

Chapter 2

Electric and Magnetic Fields from Simple Circuit Shapes

If one wants to avoid empirical recipes and the “wait and see if it passes” strategy, the calculation of radiated fields from electric circuits and their associated transmission cables is of paramount importance to proper EMI control. Unfortunately, precisely calculating the fields radiated by a modern electronic equipment is a hopeless challenge. In contrast to a CW transmitter, where the radiation source characteristics (e.g., transmitter output, antenna gain and pattern, spurious harmonics, feeder and coupler losses, etc.) are well identified, a digital electronic assembly, with its millions of input/output circuits, printed traces, flat cables, and so forth, is impossible to mathematically model with accuracy, at least within a reasonable computing time by today’s state of the art. The exact calculation of the E and H fields radiated by a simple parallel pair excited by a pulse train is already a complex mathematical process.

However, if we accept some drastic simplification, it is possible to establish an order of magnitude of the field by using fairly simple formulas. Such simplification includes:

1. Retaining only the value of the field in the optimum direction
2. Having the receiving antenna aligned with the maximum polarization
3. Assuming a uniform current distribution over the wire length, which can be acceptable by using an average equivalent current instead of the maximum value
4. Ignoring dielectric and resistive losses in the wires or traces

The formulas described hereafter were derived by S. Schelkunoff [5] from more complex equations found in the many books on antenna theory. They allow us to resolve most of the practical cases, which can be reduced to one of the two basic configurations:

1. The closed loop (i.e., magnetic excitation)
2. The straight open wire (i.e., electric excitation)

2.1 FIELD RADIATED BY A LOOP

An electromagnetic field can be created by a circular loop carrying a current I (Fig. 2.1). Assuming that:

- I is uniform along the loop.
- There is no impedance in the loop other than its own reactance.
- The loop size is $\ll \lambda$.
- The loop size is $< D$, the observation distance.
- The loop is in free space, not close to a metallic surface.

E and H can be found by using the simple solutions that Schelkunoff derived from Maxwell's equations. Replacing some terms by more practical expressions:

$$H_r \text{ A/m} = \frac{IA}{\lambda} \left[\frac{j}{D^2} + \frac{\lambda}{2\pi D^3} \right] \cos \sigma \quad (2.1)$$

$$H_\sigma \text{ A/m} = \frac{\pi IA}{\lambda^2 D} \sqrt{1 - \left(\frac{\lambda}{2\pi D} \right)^2 + \left(\frac{\lambda}{2\pi D} \right)^4} \sin \sigma \quad (2.2)$$

$$E_\phi \text{ V/m} = \frac{Z_0 \pi IA}{\lambda^2 D} \sqrt{1 + \left(\frac{\lambda}{2\pi D} \right)^2} \sin \sigma \quad (2.3)$$

where:

I = loop current, in amperes

A = loop area in m^2

λ = wavelength in meters = $300/F_{(\text{MHz})}$

D = distance to observation point, in meters

Z_0 = free space impedance = 120π or 377Ω

Comparing this with Fig. 2.1, we see that for $\sigma = 0$, E_ϕ and H_ϕ are null ($\sin \sigma = 0$), while H_r is maximum ($\cos \sigma = 1$). Except near the center of a solenoid or a transmitting loop antenna, this H_r term in the Z -axis direction is of little interest because it vanishes rapidly, by its $1/D^2$ and $1/D^3$ multipliers. Notice also that there is no E_r term.

To the contrary, in the equatorial plane, for $\sigma = \pi/2$, H_r is null, and E_ϕ , H_ϕ get their maximum value. So from now on, we will consider systematically this worst-case azimuth angle.

Looking at Equ. (2.2) and Equ. (2.3) and concentrating on boundary conditions, we see two domains, near field and far field, plus a transition region.

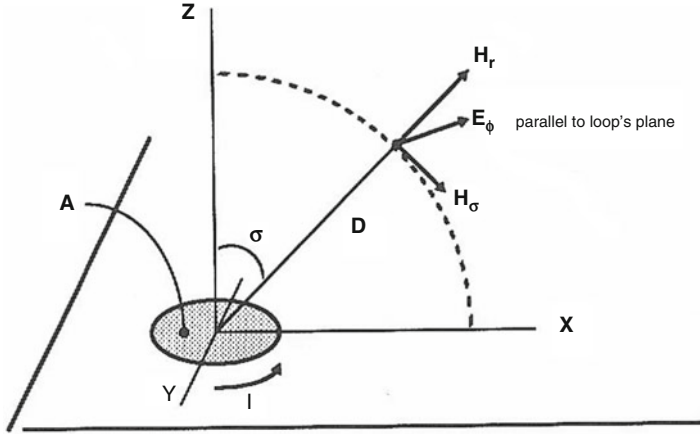


Fig. 2.1 Radiation from a small magnetic loop

Near Field: For $\lambda/2\pi D > 1$, i.e., $D < \lambda/2\pi$ or $D < 48/F(\text{MHz})$

Under the square root in Equ. (2.2) and Equ. (2.3), the larger terms are the ones with the higher exponent. Thus, neglecting the other second- or third-order terms, we have:

$$H_{A/m} = \frac{IA}{4\pi D^3} \quad (2.4)$$

$$E_{V/m} = \frac{Z_0 IA}{2\lambda D^2} \quad (2.5)$$

We remark that H is independent of λ , i.e., independent of frequency: the formula remains valid down to DC. H falls off as $1/D^3$. E increases with F and falls off as $1/D^2$.

In this region called near-field or induction zone, fields are strongly dependent on distance. Any move toward or away from the source will cause a drastic change in the received field. Getting ten times closer, for instance, will increase the H field strength 1,000 times.

Since dividing volt/m by amp/m produces ohms, the E/H ratio, called the wave impedance for a radiating loop, is

$$Z_w(\text{near loop}) = Z_0 \frac{2\pi D}{\lambda} \quad (2.6)$$

When D is small and λ is large, the wave impedance is low. We may say that in the near field, Z_w relates to the impedance of the closed loop circuit which created the field, i.e., almost a short. As D or F increases, Z_w increases.

Far Field: For $\lambda/2\pi D < 1$, i.e., $D > \lambda/2\pi$, or $D > 48/F(\text{MHz})$

The expressions under the square roots in Equ. (2.2) and Equ. (2.3) are dominated by the terms with the smallest exponent. Neglecting the second- and third-order terms, only the “1” remains, so:

$$H_{A/m} = \frac{\pi I A}{\lambda^2 D} \quad (2.7)$$

$$E_{V/m} = \frac{Z_o \pi I A}{\lambda^2 D} \quad (2.8)$$

In this region, often called the *far-field*, *radiated-field*, or *plane wave region*¹, both E and H fields decrease as $1/D$ (see Fig. 2.2). Their ratio is constant, so the wave impedance is

$$Z_w = E/H = 120\pi \text{ or } 377\Omega$$

This term can be regarded as a real impedance since E and H vectors are in the same plane and can be multiplied to produce a radiated power density, in W/m^2 . E and H increase with F^2 , an important aspect that we will discuss further in our applications.

Transition Region: For $\lambda/2\pi D = 1$ or $D = 48/F(\text{MHz})$

In this region, all the real and imaginary terms in field equations are equal, so all terms in $1/D$, $1/D^2$, and $1/D^3$ are equal and summed with their sign. This zone is rather critical because of the following:

1. With MIL-STD-461 testing (RE102 test for instance), the test distance being 1 m, the near-far-field transition is taking place around 48 MHz, which complicates the prediction.
2. Speculations concerning the wave impedance are hazardous due to very abrupt changes caused by the combination of real and imaginary terms for E and H .

¹ However, “plane wave” does not have exactly the same meaning. Another condition is governing the near- or far-field situation that is related to the physical length of the antenna. If l , the largest dimension of the radiating element, is not small compared to distance D , another near-field condition exists due to the curvature of the wavefront. To have less than 1 dB (11%) error in the fields calculated by Equ. (2.7) and Equ. (2.8), another requirement stipulates that $D > \ell^2/2\lambda$. However, if the dimension of the radiating element is less than $\lambda/2$, both length and distance far-field conditions are met for D (far field) $> \lambda/2\pi$.

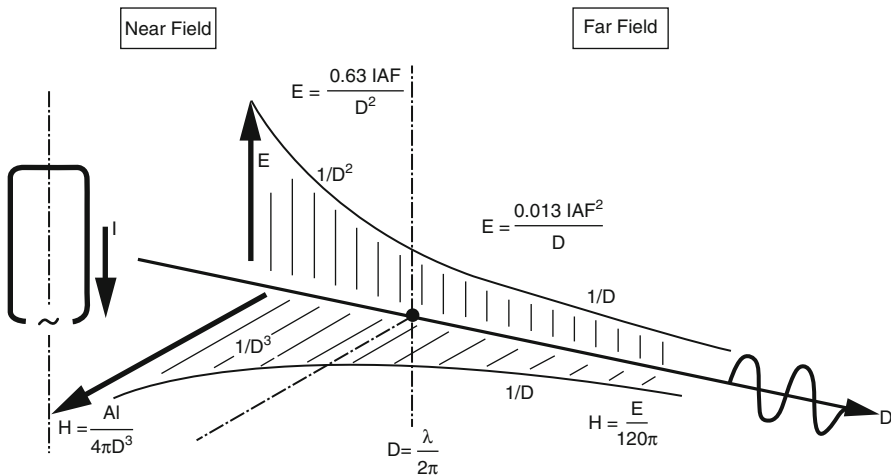


Fig. 2.2 E and H fields from a perfect loop

2.2 FIELDS RADIATED BY A STRAIGHT WIRE

It does not take a closed loop to create an electromagnetic field. A straight wire carrying a current, I , creates an electromagnetic field (most radio communication antennas are wire antennas). The practical difficulty is that, in contrast to the closed loop, it is impossible to realize an isolated dipole with a DC current: only AC current can circulate in an open-wire self-capacitance. Fields generated from a short, straight wire are shown in Fig. 2.3.

E and H can be derived from Maxwell's equations with the same assumptions as the elementary loop, i.e.:

- Current I is uniform.
- The wire length $\ll \lambda$.
- The wire length $< D$, the observation distance.
- The wire is in free space, not close from a ground plane.

