RELEVANCE OF ELECTROMAGNETICS IN WIRELESS SYSTEM DESIGN

REFERENCE: Tapan K Sarkar, Eric L Mokole and Magdalena Salazar Palma, "Relevance of Electromagnetics in wireless system design", IEEE Aerospace and Electronic Systems Magazine, Vol. 31, No. 10, pp. 8-19, 2016

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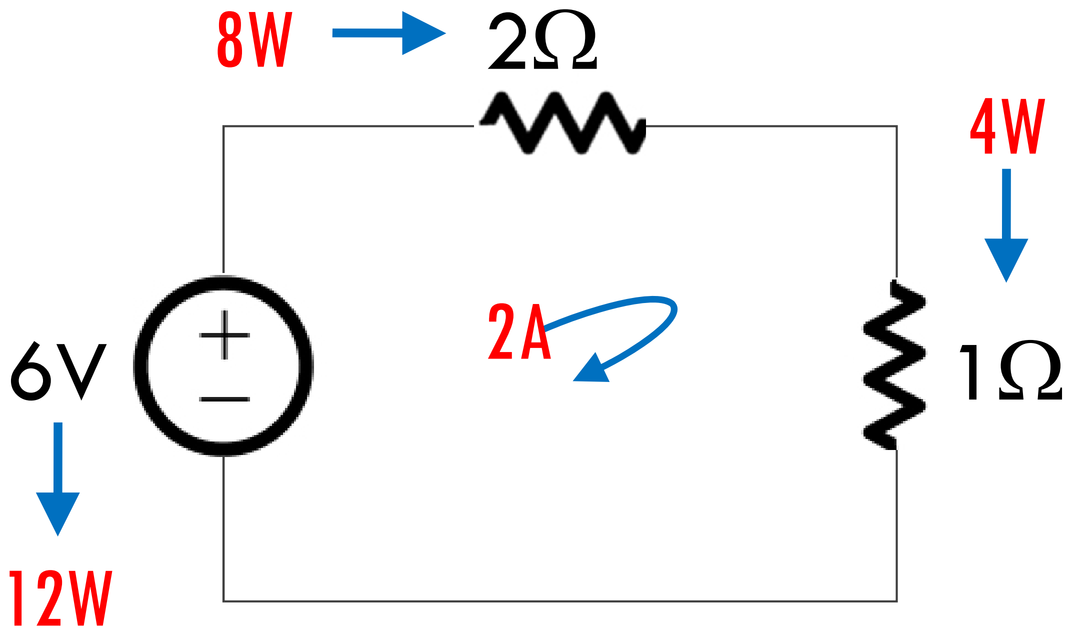
- Principles of Electrical Engineering
- ANTENNA: Its Relevance in
  - Transmission
  - Reception
  - Propagation
- Ultrawideband Transmission & Reception without Distortion
- CHANNEL CAPACITY

- **CONVOLUTION and CORRELATION**

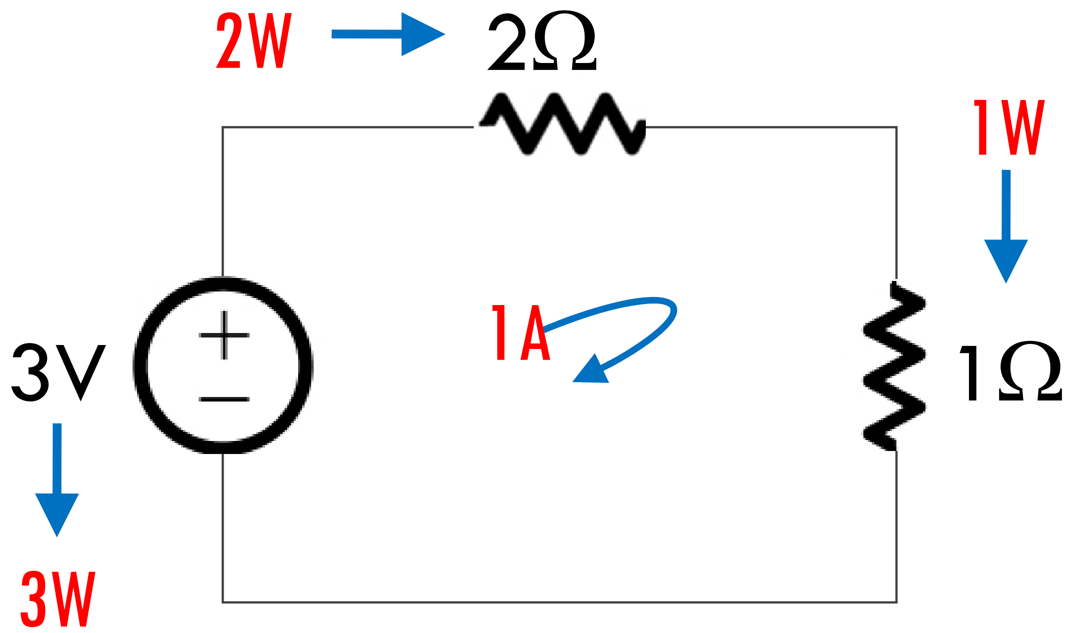
- **Principle of Superposition APPLIES to CONVOLUTION**
  It DOES NOT APPLY to CORRELATION

