

# Agilent Low-Level RF Leakage Measurement

Application Note 1204





# Introduction

There are many advancements in two-way radios and pagers including greater sensitivities, and higher frequencies. A radio or pager which can receive very low level signals far from the transmitter is considered more valuable and can command a higher price. Regulatory agencies are opening channels to higher frequencies. In order to design and produce these new products, component, device and product manufacturers are requiring signal sources which can accurately simulate the low-level signals in the right frequency range which their new products are being designed to work with.

The impact of greater receiver sensitivity is subtle. The obvious need is for sufficient output level attenuation to obtain the new lower test signal levels. The less obvious need is for reduced RF leakage from the signal generator. The test signal should leave the signal generator by the RF output connector only. But the signal may also leave the signal generator by a radiated leakage path. If the radio under test is not well shielded, or if the radio's shielding covers are off for purposes of test, the leakage signal can enter the radio along with desired signal from the RF output connector. The two signals can add either constructively or destructively, resulting in a total signal of unknown amplitude. This reduces the accuracy of the test.

Higher frequencies also demand more of a signal generator. Once again, there is an obvious need—test signals at higher frequencies. But higher frequencies also make leakage from the signal generators worse. For example, a cabinet seam becomes a larger fraction of a wavelength as the frequency increases, and its shielding effectiveness goes down. Thus, increasing leakage with increasing frequency.

The task for the manufacturer of the signal generator is to measure leakage at higher frequencies and lower levels, while improving the accuracy of those measurements. This paper addresses the measurement problems, and briefly discusses some of the work done to reduce leakage from signal generators.

Many are familiar with the problems of radiated RF leakage from computer equipment and its effects on radio and television reception. Reception can be degraded even at some distance away, say 10 meters or more. The radio regulatory agencies in various countries, for example the Federal Communications Commission in the United States, have set limits on the electric field emissions from electronic equipment. Agilent has generated a field strength specification that assures compliance in practically any country by combining the most stringent requirements of all these countries into one field strength curve.

Agilent signal generators have been designed to meet these same regulatory limits. How do these regulatory field strength limits compare to radio sensitivities? Let's look at the portion of the regulatory field

strength curves from 100 MHz to 1000 MHz, and compare it to typical VHF and UHF radio sensitivities (.25 mV) in this same frequency range. Assume the radios use a resonant dipole antenna.

Figure 1 shows that even at 10 meters away, emissions from equipment passing the regulatory requirements for RF leakage can be detected by radios quite easily. Typically, radio tests are done at distances much less than this, often times within one meter of the signal generator.

What is this field, and how do we compare field strength to radio sensitivity? How do we quantify this?

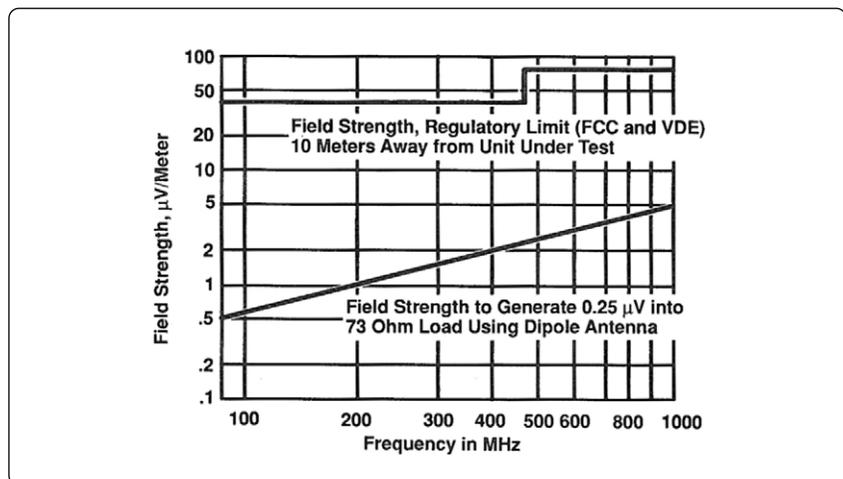


Figure 1. Regulatory field strength limits vs. typical radio sensitivities

The electromagnetic fields we are discussing are made up of two components—an electric field (E), and a magnetic field (H). The strength of the E field is measured in volts per meter, and the H field is measured in units of amps per meter. The origin of these fields may be unintentional radiation, such as RF leakage from a signal generator or computer, or may be intentional radiation, such as from a radio transmitter. At a distance of several wavelengths from an electromagnetic field source, perhaps a transmitting antenna, the magnitudes of the E and H fields are found to be in a fixed ratio:

$$(1) \frac{|E|}{|H|} = 377 \text{ ohms,}$$

and 377 ohms is defined as the impedance of free space. Multiplying the magnitude of the E field with the magnitude of the H field yields the power flux density in watts/sq. meter:

$$(2) P_f = |E| |H| \frac{\text{watts}}{\text{meter}_2}$$

$$(3) = \frac{|E|^2}{377} \frac{\text{watts}}{\text{meter}_2}$$

$$(4) = 377 |H|^2 \frac{\text{watts}}{\text{meter}_2}$$

These are the fundamental units of electromagnetic fields. They allow us to relate RF leakage, regulatory requirements, and radio sensitivities to each other as we shall see.

