Electromagnetic near field and the far field, 
Chapter two

Pawel Bienkowski and Hubert Trzaska - January 28, 2013

Editor’s note: This book is useful for all designers who need to test or have their designs tested for emissions, radiation and susceptibility for EMC. It can also help a wide range of specialists in biology, medicine, labor safety, environmental protection, metrologists, EMF meter designers, testers and users, and even for those who must make legal decisions on the grounds of measurement results interpretation.

I will publish this chapter two which follows and another chapter 7, “Directional Pattern Synthesis” in the near future. Dust off your Maxwell’s equations and dig in!

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Chapter 2

The Near Field and the Far Field

The essential information for practical metrology is presented in this chapter, including a brief summary of the near-field properties as well as the basic equations and formulas related to fields generated by simple radiation sources.
2.1 AN EMF GENERATED BY A SYSTEM OF CURRENTS

Consider the calculation of the EMF at an arbitrary point, \( P \), external to a volume containing arbitrary currents. Figure 2.1 shows a volume \( V \), of arbitrary cross section, containing a system of arbitrarily oriented electric and magnetic currents, \( J \) and \( \ast J \), respectively. The volume \( V \) is surrounded by an infinitely large, homogeneous, isotropic, linear, lossless medium of permeability \( \mu \) and permittivity \( \varepsilon \). Electrical parameters are continuous on the boundary surface. The maximal linear size of the volume, \( V \), is \( D \) (Fig. 2.1).

The \( E \) and \( H \) fields can be determined at an arbitrary point situated outside the volume, \( V \), by solving Maxwell’s equations for the angular frequency, \( \omega \) [1]:

\[
E = \nabla \times \nabla \times \Pi + j\omega \mu \nabla \times \ast \Pi \tag{2.1}
\]

\[
H = \nabla \times \nabla \times \ast \Pi - j\omega \varepsilon \nabla \times \Pi \tag{2.2}
\]

Where \( \Pi \) and \( \ast \Pi \) are electric and magnetic Hertzian vectors:

\[
\Pi = \frac{1}{j4\pi\omega\varepsilon} \int_{V} J \frac{\exp(-jkr)}{r} dV \tag{2.3}
\]

\[
\ast \Pi = \frac{1}{j4\pi\omega\mu} \int_{V} \ast J \frac{\exp(-jkr)}{r} dV \tag{2.4}
\]

Where \( k \) = the propagation constant

\[
k = \omega \sqrt{\varepsilon\mu} \tag{2.5}
\]
\[ r = \text{the distance from the observation point } P \text{ to a general point in the volume } Q (R', \Theta', \vartheta') \text{ so} \]
\[ r = R - R' \]
with a resultant magnitude of:
\[ r = \sqrt{R^2 + R'^2 - 2RR' \cos \beta} \]

(2.6)

where:
\[ \beta = \text{an angle between } R \text{ and } RR', \]
\[ R = \text{the distance from the observation point to the center of the coordinate system}, \]
\[ R' = \text{the distance from the general point to the center of the coordinate system}. \]

Under the condition \( R' < R \), the distance \( r \) may be presented with the use of a series expansion [Eq. (2.7)]:

\[
r = R \left[ 1 - \frac{R'}{R} \cos \beta + \frac{R'^2}{2R^2} (1 - \cos^2 \beta) + \frac{R'^3}{2R^3} \cos \beta \left( 1 - \cos^2 \beta \right) - \cdots \right]
\]

(2.7)

If \( R >> D \) (where \( D \) is the maximal size of an arbitrary cross section of the volume \( V \)), it is possible to assume that \( r \) is parallel to \( R \), so \( r = R - R' \cos \beta \). Then:

\[
\Pi_\infty = \frac{\exp \left( -jkr \right)}{j4\pi\omega R} \int_V J \exp \left( -jkr' \cos \beta \right) \, dV
\]

(2.8)

\[
{^*\Pi_\infty} = \frac{\exp \left( -jkr \right)}{j4\pi\omega R} {^*} \int_V {^*}J \exp \left( -jkr' \cos \beta \right) \, dV
\]

(2.9)