Using Schelkunoff's solutions for a small electric dipole, expressed in more practical units,

$$E_r = 60I\ell \left(\frac{1}{D^2} - \frac{j\lambda}{2\pi D^3} \right) \cos \sigma \quad (2.9)$$

$$E_\sigma = \frac{Z_0 I \ell}{2\lambda D} \sqrt{1 - \left(\frac{\lambda}{2\pi D} \right)^2 + \left(\frac{\lambda}{2\pi D} \right)^4} \sin \sigma \quad (2.10)$$

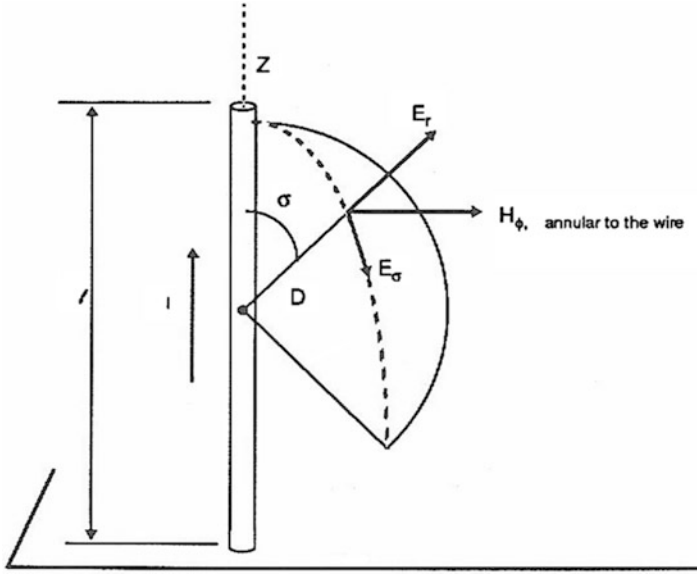


Fig. 2.3 E and H fields from a small, straight wire

$$H_{\phi} = \frac{I\ell}{2\lambda D} \sqrt{1 + \left(\frac{\lambda}{2\pi D}\right)^2} \sin \sigma \quad (2.11)$$

where:

I = wire current, in amperes

ℓ = dipole length in meters

λ = wavelength in meter, $300/F_{\text{MHz}}$

D = distance to observation point in meters

Z_0 = free-space impedance, 120π or 377Ω

As for the loop, we remark that for $\sigma = 0$, E and H_{ϕ} are null ($\sin \sigma = 0$) while E_r is maximum ($\cos \sigma = 1$). E_r , in the axis of the wire, is of little interest because it drops off rapidly, as $1/D^2$, $1/D^3$

In the equatorial plane for $\sigma = \pi/2$, E and H have their maximum values. From now on we will consider this worst-case azimuth angle. In fact, for $\sigma = 90^\circ \pm 25^\circ$, the error would be less than 10%. As for the loop, we can see two domains plus a transition region.

Near Field: For $\lambda/2\pi D > 1$, i.e., $D < \lambda/2\pi$ or $D < 48/F(\text{MHz})$

As for the loop, the terms with the higher exponent prevail, under the square root.

Neglecting the other second or third terms,

$$H_{A/m} = \frac{I\ell}{4\pi D^2} \quad (2.12)$$

$$E_{V/m} = \frac{Z_o I \ell \lambda}{8\pi^2 D^3} \quad (2.13)$$

Here again, we remark that H annular around the dipole is independent of F . This formula holds down to DC, where it equals the well-known result of the Biot and Savart law for a small element. This time, it is to H to fall off as $1/D^2$ while E falls as $1/D^3$. Both are strongly dependent on distance. Since, for a current kept constant, E decreases when F increases, the wave impedance decreases when D or F increases:

$$Z_w = \frac{E}{H} = Z_0 \frac{\lambda}{2\pi D} \quad (2.14)$$

As for the loop, Z_w near the source relates to the source impedance itself which, this time, becomes infinite when F gets down to DC.

Far Field: For $\lambda/2\pi D < 1$ (i.e., $D > \lambda/2\pi$ or $D > 48/F(\text{MHz})$)

The terms with higher exponents can be neglected under the square root, so

$$H_{A/m} = \frac{I\ell}{2\lambda D} \quad (2.15)$$

$$E(V/m) = Z_0 \times I \times \ell(2\lambda D) \quad (2.16)$$

Both E and H decrease as $1/D$. This ratio, as for the loop in far field, remains constant:

$$E/H = Z_0 = 120\pi \text{ or } 377 \Omega$$

It is worth noticing that for a single wire in free space, E and H increase as F (instead of F^2 for the loop).

Transition Region: For $\lambda/2\pi D < 1$ or $D = 48 F(\text{MHz})$

The same remarks apply as for the loop.

Figure 2.4 summarizes the evolution of Z_w for wires and loops as D/λ increases.

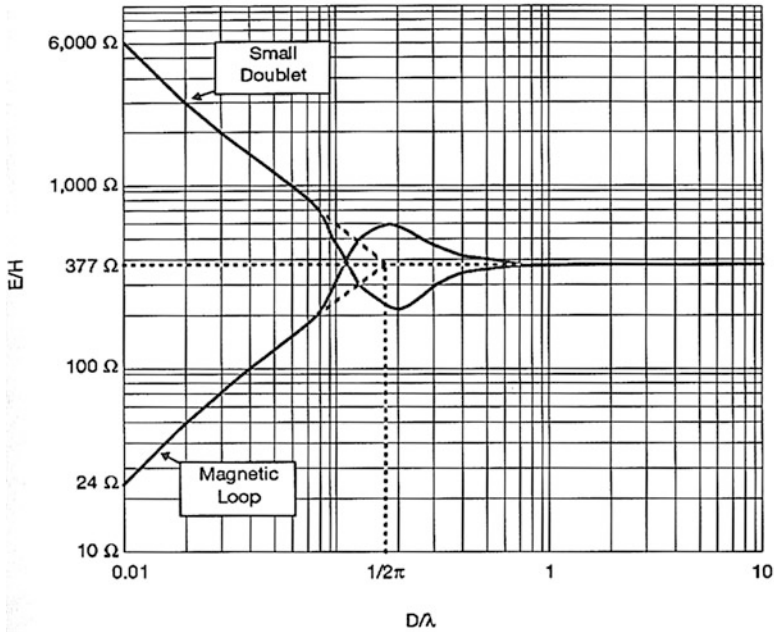


Fig. 2.4 Wave impedance vs. distance/wavelength

2.3 EXTENSION TO PRACTICAL, REAL-LIFE CIRCUITS

Although theoretically correct, ideal loop or doublet models have a limited practical applicability in EMC due to the restrictions associated with their formulas:

1. Distance D should be large compared to the circuit dimensions.
2. The circuit length should be less than $\lambda/2$, and preferably less than $\lambda/10$, for the assumption of uniform current to be acceptable.
3. The single-wire model corresponds ideally to a piece of wire floating in the air, in which a current is forced, a situation seldom seen in practice.
4. The single-wire model assumes that the circuit impedance is infinite in near field or at least larger than the wire reactance alone; this condition is rarely met except in dipole or whip antennas.
5. Restrictions 3 and 4 seem alleviated if one switches to the loop model. The loop, indeed, is a more workable model for practical, non-radio applications because it does not carry the premise of a wire coming from nowhere and going nowhere. But it bears a serious constraint, too: the loop must be a short circuit, such as the wave impedance, and hence the E field is only dictated by the coefficients in Maxwell's equational solutions. If this condition is not met (it is seldom met

except in the case of a coil with only one or few turns and no other impedance), the H field found by Equ. (2.6) and Equ. (2.7) will be correct, but the actual associated E field will be greater than its calculated value.

In reality, we deal with neither purely open wires nor perfect loops, but with circuit configurations which are in between. Therefore, predictions in the near field would produce:

- An E field higher than reality, if based on open-wire model (pessimistic error)
- An E field lower than reality, if based on ideal loop model (optimistic error)

Measurements have proven that the latter can cause underestimates as large as 60 dB or more. Therefore, certain adjustments need to be made. Assuming these adjustments, the modified equations and models can be usable by the designer for most of the actual circuits and cable configurations encountered, like the one in Fig. 2.5.

2.3.1 Fields Radiated by Actual Conductor Pairs

The core of this modeling is the “modified single-wire model” where, instead of a straight wire or circular loop, we have a more practical vehicle, where an area $\ell \times s$ can be treated by the loop equation or regarded as two single wires with a radiation phase shift equal to $\sin(2\pi s/\lambda)$. Depending on the circuit impedance, we will use one or the other, as explained next. The basis for the simplification is that, in near field, the wave impedance $Z_w = E/H$ is “driven” by the circuit impedance Z_c every time this circuit impedance is in between an ideal dipole that creates a high Z_w and an ideal loop creating a low Z_w .

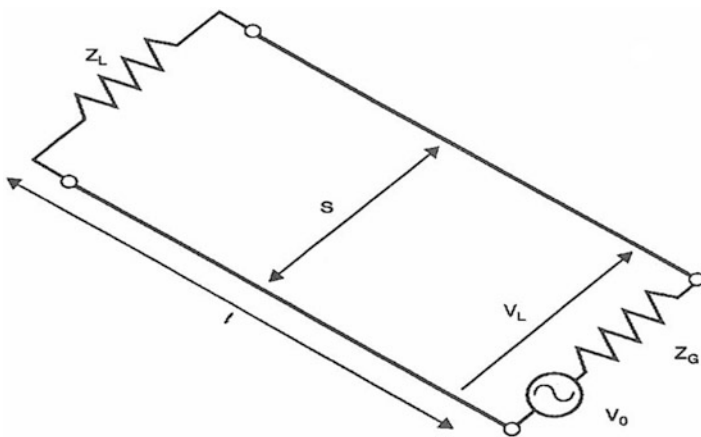


Fig. 2.5 The modified single-wire model

In the near field, given the total circuit impedance Z_c (wiring plus load),

$$Z_c = Z_g + Z_L$$

A. If $Z_c > 7.9 Dm \times F(MHz)$, we will use the modified wire model:

$$E_{V/m} = \frac{VA}{4\pi D^3} \quad (2.17)$$

where,

V = driving voltage in volts (actual line voltage, not the open circuit voltage)

A = circuit area $\ell \times s$, in m^2

D = observation distance, in m

Very often, more practical units are welcome:

$$E_{(\mu V/m)} = \frac{7.96 VA}{D^3} \quad (2.17a)$$

for V in volts, A in square centimeters, and D in meters.

B. If $Z_c < 7.9 D \times F(MHz)$, we will use the ideal loop formulas, since the circuit impedance is low enough for this model to hold.

$$E_{V/m} = \frac{0.63IAF_{MHz}}{D^2} \quad (2.18)$$

for current I in amperes, A in square meters, and D in meters. Or, using more convenient units:

$$E_{(\mu V/m)} = \frac{63IAF_{MHz}}{D^2} \quad (2.18a)$$

for current I in amperes, A in square centimeters, and D in meters.

Because, with such low-impedance radiating circuits, this is the H field which is of concern, we can employ a straightforward application of Equ. (2.4):

$$H(A/m) = I \times A / (4\pi D^3) \quad (2.18b)$$

for A in square meters and D in meters. Notice that this near-field expression is exactly the mirror image of the high-impedance loop E field in Equ. (2.17). The product $I \times A$ is often referred to as the “magnetic moment.” Again, using more convenient units,

$$H_{(\mu\text{A/m})} = \frac{7.96 \times I \times A}{D^3} \quad (2.19)$$

for current (I) in amperes, A in square centimeters, and D in meters. Therefore a *magnetic moment* is defined, in $\text{amp} \times \text{cm}^2$, abbreviated as A-cm^2 .