Radio sensitivity is often specified as microvolts into a given input impedance) for example .5μV into 60 ohms. It can also be specified in input power, for example 5 femtowatts. These sensitivity figures are usually given for 12 dB SINAD out of the radio. How do we convert from field strength to radio sensitivity? Antennas convert electromagnetic fields to power (or voltage) into the radio input terminals. Let us review some antenna basics. Let's start with a transmitting antenna. If we assume an isotropic antenna, power will be radiated uniformly in all directions. The power flux density ( $P_f$ ) at a distance r from the antenna is then;

$$(5) P_f = \frac{\text{power to antenna}}{\text{surface area of sphere}} \\ = \frac{P_t}{4\pi r^2} \frac{\text{watts}}{\text{meter}^2}$$

Most antennas are not isotropic, and will radiate more power in some directions than others. The field strength will be increased in those directions by the amount of the gain over isotropic. This number is the antenna gain. The power flux density in the preferred direction will then be;

$$(6) P_f = \frac{G_t P_t}{4\pi r^2} \frac{\text{watts}}{\text{meter}^2}$$

where  $G_t$  = transmit antenna gain.

A resonant half wave dipole has a gain of 1.64, or 2.15 dB. This gain factor holds for both transmit and receive antennas. A transmitting dipole will radiate fields in its preferred direction 1.64 time as strong as an isotropic antenna. Similarly, a receiving dipole will provide a signal at its terminals 1.64 times as strong as the signal from an isotropic antenna.

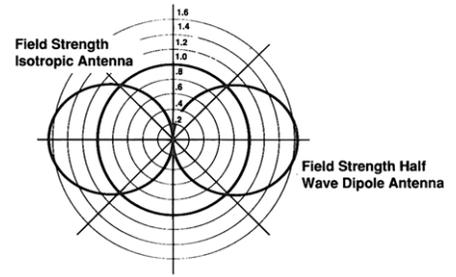


Figure 2. Dipole field strength, pattern

To calculate the relationship between field strength and antenna terminal voltage, we use the power flux density equation (6) and the equation for path loss between two antennas. Path loss is the reduction in power of the signal at the receive antenna terminals compared to the transmit antenna terminals. It is due to the spreading of the signal power over a larger area as the distance from the source increases. A given receive antenna will then intercept less power as this distance increases. The equation for path loss is;

$$(7) \frac{P_r}{P_t} = \frac{G_r P_t \lambda^2}{(4\pi r)^2} = \frac{A_{er} A_{et}}{(\lambda r)^2}$$

where:

- $A_{er}$  = Effective area, receive antenna
- $\lambda$  = Wavelength in meters.
- $A_{et}$  = Effective area, transmit antenna
- $G_r$  = Receive antenna gain
- $G_t$  = Transmit antenna gain
- $P_f$  = Power flux density
- $P_r$  = Received power
- $P_t$  = Power transmitted by transmit antenna

Rearranging equation (7) gives:

$$(8) P_r = \frac{P_t G_r G_t \lambda^2}{16(\pi r)^2} = \text{watts}$$

Note that antenna gain and effective area can be related by:

$$(9) G = \frac{4\pi A_e}{\lambda^2}$$

To calculate the amount of power received by the antenna in a field with a given power flux density, we take the ratio of the received power to the power flux density,  $P_r/P_f$  [ratio of equation (8) to equation (6)].

Thus:

$$(10) \frac{P_r}{P_f} = \frac{\frac{P_t G_r G_t \lambda^2}{16(\pi r)^2}}{\frac{P_t G_t}{4\pi r^2}} = \frac{G_r \lambda^2}{4\pi}$$

Or equivalently:

$$(11) P_r = P_f \frac{G_r \lambda^2}{4\pi} = \frac{|E|^2}{377} \frac{G_r \lambda^2}{4\pi^2} \text{ watts}$$

We see that the received power does not depend on the nature of the transmit antenna, but only the field strength. For a resonant half wave dipole,  $Z_{in} = 73$  ohms (pure resistive).

Thus:

$$(12) E_r = \sqrt{73 P_r}$$

$$= \sqrt{73 \frac{E^2}{377} \frac{G_r \lambda^2}{4\pi}} \text{ volts}$$

$$(13) = E \lambda \sqrt{\frac{73 G_r}{377 \cdot 4\pi}} \text{ volts}$$

For a half wave resonant dipole,  $G_r = 1.64$ .