The index \( \infty \) in the formulas indicates that they are valid for \( R >> D \).
In this case the spatial components of $E$ and $H$ are given by:

$$E_{\theta\infty} = -j k (Z \Pi_{\theta\infty} + \Pi_{\phi\infty}) \quad (2.10)$$

$$E_{\phi\infty} = -j k (Z \Pi_{\phi\infty} - \Pi_{\theta\infty}) \quad (2.11)$$

$$E_{R\infty} = H_{R\infty} = 0 \quad (2.12)$$

$$H_{\phi\infty} = \frac{E_{\theta\infty}}{Z} \quad (2.13)$$

$$H_{\theta\infty} = -\frac{E_{\phi\infty}}{Z} \quad (2.14)$$

where:

$\Pi_{\theta\infty}, \Pi_{\phi\infty}, *\Pi_{\theta\infty}, *\Pi_{\phi\infty}$ = the spatial components of vector $\Pi_{\infty}$ and * $\Pi_{\infty}$.

$Z =$ wave impedance of the medium:

$$Z = \frac{\sqrt{\mu}}{\varepsilon} = Z_0 \frac{\sqrt{\mu}}{\varepsilon_0} \quad (2.15)$$

$Z_0 =$ intrinsic impedance of free space:

$$Z_0 = \frac{\sqrt{\mu_0}}{\varepsilon_0}$$

Equations (2.10–2.14) allow us to find the far-field EMF components of an arbitrary system of currents in volume $V$. The field may be characterized as follows:

- The EMF in the far-field is a transverse field [Eq. (2.12)].
- At an arbitrary point the EMF has a shape of the TEM wave [Eqs. (2.13) and (2.14)].
- Vectors $E$ and $H$ can have two spatial components that are shifted in phase; as a result, the field is
• The dependence of the E and H fields on . and . is described by the normalized directional pattern that is independent of R.

• The E and H components are mutually perpendicular and related by the wave impedance of a medium.

• The Poynting vector $S = E \times H$ is real and oriented radially.

To characterize the EMF properties in the far field, we have presented a straightforward solution of Maxwell’s equations. To get a fully generalized solution of the equations, it would be necessary to take into account a number of additional factors including: the diffraction of a wave caused by irregularities in a nonhomogeneous medium, the dispersion and nonlinear properties of the medium, the anisotropy of the material, and the superposition of waves when nonmonochromatic fields are being considered. Such a solution has not been fully formulated. However, a fully general solution is not crucial for metrology and the following sections are based on the previous analysis.

THE FAR FIELD AND THE NEAR FIELD

2.2 THE FAR FIELD AND THE NEAR FIELD

The considerations presented above lead us to the description of several features that characterize the far-field. There are no actual discontinuities between the far field, the intermediate field, and the near field. However, in order to adequately describe these regions, one of the criteria for their delimitation is presented below [2].

If we calculate the difference between the distance $r$ given by Eq. (2.6) and its approximate magnitude given by the first two terms of the series in Eq. (2.7), we get an error distance. If we then multiply this by the wavenumber $k$, we have the phase error, $\Delta \Psi$. The limits to the use of the approximation $R >> D$ is defined by the error and may be expressed in the form:

$$\Delta \Psi = k \left( \frac{R^2}{2R} \sin^2 \beta + \frac{R^3}{2R^2} \cos \beta \sin^2 \beta + \cdots \right)$$

(2.16)

At its maximum, $2R' = D$, and this will result in the maximum value of the error given by following formula:

$$\Delta \Psi_{\text{max}} = \frac{kD^2}{8R} = \frac{2\pi}{N}$$

(2.17)

where: $N = \text{a number defining the acceptable inaccuracy of the phase front.}$
Usually, it is assumed that $N = 16$. So, substituting this in Eq. (2.17) and rearranging, we get:

$$R \geq \frac{2D^2}{\lambda}$$

(2.18)

The condition is widely accepted as the onset of the far field where $D$ is now regarded as the maximum linear dimension of the antenna or other radiating structure. To illustrate this, consider two examples relating to antennas working at different frequencies:

- A parabolic reflector antenna of 3 m diameter working within the 10-GHz band.
- The tallest antenna long-wave transmitting antenna in the world, in Gabin (Poland), with a maximum height above $0.5\lambda$, operating at 227 kHz (unfortunately, the antenna collapsed during guy-wire renovation, several years ago).