In the far field, regardless of the type of excitation (i.e., circuit impedance), E and H are given by Equ. (2.7) and Equ. (2.8), which we will reformulate in terms of frequency rather than wavelength:

$$E_{\text{V/m}} = \frac{0.013 V A F_{\text{MHz}}^2}{D \times Z_c} \quad (2.20)$$

$$H_{\text{A/m}} = \frac{E}{120\pi} = \frac{35.10^{-6} \times I \times A \times F_{\text{MHz}}^2}{D} \quad (2.21)$$

with V in volts, I in amperes, A in square meters, and D in meters.

Or, once again using more practical units of measurement,

$$E_{(\mu\text{V/m})} = \frac{1.3}{D} \times \frac{V}{Z_c} \times A \times F_{\text{MHz}}^2 \quad (2.22)$$

for V in volts, A in square centimeters, and D in meters.

At this point, a few remarks are in order:

1. We now have an expression for E fields that can be calculated by entering the drive voltage, which often is more readily known to the circuit designer than the current.
2. Except for very low-impedance loops (less than 7.9Ω at 1 MHz, less than $7.9 \text{ m}\Omega$ at 1 kHz), i.e., low-voltage circuits carrying large sinusoidal or pulsed currents, it is generally the wire pair model Equ. (2.17) and Equ. (2.19) that applies.
3. *In the near field*, for all circuits except low-impedance loops, E is independent of frequency and remains constant with V . At the extreme, if Z_c becomes extremely large, current I becomes extremely small but Z_w increases proportionally, keeping E constant when F decreases down to DC (more details in Appendix A “Modified dipole model”).
4. *In the far field*, radiation calculated for a two-wire circuit (the single-dipole formula times the weighing factor $\sin 2\pi s/\lambda$ due to the other wire carrying an opposite current) would reach exactly the same formula as the one for a radiating circular loop. Therefore, as long as its dimensions are $\ll \lambda$, the actual circuit shape has virtually no effect on the radiated field in the optimum direction. Only its area counts.
5. For $\ell \geq \lambda/4$, the circuit begins to operate like a transmission line or a folded dipole. Current is no longer uniform, and in expressing A , the length ℓ must be

clamped to $\lambda/4$, i.e., ℓ (m) is replaced by $75/F_{\text{MHz}}$. In other words, the active part of this fortuitous antenna will “shrink” as F increases. Furthermore, if the circuit does not terminate in its matched impedance, there will be standing waves, and the effective circuit impedance will vary according to transmission line theory. The radiation pattern will exhibit directional lobes.

6. When separation s is not $\ll \ell$, i.e., the loop is not a narrow rectangle but is closer to a square, the upper bound is reached [1] when $(\ell + s) = \lambda/4$, i.e., $F_{\text{max}} = 7,500/(\ell + s)$, for F in MHz and ℓ, s in centimeters. Furthermore,

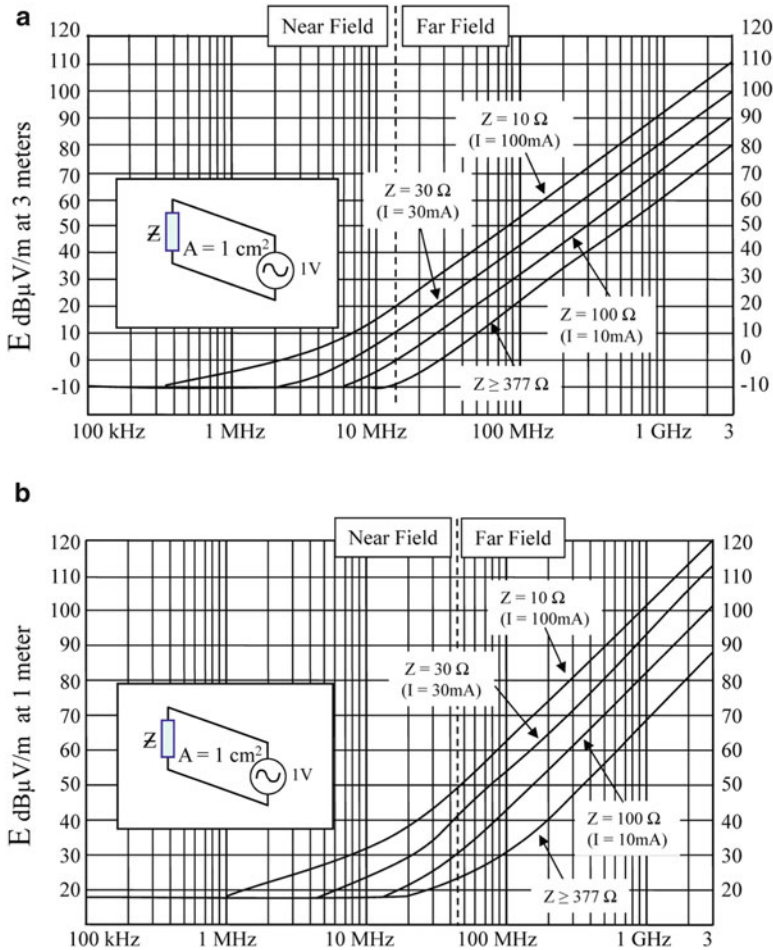


Fig. 2.6 (a) E field at 3 m from a 1 cm^2 loop, driven by 1 V. For other voltages and $(\ell \times s)$ values, apply correction: $20 \log V + 20 \log(\ell \times s)$. (b) E field at 1 m from a 1 cm^2 loop, driven by 1 V. For other voltages and $(\ell \times s)$ values, apply correction: $20 \log V + 20 \log(\ell \times s)$

- with conductors partly in dielectric, the velocity is reduced by a factor of $1/\sqrt{(1 + 0.5\epsilon_r)}$. For PVC or Mylar cables, $F_{\max} = 5,300/\ell(\text{cm})$. For PCB traces, $F_{\max} = 4,400/\ell(\text{cm})$. So an average value $F_{\max} = 5,000/\ell(\text{cm})$ could be retained, above which the physical length should be replaced by $5,000/F(\text{MHz})$.
7. In the far field, if $Z_c > 377 \Omega$, the value of 377Ω must be entered in Equ. (2.20). This acknowledges the fact that an open-ended circuit will still radiate due to the displacement current.
 8. In the far field, E increases as F^2 for a loop or a pair. This is a very important effect that we will address in the application part of this book. Equations (2.17a) to (2.22) are plotted in Fig. 2.6a–c for a “unity” electric pair of $1 \text{ V}\cdot\text{cm}^2$ and a unity magnetic moment of $1 \text{ A}\cdot\text{cm}^2$. They show E or H at typical test d distances.

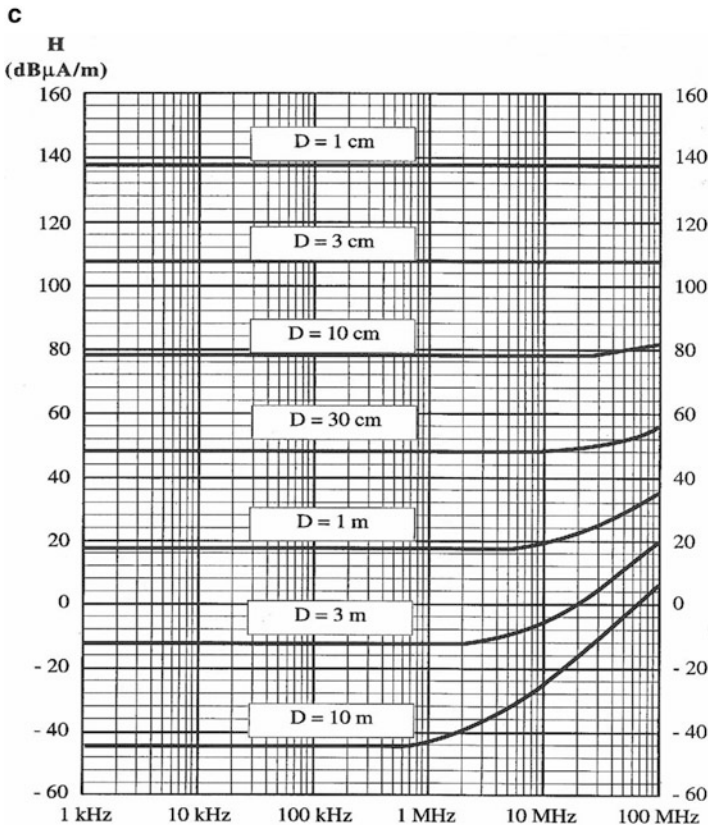


Fig. 2.6 (continued) (c) Magnetic field from a 1 amp, 1 cm^2 loop. For other currents and areas, apply correction: $20 \log [I \times A(\text{cm}^2)]$

2.3.2 Fields Radiated by a Wire or Trace Above a Ground Plane

A frequent configuration is that of a single wire or PCB trace above a conductive plane acting as return or reference conductor. From image theory, a wire at a height h above a ground plane will radiate the same E field as a wire pair with separation $2h$, driven by the same voltage [1, 4]. Therefore, one can conclude that given a same voltage/current combination, the field radiated by any conductor at a distance h above a plane could be calculated from Equ. (2.17a), Equ. (2.18a), or Equ. (2.22) simply by using $2h$ for the loop dimension. Yet, one must not forget that image theory is only valid for an infinite plane, which is seldom the case with a PCB, as will be explained next.

2.3.2.1 Finite Plane

As shown on Fig. 2.7, the finite plane allows the return current to radiate a magnetic field that surrounds the PCB, hence it radiates in all directions, that is a 360° solid angle (or 4π steradians) and not just in the half space. Close to the wire and plane, the field contour is approximately the same as if the plane was infinite. But for an observation point P , at a distance D much larger than the PCB dimensions, the trace above ground merely radiates like a 2-conductor pair with height h , the return conductor being the plane itself.

2.3.2.2 Quasi-Infinite Plane

However, there are cases where image-plane conditions actually exist:

- Typical radiated emission (RE) tests of an equipment with its external cables are performed at 1, 3, or 10 m, with a conductive ground surface (semi-anechoic room or open-area test site) extending far beyond the test setup. In this case the doubling of the field by the image mechanism is real at certain frequencies (see Sect. 2.5.2).
- When calculating E or H field in close proximity of a chip or trace, if the distance D is \ll than PCB dimensions, we can accept the finite plane conditions and use $2h$ for the loop size.