Thus:

$$(14) E_r = 0.159 |E| \lambda$$

assuming a 73 ohm impedance load. We now have a simple conversion between field strength and voltage into a 73 ohms load assuming a dipole antenna.

Well then, are we all ready to go now and make measurements? No! We have to decide at what distance we are going to measure, and what level is acceptable. As discussed earlier, receiver tests are often done close to the signal generator, usually less than one meter away. As equation (5) shows,

$$(5) P_f = \frac{\text{power to antenna}}{\text{surface area of sphere}}$$

$$= \frac{P_t}{4\pi r^2} \frac{\text{watts}}{\text{meter}^2}$$

Higher power flux density is found closer to the source. Thus, maximum leakage signal is obtained close to the signal generator. Let us assume that receiver tests are to be conducted as close as 25 mm away from the signal generator. Can we repeat the regulatory requirements tests with a new lower specification line and extrapolate to our 25 mm receiver test position? We might be able to do an extrapolation using equation (3),

$$(3) = \frac{|E|^2}{377} \frac{\text{watts}}{\text{meter}_2^2}$$

and equation (6),

$$(6) P_r = \frac{G_t P_t}{4\pi r^2} \frac{\text{watts}}{\text{meter}^2}$$

Combining the two, we find:

$$(15) |E|^2 = \frac{377 G_t P_t}{4\pi r^2} \text{ volts}^2$$

or:

$$(16) |E| = \frac{1}{r} \sqrt{\frac{377 G_t P_t}{4\pi}}$$

This analysis indicates that a magnitude of the electric field strength  $|E|$  varies as  $1/r$ . We know from equation (14) that electric field strength and antenna terminal voltage (and thus receiver terminal voltage) vary directly with each other. If we desire no more than  $0.25 \mu\text{V}$  of antenna terminal voltage at 25 mm away, by our analysis, what must we measure at 10 meters away to insure this? The  $1/r$  electric field strength relationship indicates that we would find  $0.625 \text{ nV}$  of signal at the antenna terminals. A very low noise receiver with a bandwidth of less than 1 Hz would be needed for this measurement. To make things easier, why not just make the measurement at 25 mm away?

We must look again at our analysis. The assumptions made for the analysis thus far hold true only at distances several wavelengths from the source of radiation, that is, the far field region. Several nice field properties hold in the far field region;

1. The ratio of  $|E|$  to  $|H|$  is 377 ohms, independent of source.
2. The magnitude of E and H vary as  $1/r$  along a given radius line where  $r$  is the radial distance from the source.
3. The fields are plane waves.

What happens closer in, at distances of a fraction of a wavelength? This region is called the near field region, and things get more complicated. First of all, to simplify the discussion, let us assume that the sources of radiation are point sources. Also we must now be concerned with two types of sources—an electric point source and a magnetic point source.

An infinitesimal dipole is an electric field point source. An infinitesimal loop is a magnetic field point source.

The equations for the fields generated by an electric point source are below:

$$(I_m \leftrightarrow I_o)$$

$$(-E_\theta) H_\phi = \frac{I_o h}{4\pi} e^{-jkr} \left( \frac{jk}{r} + \frac{1}{r^2} \right) \sin\theta$$

$$(H_r) E_r = \frac{I_o h}{4\pi} e^{-jkr} \left( \frac{j\eta}{r^2} + \frac{1}{j\omega\epsilon r^3} \right) \cos\theta$$

$$(H_\phi) E_\theta = \frac{I_o h}{4\pi} e^{-jkr} \left( \frac{j\omega\mu}{r} + \frac{1}{j\omega\epsilon r^3} \right) \sin\theta$$

These are the fields generated by the infinitesimal dipole current element  $I_o$ . The terms in parentheses are for the fields generated by an infinitesimal current loop. The term  $I_m$  is an imaginary magnetic "current" element, perpendicular to the current loop, whose magnitude is proportional to the magnitude of the current times the area of the loop.