In both cases, the far-field can be obtained from Eq (2.18) and begins at distance above about 600 m away from the antenna. In the first case, $R \sim 2 \times 3^2/0.03 = 600$ m, and in the second case $R \sim 2 \times 661^2/1322 = 661$ m.

If, in our consideration, three terms of series [Eq. (2.7)] are taken into account with the other terms considered as vanishingly small, we have:

$$r = R - R' \cos \beta + \frac{R'^2}{2R} \sin^2 \beta$$

(2.19)

and then similar considerations to the far field are repeated, we obtain the following condition:

$$R \leq \frac{D}{4} + \frac{D}{2}\left(\frac{D}{\lambda}\right)^{\frac{3}{2}}$$

(2.20)

Equation (2.20) gives the limit of the near field.

Figure 2.2 shows (after [3]) the field regions around an aperture antenna. Figure 2.3 illustrates the near- and the far-field boundaries as a function of $r$, $D$, and $\lambda$. The near field and the intermediate field are referred to as the Fresnel region (Fresnel zone), and the far field is referred to as the Fraunhofer region or the radiation field. When in close proximity to a radiation source, where the field may be assumed to be stationary, i.e., it is not radiating, the $E$ and $H$ fields are mutually independent. In this case, the use of the Biot-Savart law and Coulomb’s law are usually assumed as sufficient. The field, in such close proximity to the antenna, is described as an induction field. Here, the imaginary part of the Poynting vector is dominant.
A word of comment

The formulas introduced to define the near-field boundary [Eq. (2.20)] and the far-field boundary [Eq. (2.18)] require a word of comment. The series expansion given by Eq. 2.7 is true if \( R' < R \), or more precisely, if:

\[
R' < R \left( \cos \beta + \sqrt{\cos^2 \beta + 1} \right)
\]

Although the conditions are not always fulfilled, Eqs. (2.18) and (2.20) are widely applied in the
literature as definitions of the far-field and near-field limits. The accepted approximation here is a result of arbitrarily assumed permissible nonhomogeneity of the phase front $N$, rather than some inherent property of the fields themselves. At the boundary (or “border”), no actual discontinuity exists, and the expression boundary or border was introduced here so as to systematize the EMF parameters in the region surrounding a source.

The near- and far-field definitions presented above are not the only definitions that exist. The boundary criterion may be based on, for instance, the convergence of the $E/H$ ratio to $Z_0$, the Poynting vector to the electric (magnetic) power density, and others, but they are more difficult to systematize in the way that was done above, and the use of other criteria will give the boundary in the form of complex spatial function, while that given by Eq. (2.18) presents a sphere of radius $R$. Nevertheless, any criterion is based upon arbitrarily chosen values of a parameter, and the choice may be difficult to justify (e.g., why we accepted $N = 16$ instead of 15 or 17).

As an example of another point of view, a definition of the EMF region boundary around radio-station antennas is presented in Table 2.1 [4]. With no regard to the above, the role of the boundary in EMF metrology is limited only to warning that measurements performed at distances below $R$ may be more troublesome and increasingly error-prone. Due to this, the authors have formulated a new definition of the near-field boundary: the near field is everywhere, where we perform EMF measurements. The definition is based upon longtime experience and observations of interfering and superposed fields, even at distances well in excess of $R$, due to multipath propagation and other phenomena, fields where a direct ray may not exist at all.

### Table 2.1 Boundary of EMF Specific Regions around Radio-Station Antennas [4]

<table>
<thead>
<tr>
<th>Region edges, measured from antenna</th>
<th>Region I $\lambda$</th>
<th>Region II $\delta\lambda$</th>
<th>Region III $\delta\lambda$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\lambda$ = wavelength</td>
<td>$\frac{\lambda}{D^2}$</td>
<td>$\frac{\delta\lambda}{D^2}$</td>
<td>$\frac{\delta\lambda}{D^2}$</td>
</tr>
<tr>
<td>$D$ = largest dimension of the antenna</td>
<td>$\max\left(\frac{\lambda}{D}, \frac{\delta\lambda}{D}\right)$</td>
<td>$\max\left(\frac{\delta\lambda}{D}, \frac{0.6D^2}{\lambda}\right)$</td>
<td>$\max\left(\frac{\delta\lambda}{D}, \frac{0.6D^2}{\lambda}\right)$</td>
</tr>
<tr>
<td>$E \perp H$</td>
<td>No</td>
<td>Effectively yes</td>
<td>Yes</td>
</tr>
<tr>
<td>$Z = E/H$</td>
<td>$\approx Z_0$</td>
<td>$\approx Z_0$</td>
<td>$\approx Z_0$</td>
</tr>
<tr>
<td>Component to be measured</td>
<td>$E$ and $H$</td>
<td>$E$ or $H$</td>
<td>$E$ or $H$</td>
</tr>
</tbody>
</table>