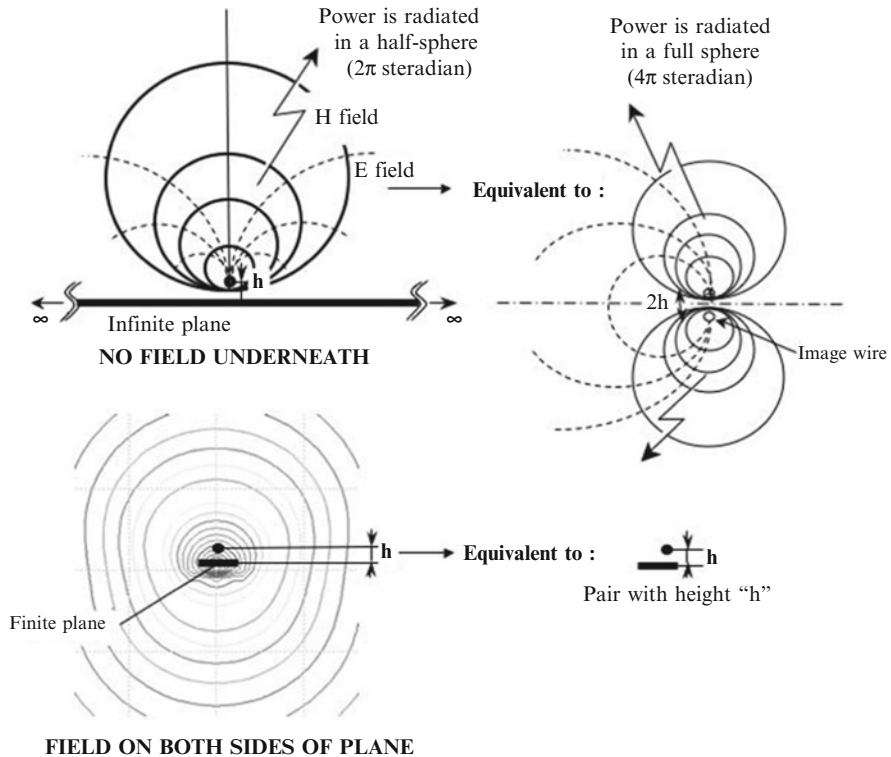


Fig. 2.7 Image mechanism with infinite and finite ground planes

2.4 DIFFERENTIAL-MODE RADIATION FROM SIMPLE CIRCUITS

The simplest radiating configuration we will encounter in practice is the small differential-mode radiator, whose largest dimension, ℓ , is smaller than both the observation distance, D , and the height above ground. Such circuits (see Fig. 2.8) are found with:

- PCB traces, truly differential or microstrip (one trace above ground plane)
- Wire wrapping or any hard-wired board or backplane
- Ribbon cables
- Discrete wire pairs (for $\ell \ll D$)

The culprit source exciting such circuits can be a digital or analog signal, a switching transistor, a relay, a motor creating transient spikes, etc. There is also a possibility that the differential pair is simply a carrier of an EMI signal that has been generated in the vicinity and coupled to it through power supply conduction or nearby crosstalk.

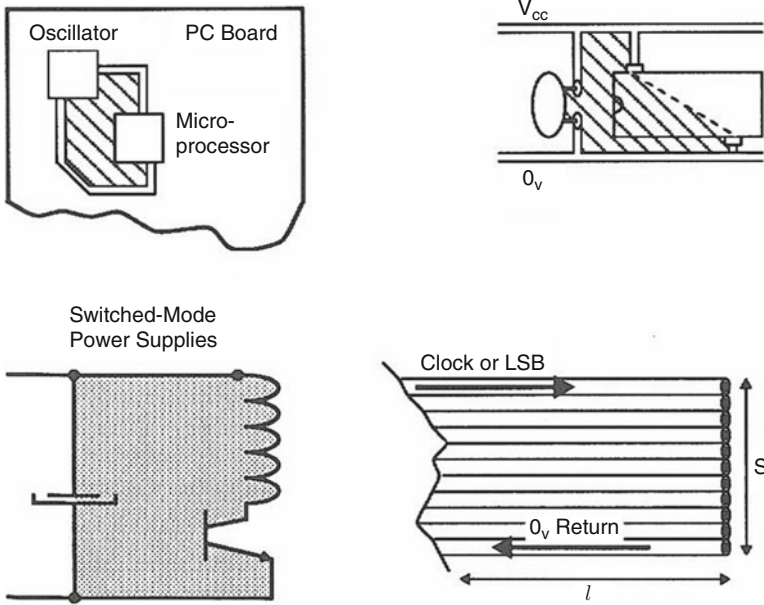


Fig. 2.8 A few typical differential-mode radiators

The procedure is then as follows:

1. Determine V_{diff} , I_{diff} at the frequencies of interest and the circuit impedance.
2. Check for test distance $D \geq 48/F_{\text{MHz}}$ (far-field conditions).
3. If far field, use the curves in Fig. 2.6 or Equ. (2.20).
4. If near field, determine if the circuit belongs to the low- Z_c , loop model (for $Z_c < 7.9 \times F \times D$) or to the wire model ($Z_c > 7.9 \times F \times D$).
5. Check if ℓ (cm) $> \lambda/4$ or $5,000/F$. If ℓ is larger, replace ℓ by $5,000/F$ for area correction.
6. Repeat Step 5 for wire or trace separation, s .
7. Calculate $A \text{ cm}^2 = \ell \times s$ and determine area correction, $20 \log A$, using adjustments (5) and (6), if needed. For a trace above a ground plane, do not apply image theory, since the plane is not infinite. Simply use $A = \ell \times h$.
8. Find the E field by: $E \text{ (dB}\mu\text{V/m)} = E_0 \text{ (from curves)} + 20 \log A + 20 \log V$.
9. If H field calculation is desired instead, use the H -field curves and add corrections.

Example 2.1 A video signal crossing a PC board is to be switched to different displays. The carrier is 100 MHz with a line voltage of 10 Vrms. The PCB has the following characteristics:

- Single-sided, one layer (no ground plane)
- Average video trace length, $\ell = 6 \text{ cm}$
- Average distance to ground trace, $s = 0.5 \text{ cm}$

Calculate the E field at 1 m vs. the RE102 limit of MIL-STD-461 when:

1. The circuit is loaded with $75\ \Omega$.
2. The circuit is “on” but standby, open-ended.

We will assume, as a starting condition, that there is no box shielding.

- $V_{\text{diff}} = 20\text{ dBV}$ (loaded) or 26 dBV when open-ended (the voltage will double).
- At 100 MHz , the near-field/far-field transition distance is

$$D_{N-F} = 48/100 = 0.48\text{ m}$$

Therefore, at 1 m we are in far-field conditions. We can use Fig. 2.6 or Equ. (2.22).

- Area correction: $20 \log(6 \times 0.5) = 10\text{ dB}$. The 6 cm length is $< \lambda/4$ at 100 MHz .
- For the $75\ \Omega$ load, we can interpolate between the 30 and $100\ \Omega$ curves. For the open circuit ($Z = \infty$), we will use the curve for $Z \geq 377\ \Omega$.

The calculations steps are detailed below:

Frequency	100 MHz (with $Z = 75\ \Omega$)	100 MHz (open circuit)
E_0 (for $1\text{ V} \times 1\text{ cm}^2$)	44 dB $\mu\text{V/m}$	30 dB $\mu\text{V/m}$
Amplitude correction	20 dBV	26 dBV
Area correct. (cm^2)	10 dB	10 dB
E (final)	74 dB $\mu\text{V/m}$	66 dB $\mu\text{V/m}$
E specification limit	29 dB $\mu\text{V/m}$	29 dB $\mu\text{V/m}$
$\Delta\text{ dB}$	45	37

The specification limit is exceeded by 37-45 dB. Such an attenuation can only be obtained, in practical terms, by using a multilayer board or a single-layer board with a ground plane. This would reduce the radiating loop width by a ten-times factor, i.e., 20 dB, and a correctly designed metal housing to provide at least 25 dB of shielding at 100 MHz . Both solutions will be discussed further in this book.

Example 2.2 A 5 V/20 A switching power supply operates at the basic frequency of 50 kHz . In the secondary loop (formed by the transformer output, the rectifier, and the electrolytic capacitor), the full-wave rectified current spikes have a repetition frequency of 100 kHz and an amplitude of 60 A (peak) on the fundamental. Loop dimensions are $3 \times 10\text{ cm}$. The loop impedance at this frequency is $0.2\ \Omega$. Calculate E and H at 100 kHz for a 1 m distance.

1. The 60 A amplitude corresponds to 36 dBA .
2. At 100 kHz , the near-far transition distance is $D_{N-F} = 48/0.1 = 480\text{ m}$. So, at 1 m, we are in very near field.
3. With $Z_c = 0.2\ \Omega$, we meet the criteria for $Z_c < 7.9 \times F \times 1\text{ m}$.
4. The area correction is $20 \log(30\text{ cm}^2) = 30\text{ dB}$.

We will use the ideal loop model, i.e., Fig. 2.7 or Equ. (2.19) for H field and Equ. (2.18a) for E field. The field is computed as follows:

Frequency	0.1 MHz	0.1 MHz
H_0 (1 A - 1 cm ²)	17 dBμA/m	
E_0 (1 A - 1 cm ²)		16 dBμV/m
Amplitude correction (amp)	36 dB	36 dB
Area correction (cm ²)	30 dB	30 dB
H (final)	83 dBμA/m	
E (final)		82 dBμV/m

Notice that the wave impedance for this predominantly magnetic field at 1 m has a value of

$$Z_w = E/H = 82 \text{ dB}\mu\text{V/m} - 83 \text{ dB}\mu\text{A/m} = -1 \text{ dB } \Omega \text{ or } 0.9\Omega$$

2.5 COMMON-MODE RADIATION FROM EXTERNAL CABLES

External cables exiting an equipment are practically always longer than the size of the equipment box, so it is predictable that they will be major contributors to radiated emissions (just as they would be for radiated susceptibility). Cables radiate by the differential-mode signals that they carry, as discussed in the previous section, but also by the currents circulating in the undesired path, that is, the ground loop.

Ground loop CM (common mode) currents are due to the unbalanced nature of ordinary transmitting and receiving circuits, the imperfect symmetry of the differential links, and, more generally, the quasi-impossibility of avoiding some CM return path, whether the loop is visible (circuit references grounded at both ends to chassis and/or earth) or invisible (floated equipments or plastic boxes). This phenomenon of common-mode excitation of external cables causing radiated emissions is one of the most overlooked one in computers and high-frequency devices interference [2].

The very simple example of Fig. 2.9 shows the unavoidable generation of a CM current. Assume that over a cable length ℓ , the wire pair (untwisted) separation is $s = 3 \text{ mm}$, and the cable height above ground is $h = 1 \text{ m}$. When a signal is sent from equipment #1 to equipment #2, although the designer believes in good faith that current is coming back via the return wire, we see no reason why some of the current (i_3) could not return by the unintended path, i.e., the ground loop.

Currents i_2 and i_3 will split in proportion to the respective impedances. If the wires in the pair are in close proximity, the return impedance by the pair is significantly less than the return impedance via the ground. But less does not mean null. Let us assume that only 10% of the current is returning by the ground loop. The differential-mode (DM) radiation is related to $0.9i \times \ell \times s$. The CM radiation is related to $0.1i \times \ell \times h$. The ratio of the two magnetic moments is

$$\frac{\text{CM}}{\text{DM}} = \frac{(0.1i \times \ell \times 1 \text{ m})}{(0.9i \times \ell \times 0.003 \text{ m})} = 37$$

$$20 \log 37 = 31 \text{ dB}$$

This CM loop, although the corresponding current is regarded as a side effect, radiates 31 dB above the DM loop (notwithstanding that this latter can be further reduced by twisting).