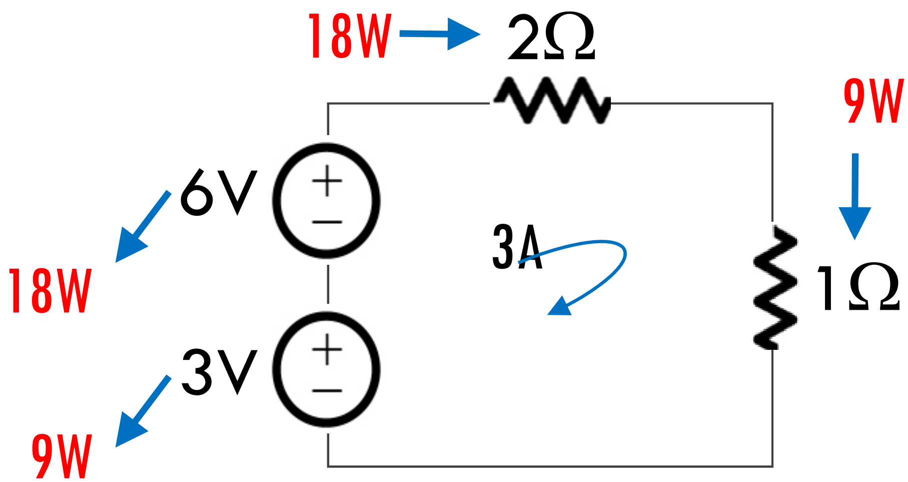
- **Maximum Power Transfer VS EFFICIENCY**