While spatial EMF components in the near-field are calculated, the rigorous use of the general dependencies [for instance, Eqs. (2.3) and (2.4)] is indispensable and appropriate precautions should be taken when any simplifications in calculations are planned. Special caution is necessary when applying software for numerical analysis without an appropriate analysis of the simplifications and assumptions that have been accepted in the procedures.

As noted in Section 2.1, properties of EMF in the far field appear partly in the intermediate field as well, although none of them appears in the near field. This results in the necessity of the specific methods use for EMF measurements in both regions. Several examples are quoted below to illustrate this point:
• In the far field, E and H measurements are fully equivalent, and they permit the calculation of the other components as well as S. In the near field, separate E and H component measurements are indispensable, and they significantly complicate the issue of the S measurement.

• The EMF polarization in the near field, especially in conditions of multipath propagation, may be quasi-ellipsoidal because of the spatial orientation variations of the polarization ellipse. This is due to, for instance, the frequency of source variations as a result of its FM modulation, Doppler effect due to reflection from a moving object, and so on.

• The radiation pattern in the far field is constant and independent of the distance to a source; on the ground of the near-field measurements, it may be calculated only for sources of regular structure using complex computations [5].

• The Poynting vector in the near field is complex, and its direction and magnitude are functions of the source structure and the distance to the source.

A comparison of the requirements in near- and far-field EMF measurements is presented in Table 2.2.

EMF FROM SIMPLE RADIATING STRUCTURES

2.3 EMF FROM SIMPLE RADIATING STRUCTURES

If in Eqs. (2.3) and (2.4) we assume that the electric current has a nonzero magnitude in the direction of axis z, i.e.: *J = 0 and |J| = J = constant, for:

\[-1/2 < z < +1/2\]

and at the same time

\[l << \lambda \text{ and } l << R\]

(where \(l = \) the length of the dipole’s arm), then the calculated \(\Pi\) is substituted into Eqs. (2.1) and (2.2) to give the components of the EMF generated by an electric elemental dipole placed in the Cartesian coordinate system as shown in Fig. 2.4.

The components are:
\[ E_R = \frac{p}{2\pi \varepsilon} \left( \frac{1}{R^3} - \frac{ik}{R^2} \right) \cos \theta e^{-jkR} \] 

(2.21)

**Table 2.2 Specificity of the Near-Field EMF Measurements**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Near field</th>
<th>Far field</th>
</tr>
</thead>
<tbody>
<tr>
<td>Measured EMF component</td>
<td>E, H, and S</td>
<td>E or H, and S on microwaves</td>
</tr>
<tr>
<td>Other magnitudes measurement</td>
<td>I, T, (SA, SAR) &quot;HESTIA&quot;</td>
<td>Unnecessary</td>
</tr>
<tr>
<td>Spatial components</td>
<td>3</td>
<td>1 or 2</td>
</tr>
<tr>
<td>Polarization</td>
<td>Quasi-ellipsoidal</td>
<td>Linear or elliptical</td>
</tr>
<tr>
<td>Environment</td>
<td>Complex, multipath propagation, and interference</td>
<td>Usually simple</td>
</tr>
<tr>
<td>Frequency spectrum</td>
<td>Wide, often unknown, many fringes</td>
<td>Usually single frequency</td>
</tr>
<tr>
<td>Antennas</td>
<td>Small, omnidirectional</td>
<td>Resonant, directional</td>
</tr>
<tr>
<td>Temporal and spatial EMF alternations</td>
<td>Significant</td>
<td>Usually negligible</td>
</tr>
<tr>
<td>Uncertainty</td>
<td>3, 6 or more dB</td>
<td>Around 1 dB</td>
</tr>
<tr>
<td>Temperature sensitivity</td>
<td>Significant</td>
<td>Unessential</td>
</tr>
<tr>
<td>Susceptibility</td>
<td>Significant</td>
<td>Omittable</td>
</tr>
<tr>
<td>Influence of surroundings</td>
<td>Significant</td>
<td>Usually omittable</td>
</tr>
<tr>
<td>Procedures</td>
<td>Complex</td>
<td>Simple</td>
</tr>
<tr>
<td>Agreement with theory</td>
<td>Reasonable</td>
<td>Good</td>
</tr>
<tr>
<td>Measured levels</td>
<td>V/m, kV/m</td>
<td>mV/m, pV/m, dBuV/m</td>
</tr>
</tbody>
</table>