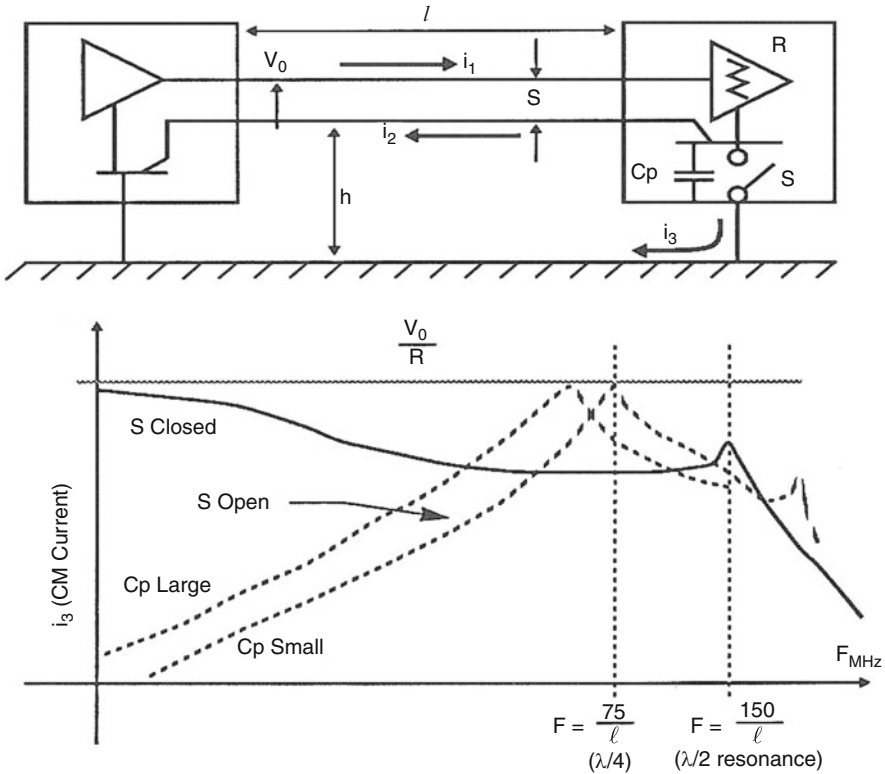


Fig. 2.9 Conceptual view of CM current generation by a differential signal

Opening the switch S , i.e., floating the PCB 0 V reference, would reduce I_{cm} at low frequencies (below a few megahertz for a 10 m cable). In this range, radiated EMI is generally not a concern; but the problem would aggravate at first cable resonance because we now have an oscillatory inductance-resistance-capacitance (LRC) circuit with a high Q (low R). The hump in CM current (Fig. 2.9) depends on the value of C_p , the PCB stray capacitance to chassis. At this occasion, we see that the traditional recipe of grounding the PCB only at one box (star grounding) is useless in the frequency range of most radiation problems. It can even be slightly worse at some specific frequencies.

The undesired CM currents that are found on external cables can be some percentage of the signal currents that would normally be expected on this interface, but, more often, cables are found to carry high-frequency harmonics *that are not at*

all part of the intentional signal (see Fig. 2.10). Rather, they have been picked up inside the equipment by crosstalk, ground pollution, or power supply DC bus pollution. Since the designer did not expect these harmonics, it is only during FCC, CISPR, MIL-STD-461, or other compliance testing that they are discovered.

Figure 2.10 shows the contrast between (a) what is normally expected: power line carries only 50/60 Hz or 400 Hz currents, I/O cable carries a slow serial bus, and the 10 MHz clock is used only internally and (b) what real life provides: I/O pairs or ribbon cable carries 10 MHz residues from the clock, picked up internally; their spectrum extends easily to 200 or 300 MHz. Because of the primary-to-secondary capacitance in the power supply transformer, power wires (phase, neutral, and ground) are also polluted by 10 MHz harmonics. The radiating loops can be ABCD, ABEF, or combinations of all.

Predicting such radiated emissions from external cables will consist of the following:

- Measuring or estimating the CM currents driving the external cables
- Estimating the geometry for the CM-driven antenna
- Applying simple, appropriate antenna formulas (loop or open wire)

These three steps of prediction are examined next.

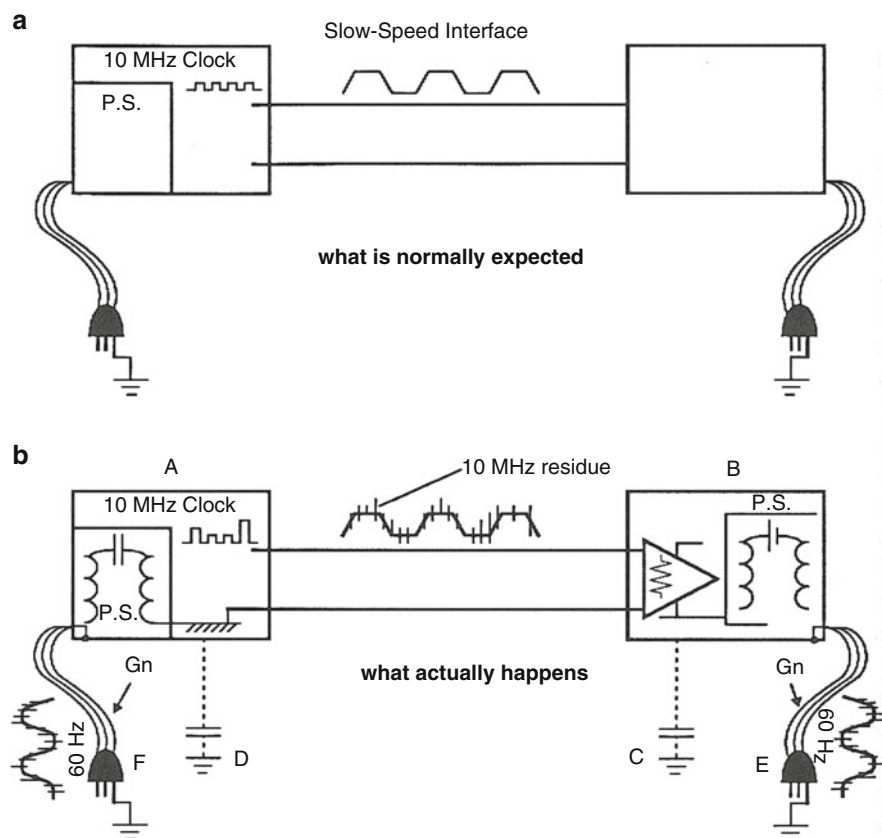


Fig. 2.10 Contamination of external cables by internal HF circuits

2.5.1 How to Estimate CM Currents on Cables

As stated before, the CM currents found on external cables can have essentially two origins. The mechanisms that are causing such currents are analyzed hereafter.

2.5.1.1 CM Current from Intentional Signals

By this, we mean the portion i_3 (generally undesired) of signal current that is returning by the external ground loop. Let us review the three principal cases.

– **Single-Wire Transmission**

The useful signal is carried on a single wire, the chassis, or other structural system ground forming the return path. This type of link is, of course, highly detrimental for EMC and almost never used anymore. Few exceptions are found in automobile, aircraft, or helicopter applications where certain signals are still carried between a hot wire and the vehicle body, to save on copper weight. In this case, the full signal spectrum is driving the single-wire antenna.

– **Unbalanced, Two-Wire Transmission** (RS232, RS423, etc.)

This time, the CM current is the % of the signal current which “elect” to return by the external path instead of using the return wire. Above a few kHz, the mutual inductance between the two wires is strong enough to attract most of the current into the intended path i_2 (Fig. 2.9), leaving only 20–30% of the current returned by the ground loop (i_3). Unless one knows the exact value of the mutual inductance, which itself varies with the cable height and wire separation, a conservative value is to expect $i_3 = i_1 - 10$ dB. This is of course assuming an unshielded cable (see Chap. 11 for the additional suppression by a shield).

– **Unbalanced Coaxial Transmission** (RF, Video, Ethernet, etc.)

This type of link is inherently low radiating, due to the high % of current—typically more than 99%—returning by the shield. This, too, is covered in Chap. 11.

– **Differential Transmission** (RS422/485, MIL-STD-1553, Diff. SCSI, CAN, USB, LVDS, IEEE 1394, Ethernet, HDMI, etc.)

With such links, the transmitter and receiver circuits have been designed to force balanced currents into the wire pair. In addition to excellent immunity and low crosstalk, this maintains also a low CM current generation, hence less EMI radiation. Along with the symmetry of the driver and receiver, a good symmetry of the wire pair is also required, such as the overall balance of the transmission is within the 1-10% range. Unless it is exactly known, we can conservatively assume 5%, i.e., the CM current i_3 will be at least 26 dB below i_1 .

2.5.1.2 Non-intentional Signals

As shown on Fig. 2.10, these are residue of clock frequencies, switch-mode power regulators, etc., coupled onto PCB I/O traces from:

- Ground traces/plane noise
- V_{cc} pollution (this in turn is partially transferred to I/O lines by some “transparency” in the driver or receiver ICs)
- Crosstalk

They can also be coupled by near-field radiation from the PCB hot spots onto the exiting cables, because of box shield leakages around the connector area. These leaks are causing the two (or more) wires to be CM-driven like a single conductor. The first difficulty is to evaluate the amplitude of these undesired signals. A pragmatic approach is to measure their voltage or current spectrum directly on the cable itself. This is easy to do at a diagnose and fix level, but it requires at least a representative prototype at the design stage. How can one do this when the hardware does not yet exist? A deterministic approach would consist in calculating every possible internal coupling between the inner circuitry and the leads corresponding to I/O ports. This is feasible but takes a considerable amount of time. A crude but effective solution is to make the following assumption by default: *Unless one knows better, it is logical to assume that the noise picked-up by internal couplings is just below the immunity level of the circuits interfacing the external link in question.*

The rationale for the above is that designers will at least make their product functionally sound; if worst-case values were exceeded (e.g., 0.4 V for a digital input noise margin), the system simply would not operate properly.

Example 2.3 A serial link between a computer and its peripheral operates at a 20 kb/s rate. The internal circuitry uses a 50 MHz clock with associated Schottky logic circuits, with the following characteristics:

- Amplitude = 3 V
- Rise time (STTL) = 3 ns
- Noise margin (worst case) = 0.3 V

Not knowing the exact layout of the inner circuit, estimate the worst possible noise picked-up by the 20 kb serial link.

Lacking of any other data, we can make the following worst-case assumptions (Fig. 2.11):

- The amplitude of any parasitic coupling from a clock-triggered pulse to a nearby trace or wire will not exceed the 0.3 V noise margin.
- Because crosstalk and ground-impedance sharing are all derivative mechanisms, the pulse width of the coupled spike on the victim trace will be in the range of 3 ns. STTL transition time. This is a worst-case guess since pulse stretching and ringing will occur due to distributed parameters of the victim line.
- If crosstalk exists, it will appear as a differential (signal-to-Gnd) noise at the I/O port.