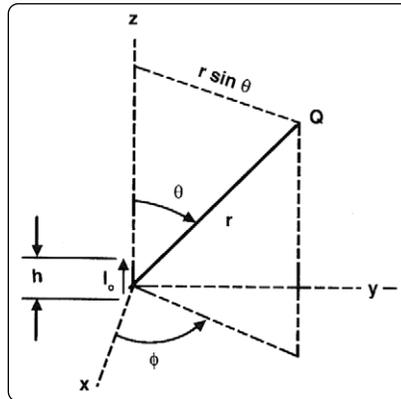


Figure 3. Electrical field source and current system

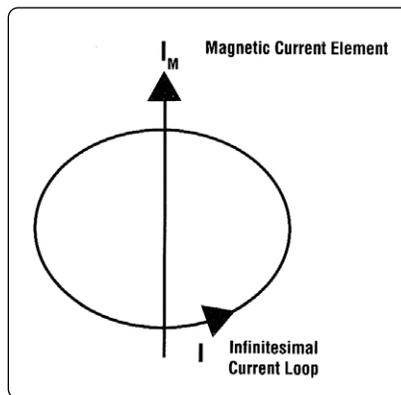


Figure 4. Magnetic current element

Looking at the equations, we now find that the field intensity varies as something other than just  $1/r$ —there are field components which vary as  $1/r^2$  and  $1/r^3$ . We find field components in all three orthogonal directions:  $r$ ,  $\phi$ , and  $\theta$ .

These equations hold in general, both near field and far field. In the far field, the  $1/r^2$  and  $1/r^3$  terms become negligible. The radial field component therefore also becomes negligible. Thus, in the far field, the fields vary only as  $1/r$ , and the ratio of  $|E|/|H|$  is 377 ohms.

The  $\phi$  and  $\theta$  terms are perpendicular to the direction of wave travel. The direction of wave travel is along radial lines. At large enough radius, the curvature in the  $\phi$  and  $\theta$  directions is negligible with respect to a wavelength, and so we have plane waves.

Dipoles are electric field antennas. Loops are magnetic field antennas. In the far field, where the E field and H field are in constant ratio to each other (377 ohms), the relative sensitivity of dipoles versus loops is always the same. In the near field, because of the additional  $1/r^2$  and  $1/r^3$  terms, this is no longer true. Let us look at the ratio of  $|E|$  and  $|H|$  in the near field so see what the difference in sensitivity between dipoles and loops would be.

The ratio of  $|E|$  and  $|H|$  is called the wave impedance. This illustration shows how the wave impedance varies as a function of distance and nature of the source, with the radial component neglected. Note that the distance is normalized to the wavelength divided by  $2\pi$ .

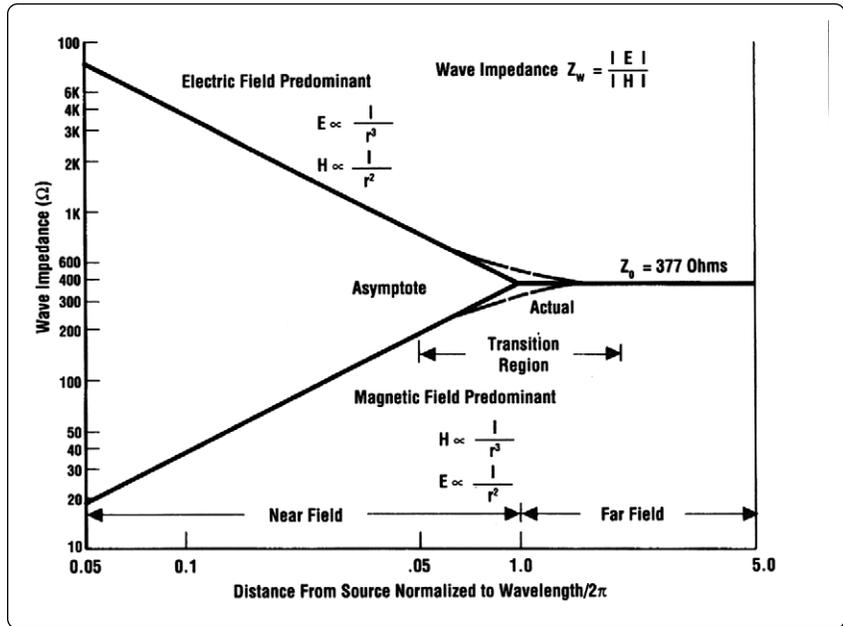


Figure 5. Wave impedance vs. distance from source

Several other factors complicate near field measurements. As we have just seen, there is no fixed ratio between the E and H fields.

We may measure different field strengths depending on the type of antenna we use. In general, the wave fronts are not plane waves in near field, but are curved. The antenna calibrations which are done in far field, where the wave fronts are plane waves, therefore will not hold. Very small dipoles can be used to make accurate near field measurements. They will minimize the field curvature effect. But they sacrifice a lot of sensitivity due to large mismatch or loss in the matching network. In practical leakage measurements, we have additional complicating factors: The nature of the source, whether electric or magnetic, may be unknown. And most likely there will be multiple sources for the leakage.