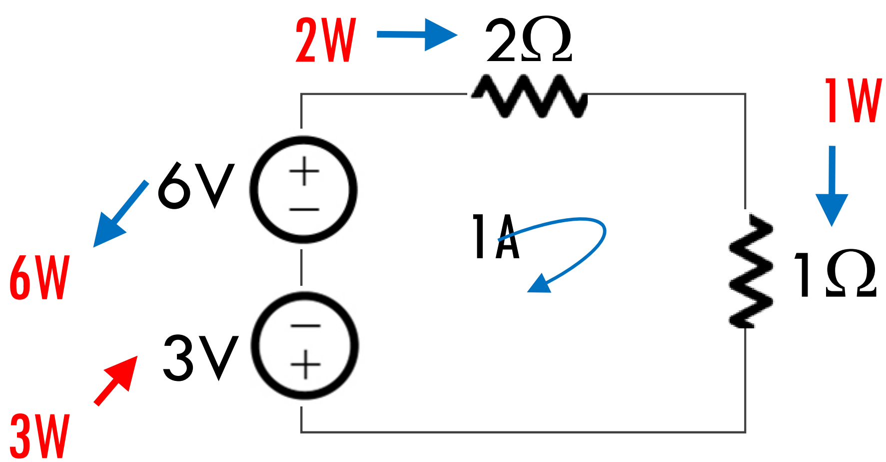
Case A



Case B



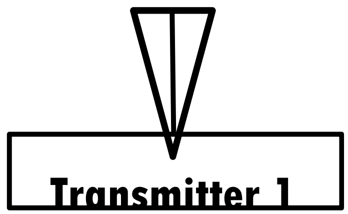




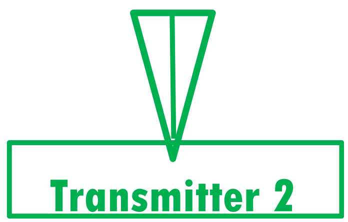
The two cases are COMPLETELY Different

Superposition of POWER DOES NOT HOLD in Electrical Engineering.

Superposition of voltages and currents are still valid



100W/m^2



1W/m^2

There will be constructive and destructive interference. So what will be the variation of power density in space?

(100 ± 1) W/m^2 ??

$(121 \leftrightarrow 81)$ W/m^2

A 40% variation even though the input power density variation is ONLY 1%!?

WHAT IS AN ANTENNA ?

A device whose primary purpose is to *radiate* or *receive* electromagnetic energy

WHAT IS RADIATION ?

Far Field (Fraunhofer region > $2L^2/\lambda$)

- the fields are transverse
- the shape of the field pattern is independent of the distance

WHAT IS THE NEAR- FIELD (FRESNEL REGION) ?

- ★The near field is in the region $D < 2 L^2/\lambda$
- ★ Near Field: power is complex (need both *E* & *H*)
- ★Far Field – Real Power: need either *E* or *H*

PROPERTIES OF THE NEAR FIELD

For a Dipole →

$E_z = -j30I_m[\exp(-jkR_1)/R_1 + \exp(-jkR_2)/R_2 - 2\cos(kH)\exp(-jk\rho)/\rho]$

The near field can never be zero for a dipole!!!!

Only the far field has pattern nulls!!

THIS IS TRUE FOR ANY ANTENNA

WHAT IS THE FAR- FIELD?

$D > 2\,L^2/\lambda$: L – Antenna region

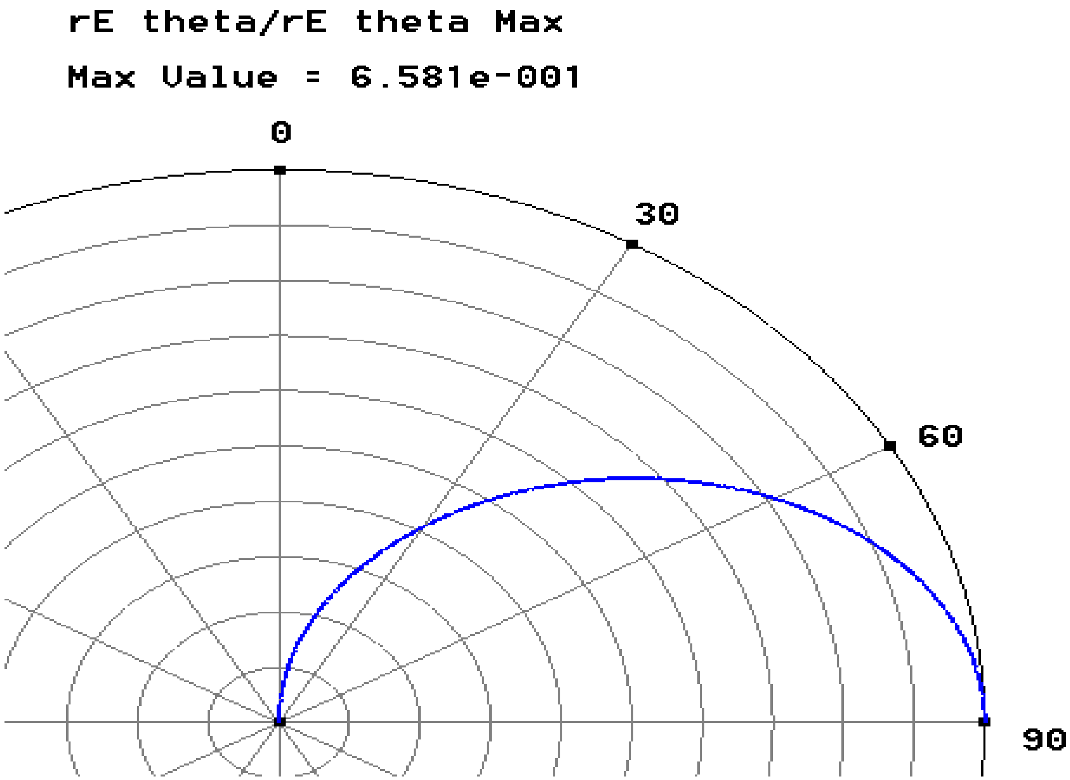
What is the far field of a half wave dipole for λ =0.3m (1 GHz) ?