Therefore a DM-to-CM reduction factor must be applied to the driven cable, like -10 dB for an ordinary unbalanced link (see previous Sect. 2.5.1.1, Unbalanced 2-Wire Transmission)

In Summary

- The crosstalk pollution of the low-speed *I/O* link by the internal 50 MHz spurious will appear as 0.3 V, 3 ns-wide spikes, with alternating positive and negative sign, riding over the 20 kb signal pulses train. Approximately 2/3 of the corresponding current will flow differentially in the pair, causing only minor radiation. The remaining 1/3 will return via the large cable-to-earth loop, causing a CM loop radiation with a spectrum populated by 50 MHz harmonics.
- The PCB ground pollution by the same 50 MHz spurious will appear as unipolar pulses (Fig. 2.11b) because they correspond to the transient overcurrent demand (“thru-current”) of digital gates at each 0-to-1 or 1-to-0 transition. The resulting PCB ground noise can be shared by several circuits in this board area, forcing CM currents into the whole *I/O* pair, returning by the local chassis or facility ground. This coupling is likely to be the greatest threat because it excites entirely the cable-to-ground loop, without DM-to-CM conversion loss.

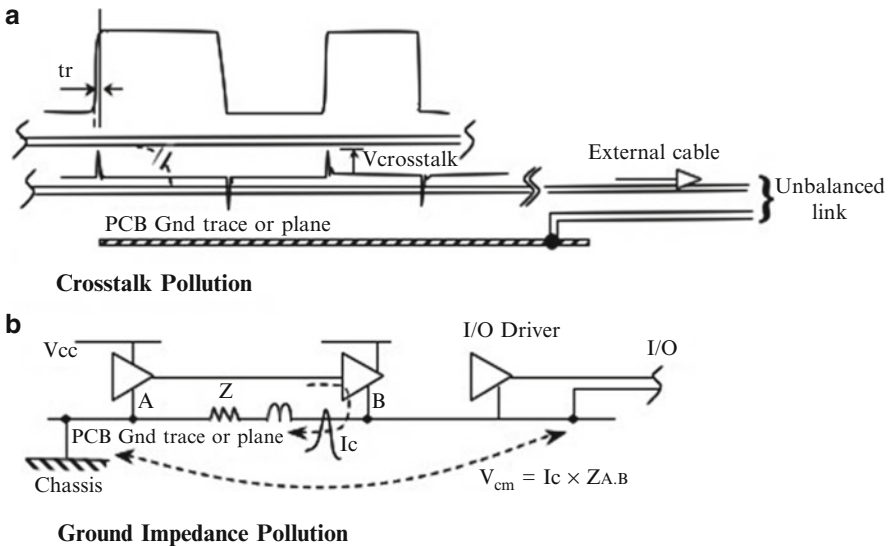


Fig. 2.11 Coupling of internal clock transitions by crosstalk to *I/O* traces

2.5.2 Approximating the Proper Radiating Geometry

Having determined the CM current driving mechanism, we now have to figure out what is the driven antenna: closed loop or open wire? Although they are gathering many configurations in two rather crude configurations, resorting to these two simple models gives remarkably good results when compared to actual validation test data.

2.5.2.1 Loop with Defined Contour

By this term, we do not imply necessarily that the loop is physically closed; the signal grounds (0 V Ref.) may or may not be connected to chassis or earth reference at both ends. If the I/O cables are connecting to metallic equipment cases (and hence are most likely grounded), the CM radiating loop is geometrically identified by its size $\ell \times h$ (Fig. 2.9). If the ground references are floated inside the box, this will increase the low-frequency impedance of the loop, yet its size remains. Therefore, depending on the loop impedance and the near-field or far-field conditions, we will apply either the loop equations or the two-wire equations of Sect. 2.3.

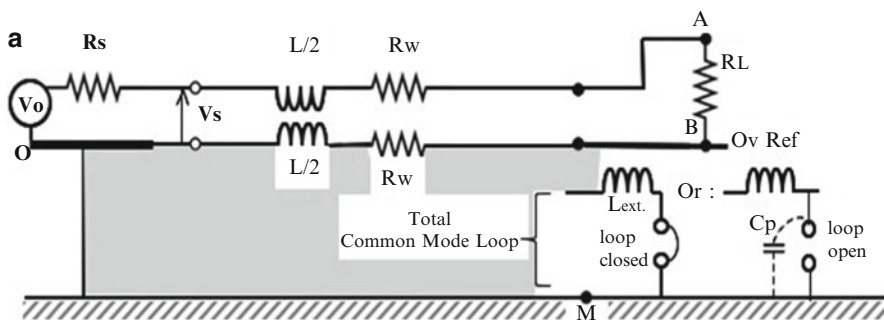


Fig. 2.12 (a) Loop Impedance of External Cables Radiation

For all-grounded ends (PCB to chassis and chassis to local earth network), the current i_1 delivered by the signal source to the line termination R_L is splitting in two possible return paths (Fig. 2.12b):

- Current i_3 returning to the source reference by the large cable-to-ground CM loop. The loop impedance seen by this i_3 current is

$$Z_{CM} = R_w + R_L + j\omega L_{CM} \quad (2.23)$$

where L_{CM} is the wire self-inductance above ground plane. R_w (typ. $0.1 \Omega/m$) can be neglected, and typical value for L_{CM} , with heights in the 0.1 to 1 m range, is $1\text{--}1.5 \mu H/m$, and so Equ. (2.23) is rewritten in a more convenient way:

$$Z_{CM} = R_L + j7.5 \Omega \times \ell_m \times F_{MHz} \quad (2.23a)$$

The amplitude of i_3 is approximately the value that the return current would take if the second wire of the pair did not exist. This current is the major contributor to cable radiation, via the CM loop.

- Current i_2 is returning back to its source via the return wire of the pair. The total loop impedance seen by this current i_2 is

$$Z_{diff} = 2R_w + R_L + j\omega L_{diff} \quad (2.23b)$$

where L_{diff} is the differential loop self-inductance of the pair. This inductance is substantially lower than that of the large CM loop, thanks to the mutual inductance between the closely spaced wires. For typical, small-gauge signal wire, this 2-wire inductance is in the order of $0.5\text{--}0.6 \mu H/m$.

The total output current i_1 is the combination of $i_2 + i_3$.

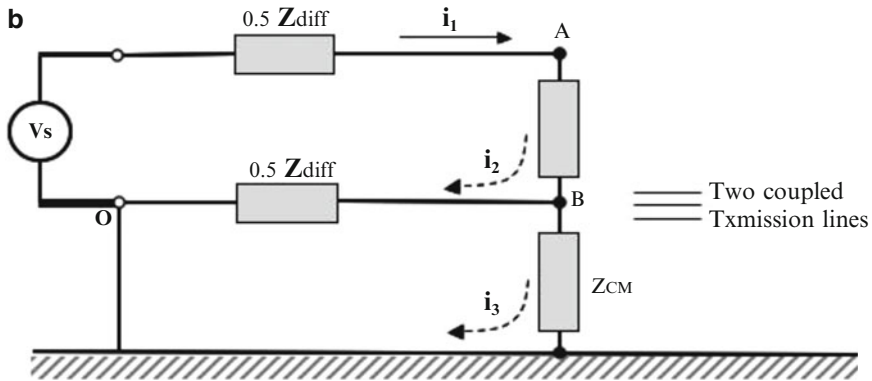


Fig. 2.12 (b) Split of return currents in the Loop Impedances of External Cables Radiation

For a floated end, the loop impedance seen by the CM current is

$$\begin{aligned} Z_{CM} &= R_w + R_L + (j\omega L_{CM} - j/C_p\omega) \\ &\approx 1/C_p\omega \text{ for low frequencies.} \end{aligned} \quad (2.24)$$

where C_p is the PCB-to-chassis stray capacitance (≈ 30 pF for small boxes, $100\text{--}200$ pF for large cabinets). When cable length exceeds $\lambda/2$ (for both ends

grounded) or $\lambda/4$ (for one end floated), Z_{CM} can be approximated by the characteristic impedance of the cable above ground:

$$\text{in the air, } Z_0 = 60 \ln(4h/d) \quad (2.25)$$

where,

\ln = natural logarithm

d = cable diameter (average contour of whole wire bundle)

Practical h/d range of 3-100 gives a span of Z_0 from 150 to 360 Ω . The low value would correspond to a typical MIL-STD-461 test setup, the high value being the extreme for a tabletop equipment in an FCC or CISPR test. A typical real-life value would be 250 Ω .

Example 2.4 A 5 MHz clock is used on a short-haul parallel bus. For a 5 V pulse, the ninth harmonic, at 45 MHz, has an amplitude of 0.3 V. The characteristics of the I/O cable between the two metallic equipments are:

1. Cable length, $\ell = 1.20$ m
2. Height, $h = 0.30$ m
3. Inductance of cable above ground $L = 1.2$ $\mu\text{H/m}$
4. Terminating resistor = 120 Ω

Calculate the 45 MHz E field at 3 m vs. the FCC Class B limit for (a) both PCBs grounded to chassis and (b) one PCB 0 V ground floated, with a total stray capacitance of 30 pF.

Solution

- $D = 3$ m $> 48/45$ MHz, so we are in far-field conditions.
- $\ell(\text{m})$ and $h(\text{m})$ are $< 75/45$ MHz, so we are below cable “antenna” resonance.

We can use directly Equ. (2.20) or Fig. 2.6 curves for 3 m distance:

$$\text{Area} = 120 \text{ cm} \times 30 \text{ cm} = 3,600 \text{ cm}^2 = 72 \text{ dB cm}^2$$

- (a) *For grounded condition*, the cable is essentially an inductance; impedance is calculated from Equ. (2.23a):

$$Z_{cm} = 120 \Omega + j(7.5 \times 45 \text{ MHz} \times 1.2 \text{ m}) = 420 \Omega$$

- (b) *For floated condition*, the cable inductance resonates with the floating PCB capacitance at

$$F_{\text{res}} = 1/2\pi\sqrt{LC} = 1/2\pi\sqrt{(1.2 \text{ m} \times 1.2 \cdot 10^{-6} \text{ H} \times 30 \cdot 10^{-12})} = 24 \text{ MHz}$$

Therefore, due to resonance downshifting caused by the stray capacitance, the cable is now beyond resonant condition. We will use a typical characteristic impedance of $250\ \Omega$, dictating the average value of the current along the cable.