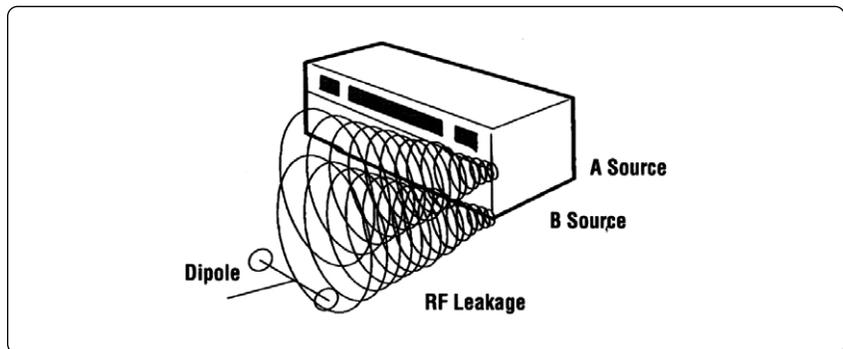


Figure 6. Sources complete measurement

Historically, we have used a two turn loop antenna, to measure rf leakage from our signal generators. Due to radios becoming more sensitive and operating at higher frequencies, it has become inadequate for rf leakage measurements. Its sensitivity is no match for a resonant dipole about 20 to 30 dB less sensitive in far field measurements. It is hard to assign a number for comparative sensitivity due to its unflatness and lack of consistency among units. At frequencies higher than several hundred MHz, the handle has been found to be a better antenna than the loop portion, This make it sensitive to small changes in orientation and hand position. Even reflections off the user's body change the response.

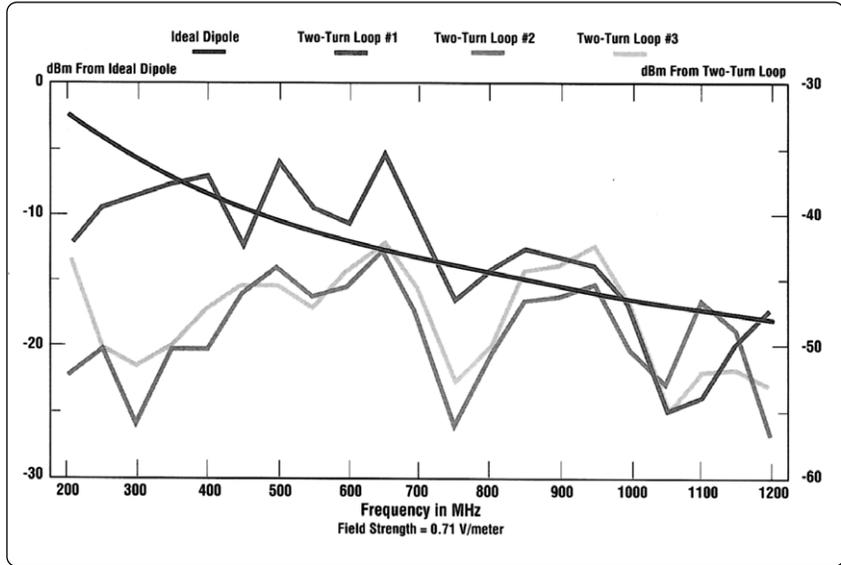


Figure 7. Two turn loop far field responses vs. ideal dipole

To improve our sensitivity and flatness for measuring our low leakage signal generators, we have decided to use resonant dipoles. The dipoles have spacers, 25 mm radius, to set a standard measuring distance. Several tests have shown this to be a good choice as we shall see.

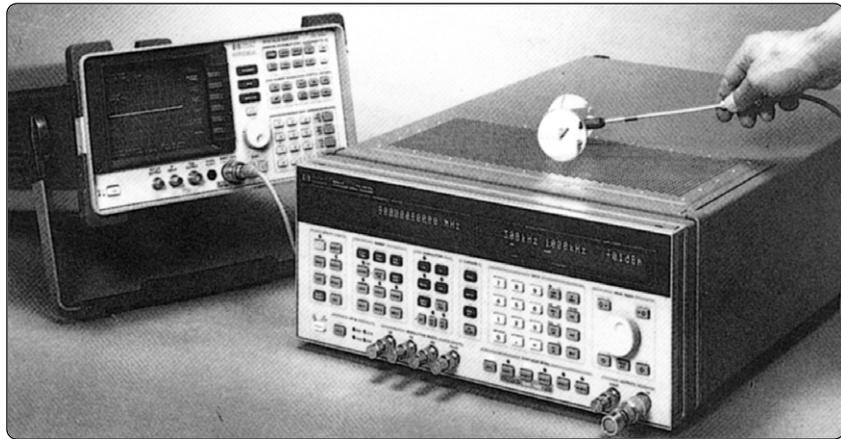


Figure 8. Photo of dipole

## Correlating receivers and dipole antennas

Our goal is to build signal generators with leakage low enough such that it cannot be detected by receivers placed a given distance from the signal generators. The problem is then to generate a field that a receiver can barely detect, and then substitute a resonant dipole in this same field. The signal level out of the dipole is then measured.

