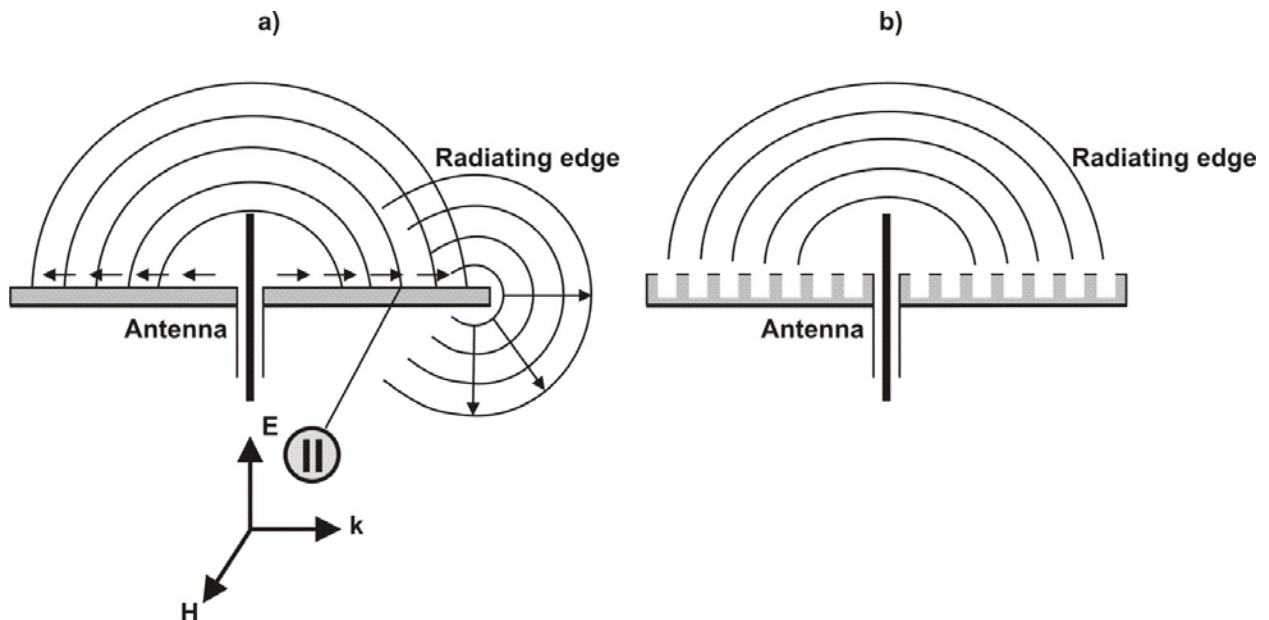


Fig. A5. A monopole-like antenna in the middle of a finite ground plane; a) –finite metal ground plane; b) – finite cutoff corrugation surface. The signal radiated from ground plane edge is schematically shown.

The edge radiation is also directed downwards, which means a significant antenna backlobe. Due to reciprocity, the same pattern holds for the receiving antenna. The receiving antenna will thus not only receive the useful signal from a positive elevation angle but also will receive all the electromagnetic noise that is emitted by ground and,

what might be more important, the noise emitted by possible interference sources located close to horizon. These sources could be due to a residential activity for larger UHF or L/S band antennas. For an antenna printed on a chip, these sources could be due to neighbor on-chip components.

On the other hand, the cutoff corrugated surface shown in Fig. A5 - right will suppress either plane or surface wave and will thus prevent the edge radiation. Hence, the backlobe radiation and radiation at very low elevation angles (from horizon) should be significantly reduced. Indeed, the present property of the corrugated surface holds only for a certain band of frequencies.

The quantitative parametric modeling of a 3D metal corrugation in Ansoft HFSS is computationally extensive. On the other hand, a 2D TM approach allows us to study the TE to z mode in Fig. A5 at different angles of incidence. This problem corresponds to a horizontal dipole with a FSS reflector. The 2D TE approach will allow us to study the TE to z mode in Fig. A5 at different angles of incidence.