$\rightarrow 2\,L^2/\lambda = 2\times0{,}15\times0.15/0.3=0.15$m

WHAT IS THE FAR FIELD WHEN THE HALF WAVE DIPOLE IS 20 M ABOVE AN INFINITE GROUND PLANE AT λ =0.3 M (1 GHZ)?

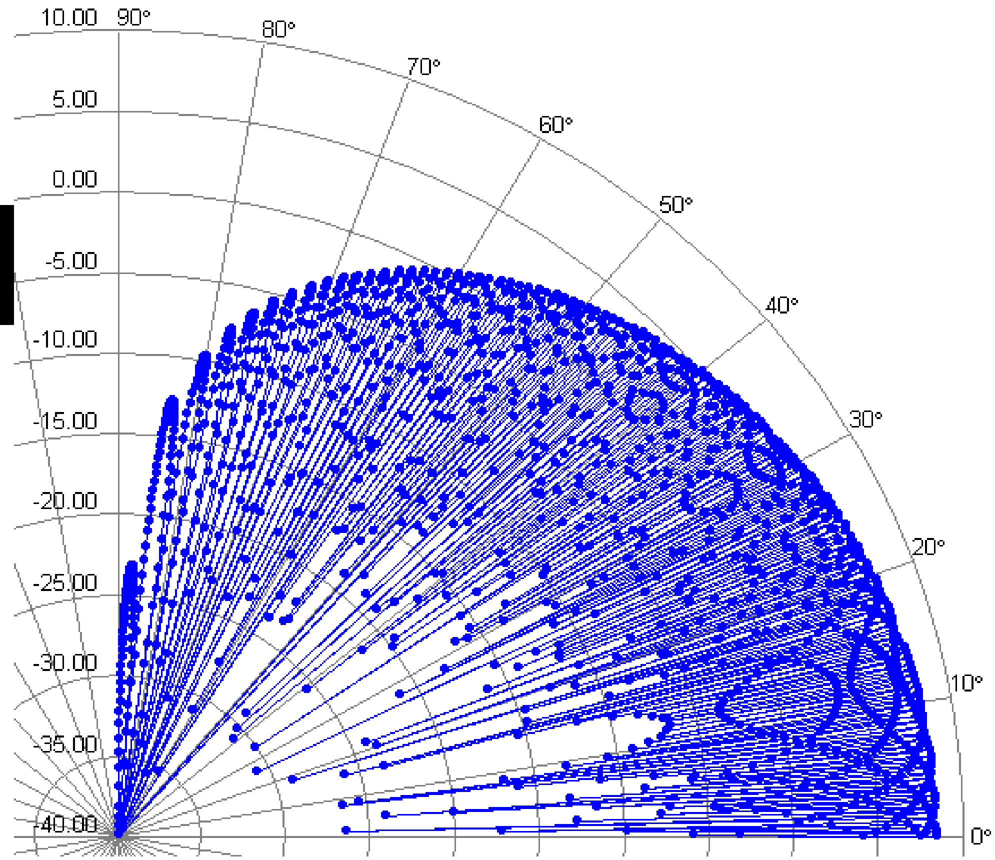
Equivalently if the dipole is on the top of a H =20 m tower above a perfect ground plane?