This is first done in the far field where all the “nice” properties hold. The test is repeated in the near field. The signal levels from the dipole in far field and near field are then compared. If they are equal, we have established equivalency between dipole field strength measurements and radio sensitivities.

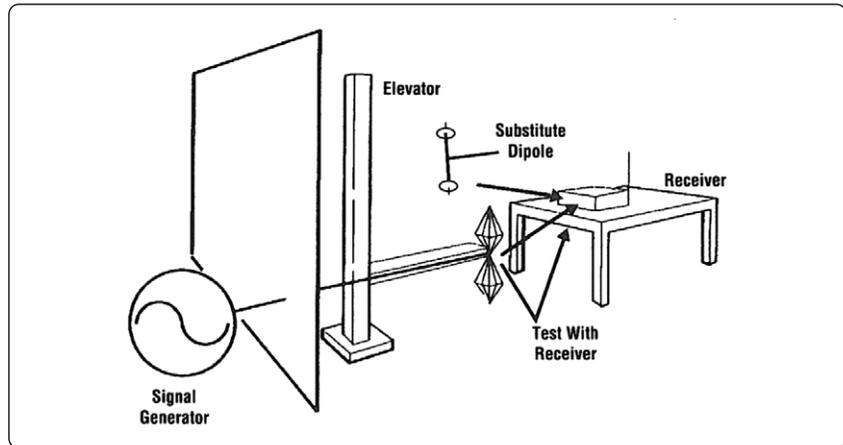


Figure 9. Calibration test between dipole and receiver.

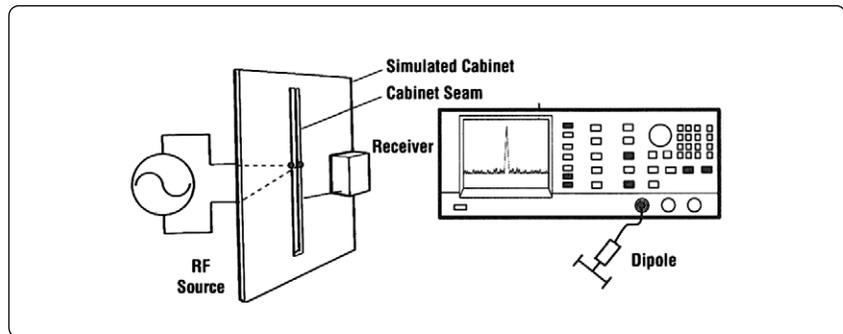


Figure 10. Leakages simulation of cabinet with seam

First, the far field tests were carried out. In a semianechoic chamber (all surfaces covered with RF absorber except the floor) a transmit antenna was placed on an elevator mechanism. Various radios and pagers, 146 MHz to 1260 MHz were placed about two meters from the transmit antenna, about 1 meter above the floor. The height of the transmit antenna was adjusted for maximum received signal in the receiver. The signal strength was adjusted at the source so that the modulation tone was just perceptible from the receiver. A dipole was then substituted at the location of the receiver, and the received signal measured.

Next, a circuit fed slot antenna was constructed. A receiver was placed 25 mm from the slot antenna. The receiver was rotated for maximum signal strength, and the signal strength was again adjusted so the modulation tone was just perceptible from the receiver. A dipole with 25 mm radius spacers was then scanned over the slot antenna. The received signal strength was then compared to that previously found in the earlier test, and was with 2 dB.

The tests so far have compared the relative sensitivity of dipoles versus receivers. While this is sufficient for accurate testing of RF leakage, we wanted to determine the absolute sensitivity of the dipole antennas.

For the far field, this is given by equation 14:

$$(14) \quad E_r = 0.159 |E| \lambda$$

How well would the dipole do in absolute terms in measuring the radiation from a slot antenna in the near field? A test was done at 500 MHz to find out. A 600 MHz resonant circuit fed slot antenna was driven with 0 dBm RF power. A resonant dipole was placed across the slot, 25 mm away. The received power was measured. The power was -3 dBm, indicating a rather efficient coupling of power. We believe that this is due to capacitive coupling of the potential difference across the slot to the antenna poles. This unexpected result shows that a dipole, an electric field antenna, is good even for measuring radiation from magnetic sources like a slot antenna.

This success inspired a similar experiment. A 500 MHz resonant loop was constructed. This too is a magnetic radiator. The loop was driven with 0 dBm RF power. When a dipole was placed 26 mm from the loop, the power received was -6 dBm. In this case, the coupling was not quite as good, but still quite acceptable to indicate when leakage should occur.

Thus, experimentally, the dipole has shown itself to be sensitive, even to magnetic sources which might be found in signal generators.