\[ E_\theta = \frac{p}{4\pi \varepsilon} \left( \frac{1}{R^3} - \frac{ik}{R^2} \right) \sin \theta e^{-jkR} \] 

(2.22)

\[ H_\phi = \frac{j\omega p}{4\pi} \left( \frac{1}{R^2} - \frac{ik}{R} \right) \sin \theta e^{-jkR} \] 

(2.23)

Where \( p = \) the dipole moment:

\[ p = \left| p \right| = \left| \int \frac{dz}{j} 1_z \right| \] 

(2.24)
and

\[ I_z = \text{the current in the dipole} \]
\[ 1_z = \text{the versor (unit vector) of axis } z \]

If we repeat similar procedures for \( J = 0 \) and \( |* J| = *J_z = \text{constant} \), we will have formulas defining the spatial field components of the elemental magnetic dipole. Using the principle of duality, the formulas may be immediately obtained from Eq. (2.21), (2.22), and (2.23) by the exchange of \( E \) to \( H \) and \( e \) to \( \mu \) and conversely.

The maximal spatial variations of the EMF components, as a function of distance to a point of observation, may be written in the following forms:

- For an elemental electric dipole:

\[
E = C_1(\alpha) \frac{\exp(-jkR)}{R^\alpha} \tag{2.25}
\]

While

\[ 3 \geq \alpha \geq 0 \tag{2.26} \]

For \( 0 < R < \infty \)

where \( C_1(\alpha) = \text{a constant dependent of } R \), and \( E = |E| \) is the complex amplitude of the electric field strength.
• And by analogy:

For an elemental magnetic dipole:

\[
H = C_2(\alpha) \frac{\exp(-jkr)}{R^\alpha}
\]

(2.27)

where \( C_2(\alpha) \)—a constant dependent of \( R \), and \( H = |H| \) is the complex amplitude of the magnetic field strength, while \( a \) and \( R \) fulfill the conditions of Eq. (2.26).

Equations (2.25) and (2.27) illustrate the curvature of the EMF and are not related to any particular field zone. The radius of curvature is proportional to \( R^{\alpha+2} \) and varies from infinity (for the plane wave) to very small magnitudes in the near field.

If we substitute \(|J| = J_z \sin k(h - |z|)\) and \(*J = 0\) in Eqs. (2.3) and (2.4) within the limits \(-h < z < +h\) (where \( h \) is the length of the dipole arm), after calculation of \( \Pi \) using Eqs. (2.1) and (2.2), we will find components of the EMF from an infinitely thin, symmetrical dipole antenna of total length \( 2h \) (Fig. 2.5). As a result of the sinusoidal current distribution assumption in the dipole, unlike the “ideal” assumption of a constant current distribution, a certain error is expected. The error is especially important for infinitesimally thin (one-dimensional) dipoles.

![Figure 2.5 Symmetrical dipole in coordinate system.](image)

The assumption

However, the assumption is fully acceptable while the radiation pattern of such antenna is being considered. The use of precise solutions of the integral equations, describing current distribution in a real antenna, is not necessary in the aspect of the EMF components’ strength calculations in the area of interest as well as in the light of the final conclusion of the considerations presented [6]. Here we will use the results of calculations available in the literature [1]. The EMF components \( E_z \), \( E_\rho \), and \( H_\phi \) are given by Eqs. (2.28), (2.29), and (2.30) respectively:
\[ E_z = -\frac{jZI_o}{4\pi \sin \kh} \left[ \frac{\exp(-jkR_1)}{R_1} + \frac{\exp(-jkR_2)}{R_2} - 2\cos \kh \frac{\exp(-jkR_o)}{R_o} \right] \]