Calculation spread sheet for 45 MHz frequency

1) E_0 in Fig. 2.6 (for 1 V - 1 cm ²)		
– For Z: 420 Ω :	8 dBμV/m	
– For Z: 250 Ω (floated):		11 dBμV/m
2) Area correction	+72	+72
3) Amplitude corr. (0.3 V)	-10	-10
4) $E = 1 + 2 + 3$	70 dB μ V/m	(73 for floated)
FCC limit, Class B	40 dB μ V/m	
Off specification	30 dB	(32 for floated)

For the all-grounded case, a quick analysis of the circuit in Fig. 2.12b, with the variables of the example, gives the following current split:

- i_1 (upper wire) = 2 mA
- $i_2 = 1.3$ mA
- $i_3 = 0.7$ mA, which is 10 dB below i_1

This is due to the DM loop inductance of the wire pair being approximately a third of the large CM loop inductance. Notice that we have passed the point where a floated PCB could be of any use. The exact calculation of I_{average} could be made using transmission line theory and would give slightly different results for each resonant condition. To reduce this excessive emission will require one of the several solutions (e.g., CM ferrites, cable shield, balanced link) that we will examine later.

2.5.2.2 Open Wire, Monopole or Dipole

We now examine the case where no geometric loop can be identified (Fig. 2.13). The external cable terminates on a small, isolated device (sensor, keypad, etc.) or into a plastic, ungrounded equipment. It may even not be terminated, waiting for a possible extension to be installed. No finite distance can be measured to a ground plane. In this case, we use the single-wire radiation model described in Sect. 2.2.

In a sense, we can say that the floating, open wire is the maximum radiating antenna that can be achieved when the height of a ground loop increases to infinity. To calculate the radiated field using Equ. (2.13) or Equ. (2.16) requires that the CM current in the wire be measured or calculated. Measurement with a high-frequency current probe is easy, but only if a prototype is available. Otherwise, one can simply use the cable self-capacitance of ≈ 10 pF/m for low-frequency modeling and the cable characteristic impedance Equ. (2.25) with high values of h , above the first resonance.

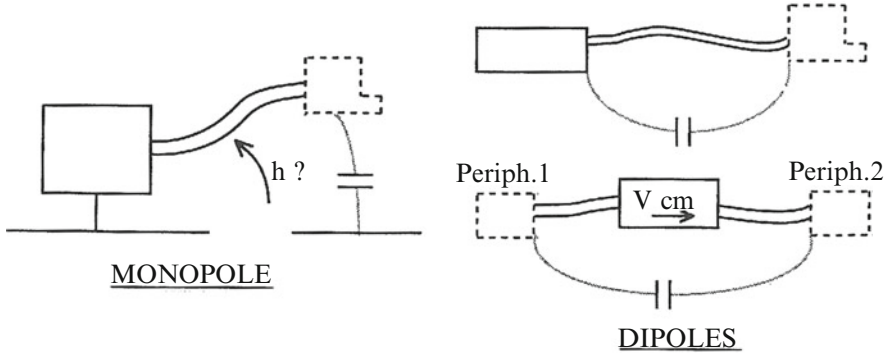


Fig. 2.13 Common-driven monopole or dipole

Using more practical units, single-wire radiation is expressed as

$$\text{In the near field, } E_{(\mu\text{V/m})} = \frac{1,430 \times I_{\mu\text{A}} \times \ell_m}{D^3 F_{\text{MHz}}} \quad (2.26)$$

$$\text{In the far field, } E_{(\mu\text{V/m})} = \frac{0.63 I_{\mu\text{A}} \times \ell_m \times F_{\text{MHz}}}{D_m} \quad (2.27)$$

If the cable interconnects two units that are completely floated and not close to any ground, the length, ℓ , is regarded as a radiating dipole length. If one of the two units is grounded or is in a metallic case close to ground, the cable has to be regarded as a radiating monopole whose length, ℓ , radiates like a dipole twice as long. In this case, 2ℓ should be entered in the formula.

Of course, wire lengths may exceed $\lambda/2$ ($\lambda/4$ for a monopole). In this case, the current can no longer be regarded as uniform over the wire length. But the “active” segment of the radiator cannot exceed a length of $\lambda/2$. The other $\lambda/2$ segments create fields that mutually cancel due to phase reversal (except for the field propagation delays, which are unequal). Everything behaves as if the antenna were electrically “shrinking” as F increases. In this case, ℓ must be replaced by $\lambda/2$ in the formula.

Applying a correction factor averaging I_{max} over the length, we have, for free space,

$$E_{(\mu\text{V/m})} = \frac{60 I_{\mu\text{A}}}{D} \quad (2.28)$$

Interestingly, we observe that E becomes independent of F and ℓ . This formula is extremely useful, and we will employ it frequently.

Example 2.5 For the same 5 Mb/s signal as in Example 2.4, assume that the 1.20 m cable now terminates into a plastic equipment on one end. The cable is far from

any ground plane. What is the maximum CM current tolerable on this cable to meet the FCC (B) 3 m limit of 100 $\mu\text{V}/\text{m}$ at 45 MHz (harm #9) and 85 MHz (harm #17)?

- At 45 MHz, $\ell < 75/F$. From Equ. (2.27), remembering that we have a monopole, a length 2ℓ is entered as radiating length:

$$E = 0.63I \times (1.20 \times 2) \times 45 \text{ MHz} / 3 \text{ m}$$

Solving for the current I ,

$$I < E/22.5 \text{ so, } I < 4.4 \mu\text{A}$$

- At 85 MHz, $\ell > \lambda/4$. We will use Equ. (2.28):

$$E = 60I/D$$

$$I \leq ED/60, \text{ so } I < 5 \mu\text{A}$$

Therefore, before running an exhaustive radiated EMI test, a simple measurement on the cable with a high-frequency current probe will indicate whether the equipment has a good chance of meeting the specification.

2.5.2.3 Simple Voltage-Driven Nomograms for Open Wire

If the CM current cannot be measured on a prototype, instead of computing the open-wire impedance at each frequency, the nomogram of Fig. 2.14 for isolated wires can be used for a quicker estimate. All that is needed is the value of the voltage(s) driving the monopole or dipole. The curves are based on cable capacitance, for electrically short lines, and a conservative 150 Ω CM impedance otherwise. Notice that, at each exact resonance, a 6 dB hump has been accounted for the 75 Ω impedance of a tuned dipole. Results are given in dB $\mu\text{V}/\text{m}$ for one μV (0 dB μV) of CM excitation.

Notice that for the 1 m test distance, the length of the longest effective dipole has been limited to 1 m. This is to prevent an overprediction of the field amplitude, since the farthest segments of the radiating wire would be at a distance greater than 1 m from the receiving antenna.

Example 2.6 Taking the same equipment as Example 2.5, assume that the 5 MHz CM current is not known, but the harmonic #9 CM driving voltage is 100 mV. This voltage may have been measured vs. chassis by a voltage probe directly at the I/O connector. We will extrapolate from the curve for a 2 m dipole (equivalent to 1 m monopole) at 45 MHz.

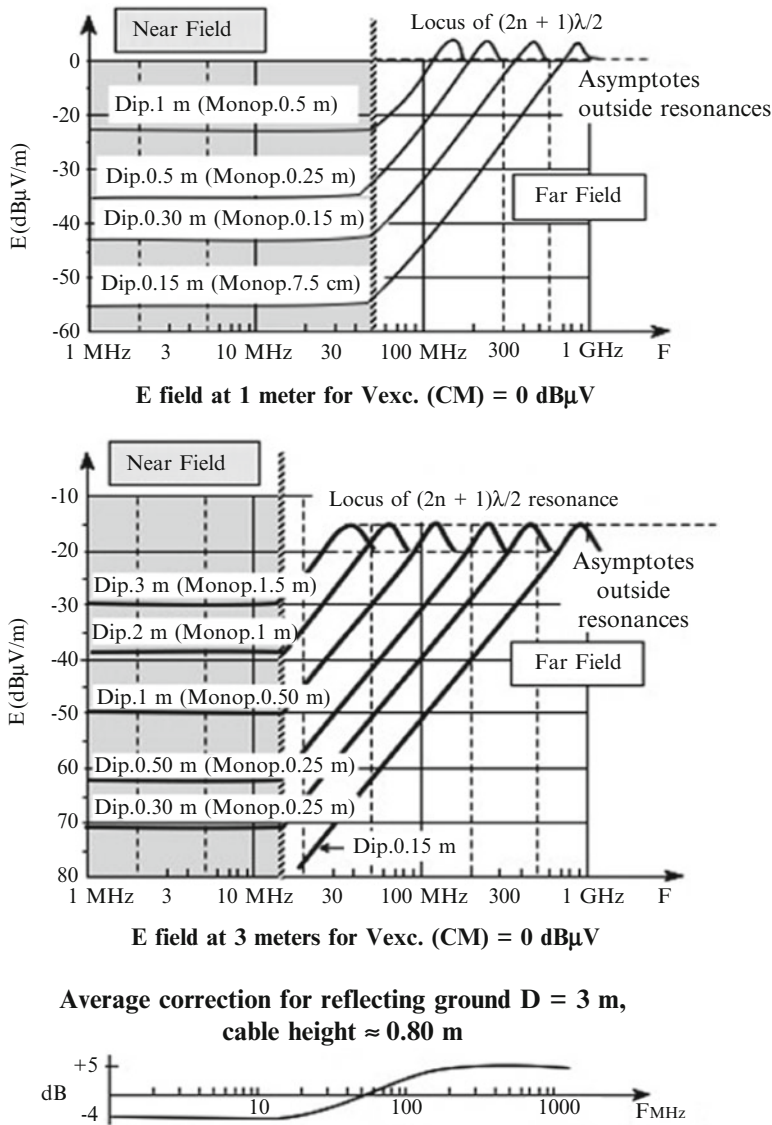


Fig. 2.14 E-field radiation at 1 and 3 m distance from voltage-driven, open wires. Field given for 0 dBμV of CM drive. Peak fields horizontal locus correspond to odd multiples of $\lambda/2$

Solution

Test distance 3 m, frequency:	45 MHz
1. E_0 for 2 m dipole and 0 dBμV (Fig. 2.14)	-23 dBμV/m
2. Length correction* $20 \log (1.20/1)^2$	+3
3. Amplitude correction ($10^5 \mu V$)	+100
$E = 1) + 2) + 3)$:	80 dBμV/m (40 dB above Class B)
*For a voltage-driven wire and below resonance, radiation efficiency increases like the square of length (relates to equivalent radiating area).	

We see that with approximately the same CM current, the field is approximately twice as large than with a closed loop, 0.30 m above ground. This can be considered as the upper bound for a CM-driven cable infinitely far from ground. Notice how little voltage it takes to excite an open wire above the radiated limit.

2.5.2.4 Influence of a Nearby Ground Plane

If there is a conductive plane near the cable, this proximity causes a reflected wave with a phase shift (see Fig. 2.15). If the plane is sufficiently close, this shift is always at phase reversal with the directly radiated wave, and the total field equals $E_0 - E_r$. It is not necessary that the source or load be referenced to this plane, but the plane must be quasi-infinite. In practice, it must extend far enough around the cable projection and farther than the cable-to-antenna distance.