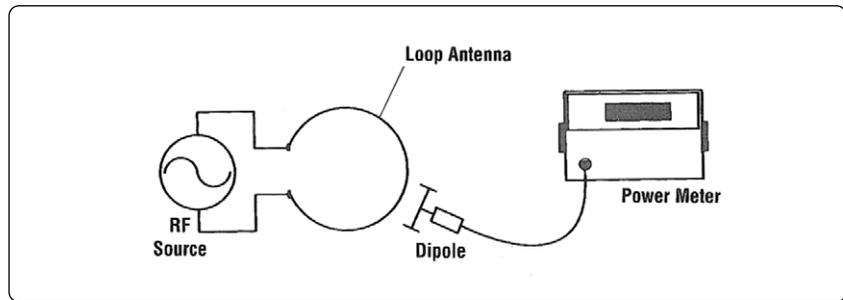


Figure 11, Absolute measurement loss – loop antenna to dipole antenna

## Sources of leakage from a signal generator

If the cabinet containing a signal generator were a welded metal box, for all practical purposes there would be no leakage. Of course, this is impractical, for we must be able to get signals out of the box, and modulation, control, and power into it. In addition, in order to service it, the covers must be removable, and not welded. These considerations cause us to compromise the ultimate shielded box.

Wires and cables passing through a shielding enclosure are commonly called penetrations. If an insulated wire is simply passed from the inside to the outside of a shielded box through a hole, it will conduct out the rf fields present in the box and then radiate them on the outside. Even a terminated coax cable, if the outer conductor is not grounded at the hole, can conduct out the fields.

For low leakage along seams, the ideal is a continuous low impedance contact. In practice, this is difficult to achieve. The goal can be approached with some reduction in performance by using closely spaced periodic contact points of the sort provided by multiple screws or contact fingers.

In the critical front seam area of our low leakage signal generators, the periodic contact between the instrument cover and a front bulk-head is supplied by screws spaced 26 mm apart. This is about .08 wavelength apart at 1 GHz.

The rear area was deemed less critical for leakage, as it is further away from the area where radio testing would be done. Here, a spiral strip is used to make contact between the instrument cover and rear panel assembly.

The cover is a one piece sleeve, with a spotwelded seam. The air vent hole area in the cover is backed by aluminum mesh riveted to the cover. The mesh lowers rf leakage through the vent holes. As mentioned earlier, behind the front panel, there is a bulkhead which supplies the shielding in front. The instrument cover makes contact with this bulkhead, forming a shielded box. Through this bulkhead pass the rf output cable and several modulation input cables. Also passing through the bulkhead is a digital control cable for the front panel keyboard and display.

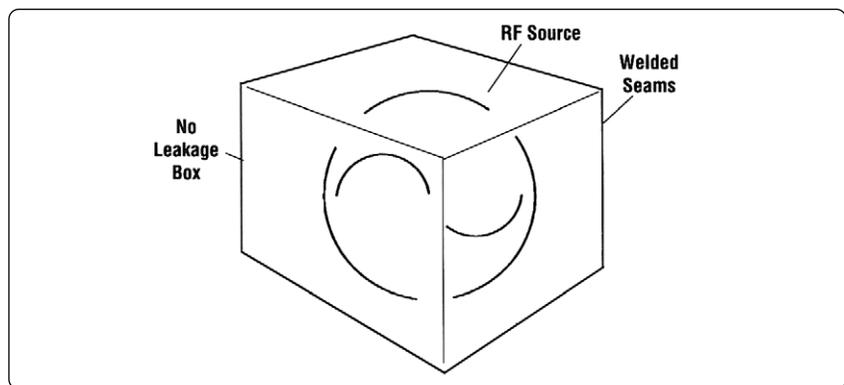


Figure 12. Perfect cabinet

The solution for coax cables then is to ground the outer conductor at the bulkhead wall. This is done by using a bulkhead connector in the case of the modulation inputs. The rf output cable is soldered to a grounding plate which is then bolted to the bulkhead. The digital control cable for the front panel keyboard and display is brought through the bulkhead with a shielded connector having an internal low pass filter for each line.