\[ E_r = \frac{jZI_o}{4\pi \sin \kh} \left[ \frac{\exp(-jkR_1)}{R_1} (z-h) + \frac{\exp(-jkR_2)}{R_2} (z+h) - 2\cos \kh \frac{\exp(-jkR_o)}{R_o} \right] \]

\[ H_r = \frac{jI_o}{4\pi \sin \kh} \left[ \exp(-jkR_1) + \exp(-jkR_2) - 2\cos \kh \exp(-jkR_o) \right] \]

where \( I_o = \) current intensity at the dipole input,

\[ R_1 = \sqrt{\rho^2 + (z-h)^2} \]
\[ R_2 = \sqrt{\rho^2 + (z+h)^2} \]
\[ R_o = \sqrt{\rho^2 + z^2} \]

and other indications as in Fig. 2.5.

If we continue here considerations similar to the above, in the case of the electric and magnetic elemental dipoles, and generalizing on the basis of Eqs. (2.28–2.30), per analogy to Eqs. (2.25) and (2.27), we can write the relations describing EMF variations in the proximity of the symmetrical dipole:

\[ E = C_3 \alpha \frac{\exp(-jkR)}{R^\alpha} \]

\[ H = C_4 \alpha \frac{\exp(-jkR)}{R^\alpha} \]

where:
\[ 1 \geq \alpha \geq 0 \]

\( C_3(\alpha) \) and \( C_4(\alpha) \) = constants dependent of \( R \),

\( R = \) the distance to a point of observation;

while for EMF close to the antenna surface, one may assume \( R \approx \rho \)

If in Eqs. (2.25) and (2.27) we substitute \( \alpha = 3 \), we will have a relationship defining the EMF variations as a function of distance in the near field of elemental dipoles where \( \alpha = 2 \) represents the intermediate-field of the dipoles. The far field of the elemental dipoles and the near field of a thin symmetrical dipole antenna are characterized by \( \alpha = 1 \).

The variability of the later versus distance is specific to a spherical wave. Rigorous analysis of Eqs. (2.21–2.23) and (2.28–2.30) does not justify an assumption in the formulas that \( \alpha = 0 \) for \( R \rightarrow \infty \), which would represent the plane wave. Such a simplification is often accepted when an EMF in a limited area, sufficiently far from a source, is being considered. In that area, amplitude variations of \( E \) and \( H \) vectors in any direction are negligibly small.

The simplification is equivalent to the assumption that the radius of curvature of the field considered is equal to infinity. The maximal phase variations are independent on \( \alpha \) if one assumes \( \alpha = \text{const} \); such a case is most interesting from the point of view of metrological practice.

The comparison of Eqs. (2.25) and (2.27) as well as (2.31) and (2.32) permits us to come to the conclusion that the EMF “variability” in proximity to sources much smaller in the comparison to the wavelength \( (\alpha = 3) \) is the largest. Thus, if we estimate the errors of the EMF measurements near the sources, we will have majorants of the errors for an arbitrary source.

The conclusion is, in some sense, an intuitive one, and it is a result of the presence of the quasi-stationary field in proximity to sources whose sizes are comparable or larger than the wavelength (induction field). A particular example of the case is the EMF in the proximity of AC power devices and especially around overhead transmission lines. Usually, only Coulomb’s and Biot-Savart’s laws are applied, and this approach is equivalent to the assumption that the EMF does not exist, and the field is sufficiently represented by \( E \) and \( H \) fields only.

Doubt may arise, under these considerations, relating to the presence of higher powers of \( a \) when a multipole expansion is applied. The approach may make more precise calculations of EMF generated by elemental sources more possible. However, even if appropriate corrections are applied, it does not make a substantial difference to the maximum sources are considered. It should be emphasized here that only physical sources have a practical importance because of the efficiency of the EM energy radiation.

Good examples here are formulas describing the standard EMF near, for instance, a standard loop antenna. In this case, apart from the finite sizes of the antenna that are much larger than the elemental dipole, no term exceeds a power of 3. This is a matter of practical importance [7]. Having analyzed the near- and far-field terms, the next chapter considers methods of measurement.
References


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