The radiation reduction, for $h < 0.1\lambda$ (i.e., $h(m) < 30/F_{\text{MHz}}$), is

$$\frac{E_{\text{total}}}{E_0} = \frac{h}{0.1\lambda} \text{ or } \frac{10h}{\lambda} \quad (2.29)$$

Entering this factor into Equ. (2.27), for far field:

$$E_{\mu(\text{V/m})} = \frac{0.021 \times I_{\mu\text{A}} \times \ell_m \times h_m \times F_{\text{MHz}}^2}{D_m} \quad (2.30)$$

If $h(m)$ is $> 30/F(\text{MHz})$, the reflected field is alternatively additive or subtractive, and the field is not reduced but doubled at certain frequencies. Incidentally, Equ. (2.30) shows a similarity with loop radiation from a since E now also depends on the area $\ell \times h$, and F^2 .

Example 2.7 For the equipment of Example 2.5, recalculate the maximum acceptable value for CM current with the cable now located 5 cm from a ground plane. The criterion to meet is MIL-STD-461-RE102, at 1 m distance. At 45 MHz, the limit is 24 dB $\mu\text{V/m}$. At 150 MHz, the limit is 30 dB $\mu\text{V/m}$. We can accept the far-field assumption for both frequencies.

Solution

- At 45 MHz, $\lambda = 6.6$ m, so the 1.20 m cable length is $< \lambda/4$. Since h is 0.05 m, it is $< 30/F$. Per (2.30) and remembering that we have a monopole,

$$E = 0.021 \times I(1.20\text{ m} \times 2) \times 0.05 \times 45^2 = 5 \times I = I \text{ dB}\mu\text{A} + 14 \text{ dB}$$

Therefore, for limit compliance, we must satisfy $I = E_{\text{limit}} - 14 \text{ dB}$

$$I \leq 24 \text{ dB}\mu\text{V/m} - 14 \quad I \leq 11 \text{ dB}\mu\text{A}$$

- At 150 MHz, $\ell > \lambda/4$, and h is still $< 30/F$. Equation (2.28) will be corrected by ground reflection factor $h/0.1\lambda$ that we replace by $h/(0.1 \times 300/F)$ or $h F/30$

$$E = (60I/D) \times 0.05 \times 150/30$$

$$= 15I \text{ or } I + 23.5 \text{ dB}$$

Therefore, the E -field limit translated into a CM current limit is $I \leq E_{\text{limit}} - 23.5 \text{ dB}$

$$I \leq 30 \text{ dB}\mu\text{V/m} - 23.5 \text{ dB} \quad I \leq 6.5 \text{ dB}\mu\text{A}$$

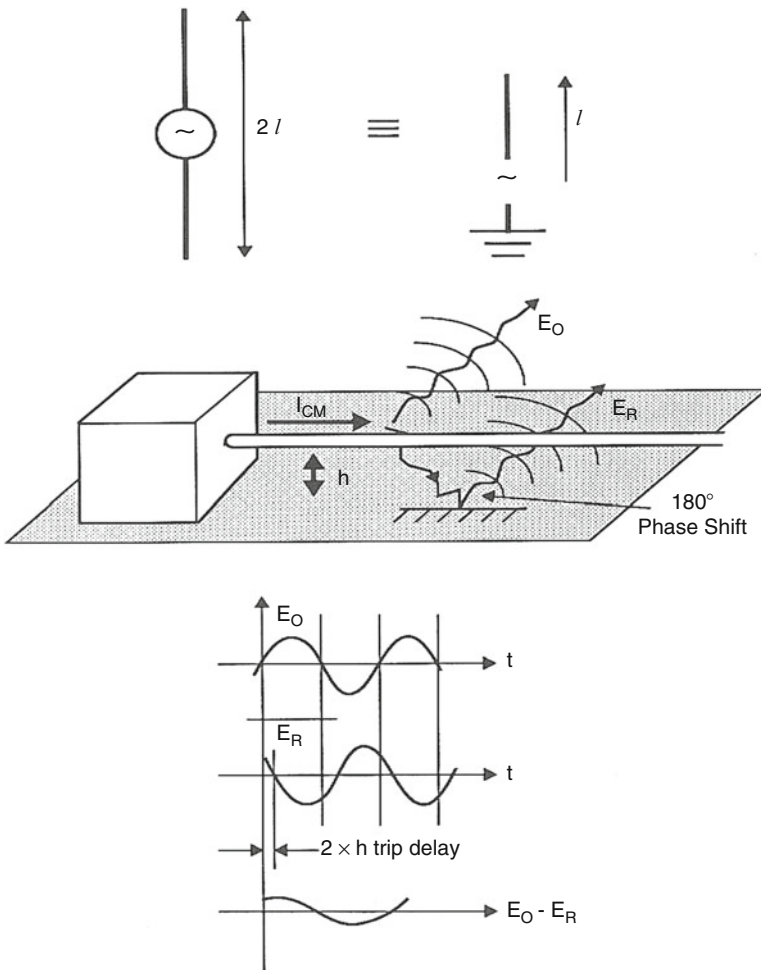


Fig. 2.15 Equivalent antenna for wire floated at both ends (dipole) or grounded at one end (monopole). Effect of a nearby ground plane for $h < \lambda/10$

2.5.2.5 Ground Reflection with FCC/CISPR Tests

The reflected field addition on an *infinite ground plane*, practically doubling the value of E , is typical of FCC 15B or CISPR 22 emission testing. The equipment under test (EUT) and its cables are located above a ground plane, while the receiving antenna is cranked up and down to find the maximum level. A typical test configuration is the following:

- EUT + cables: 0.5–1 m above ground
- Test distance: 3 m
- Antenna set for vertical and horizontal polarization, with height scan 1–4 m

The corresponding correction for $(E_{\text{dirr.}}, E_{\text{refn}})$ max. vs. free-space field is given below [3]

EUT height above ground	Horiz. polar.		Vert. polar.	
	0.5 m	1 m	0.5 m	1 m
F_{MHz}	Corr. Δ dB		Correction Δ dB	
30	+5	+4	-8	-4
50	+5	+4	-5	-1
70	+5	+3	-2	+2
100	+4	+2	0	+4
150	+3.5	0	+3	+5
200	+3		+4	+5

Below 100 MHz, the strongest reflection is detected with horizontally polarized sources (seldom the case with CM-excited cables), typical of DM sources inside the EUT. Above 100 MHz, the additive reflections are caused by vertically polarized sources, adding a maximum of 5 dB to the theoretical free-space value. We therefore can retain, as a reasonable worst case for all FCC- or CISPR-like measurements, a +5 dB ground reflection addition in our emission calculations.

As a concluding remark, Fig. 2.16 is showing the respective field contributions of PCB vs. external cables in two situations. The first plot, at top, corresponds to a PCB with 30 MHz clock circuit, attached to a 1.5 m external cable carrying typical spurious contents. As explained in this section, the cable CM radiation dominates the PCB DM contribution by 25–30 dB and is the sole violator of FCC/CISPR limit. In the bottom plot, the same geometry with a 150 MHz PCB shows a different split. Above 100 MHz, the spurious spectrum carried by the cable starts decreasing because:

- The maximum possible crosstalk coefficient in the PCB has been reached.
- Cable starts exhibiting HF losses.
- Around these frequencies, the cable radiating efficiency for 1.2–1.5 m lengths has reached its maximum.

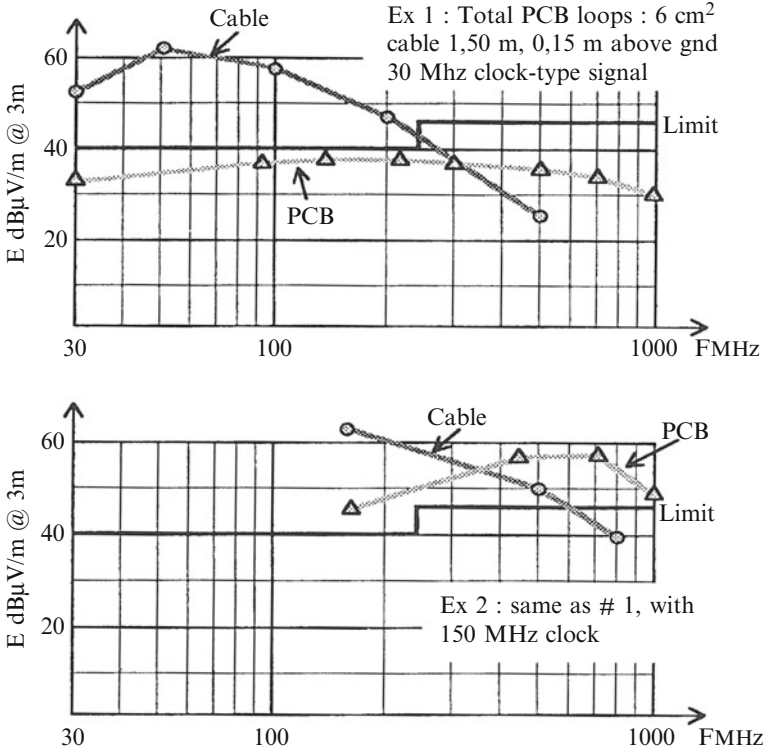


Fig. 2.16 Comparison of PCB vs. cable contribution to radiated spectrum

2.5.3 Radiation from a Long Wire

The distance restriction imposed for using the loop model or the wire models (i.e., $\ell < D$) rapidly becomes an obstacle to calculations in many configurations where cable lengths exceed a few meters. In this case, the physical length of the wire is such that it cannot be considered as a small element with respect to the observation distance. We can use a practical expression, taking into account the wide viewing angle from the observation point to the cable (infinite wire model):

$$H(A/m) = I(A)/2\pi D \quad (2.31)$$

Only the H field can be correctly determined by this Ampere's law. An "equivalent" E field could be derived using a $120 \pi \Omega$ wave impedance, but it would be inaccurate in this very near-field zone from such long antenna. Equation (2.31) assumes the wire is far from a ground plane, with respect to observation distance (i.e., in practice, $h \gg D$). If the wire is close to a ground plane and the observation point is at the same height, $h \ll D$:

$$H \approx \frac{I}{2\pi D} \left(1 - \cos \frac{2h}{D} \right) \quad (2.32)$$

This formula is valid for low frequency, a condition being that the phase shift due to the offset h be totally negligible. Notice that for small values of $2h/d$, cosine is approaching 1 and the H field tends to zero.

The criterion for deciding when a wire has to be considered to be “long” in comparison to D is simple: the maximum field is reached when $\ell \geq \pi D$. If one takes $\ell > D$ as criteria, the error would be only 16% (1.3 dB). This is because a length increase from $\ell = D$ to $\ell = \pi D$ corresponds only to an argument viewing angle variation from $\cos \alpha = 0.84$ to $\cos \alpha = 1$.

Case of the Long-Wire Pair (DM Radiation)

The infinite wire equation, when transposed to a long-wire pair carrying equal and opposite currents, becomes

$$\begin{aligned} H(\text{A/m}) &= 0.16I [s/(D^2) - s^2/4] \\ &\approx 0.16I (s/D^2) \text{ for } D \gg s \end{aligned} \quad (2.33)$$

with:

D = pair to receiver distance, measured from pair axis

s = wires separation

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