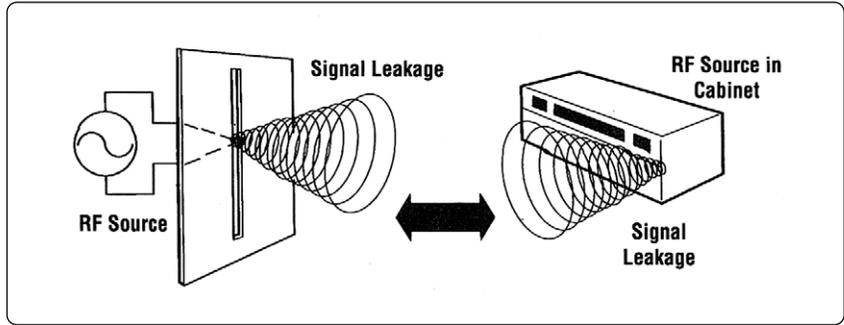


Figure 13. Simulation equivalents to signal generator cabinet

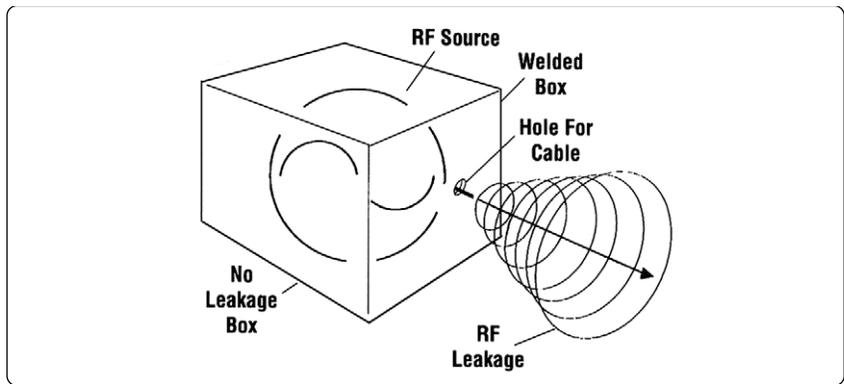


Figure 14. Wire causes cabinet leakage

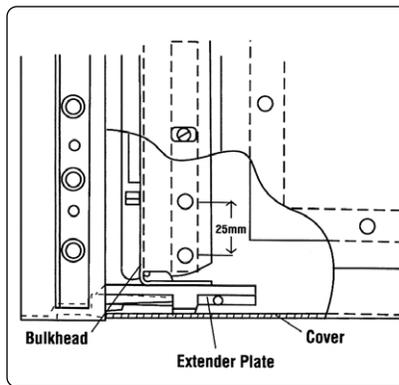


Figure 15. Detail of cabinet optionion 10

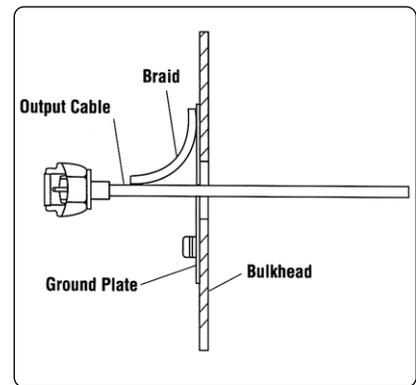


Figure 16. Rf leakage wire passage option 10

## Dipole discussion

What are some of the sources of inaccuracy with dipole measurements?

1. If the dipole impedance is not matched to the measurement system impedance, there will be loss due to mismatch. The magnitude of this loss is:

$$\text{Mismatch loss in dB} = -10 \log(1-r^2)$$

where:

$r$  = reflection coefficient magnitude  
log = base 10 logarithm

As an example, 73 ohms is the impedance of a dipole in free space, slightly shortened from one half wave length. If this is connected to a 50 system, there will be 0.15 dB mismatch loss.

2. A more general form of mismatch loss is mismatch uncertainty. This problem is often seen in systems with cables, where the angle of the reflection coefficient will rotate with frequency. The limits of this uncertainty are:

$$\text{Mismatch uncertainty} = 20 \log(r_1 r_2)$$

$r_1$  = reflection coefficient magnitude of source (antenna)

$r_2$  = reflection coefficient magnitude of load (receiver)

Analysis shows that resistive loss in a half wave dipole is negligible. Let the pole elements be made of copper. Assume all the current flows uniformly on the surface of the pole to a thickness of one skin depth. The surface resistivity is:

Surface resistivity =

where  $s$  = conductivity  
 $d$  = skin depth

Skin depth is as follows:

$$d = \frac{1}{\sqrt{\pi f \mu \sigma}}$$

Where  $f$  = frequency  
 $\mu$  = permeability

Thus;

Surface resistivity =

$$\sqrt{\frac{\pi f \mu \sigma}{\sigma}} \quad \frac{\text{ohms}}{\text{square}}$$

For copper,  $\mu = 4\pi \times 10^{-7}$  henry/meter (permeability of free space),

$s = 5.8 \times 10^7$  siemens at room temperature.

Surface resistivity =

$$\sqrt{\frac{3.1416 \times 225 \times 10^6 \times 4 \times 3.1416 \times 10^{-7}}{5.8 \times 10^7}} = 3.9 \frac{\text{milliohms}}{\text{square}}$$

Assume the pole diameter is 3.58 mm (.141 inch). The circumference is 11.25 mm. For poles of circumference 11.25 mm:

Resistance per mm =

$$3.9 \times 10^{-3} = .35$$





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