$\rightarrow 2\,(2H)^2/\lambda\ 2\times40\times40/0.3=10{,}666 \approx 10.6$km

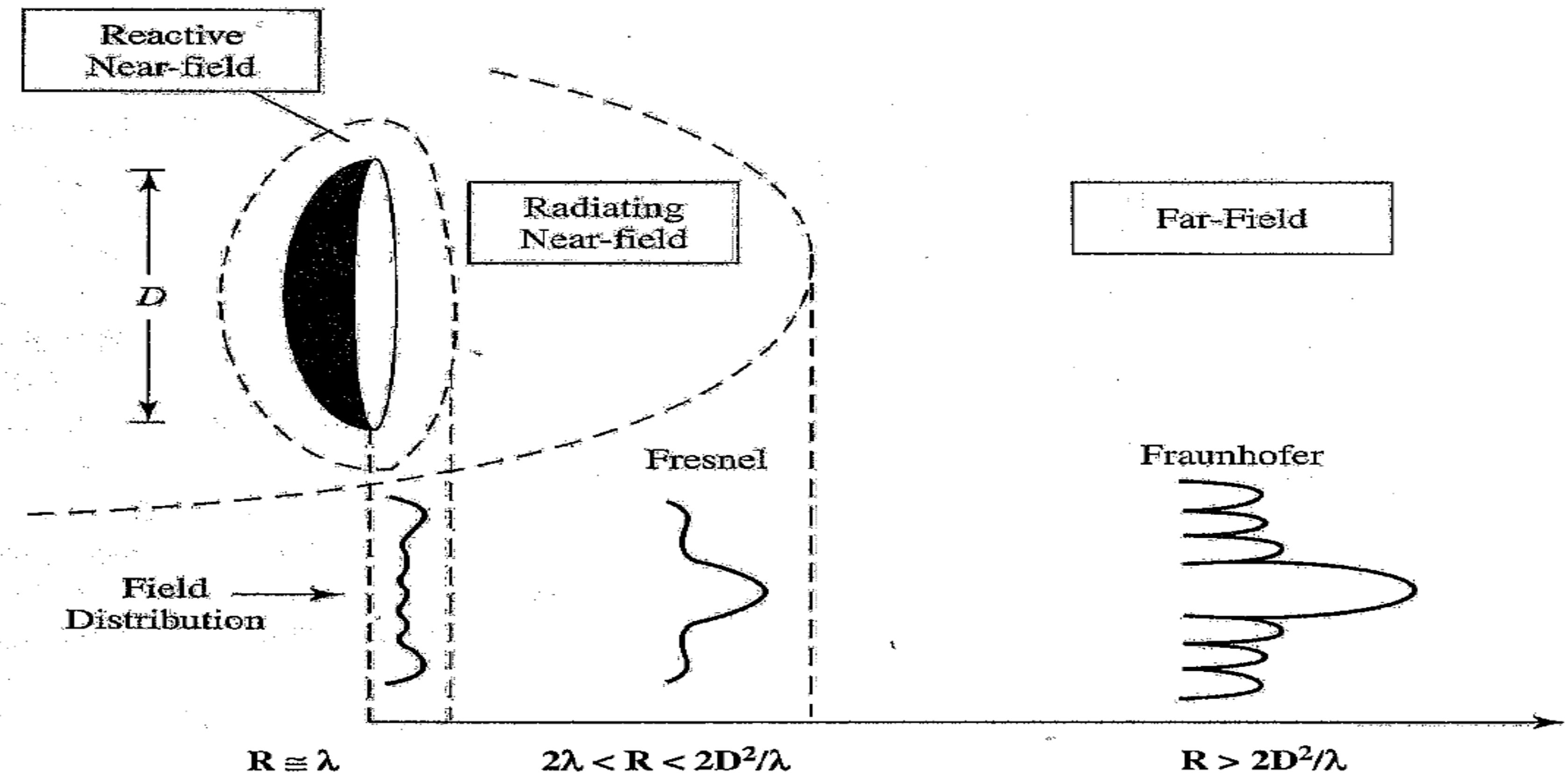


The radiation pattern of a half wave dipole in free space (only one fourth shown)

[redacted]

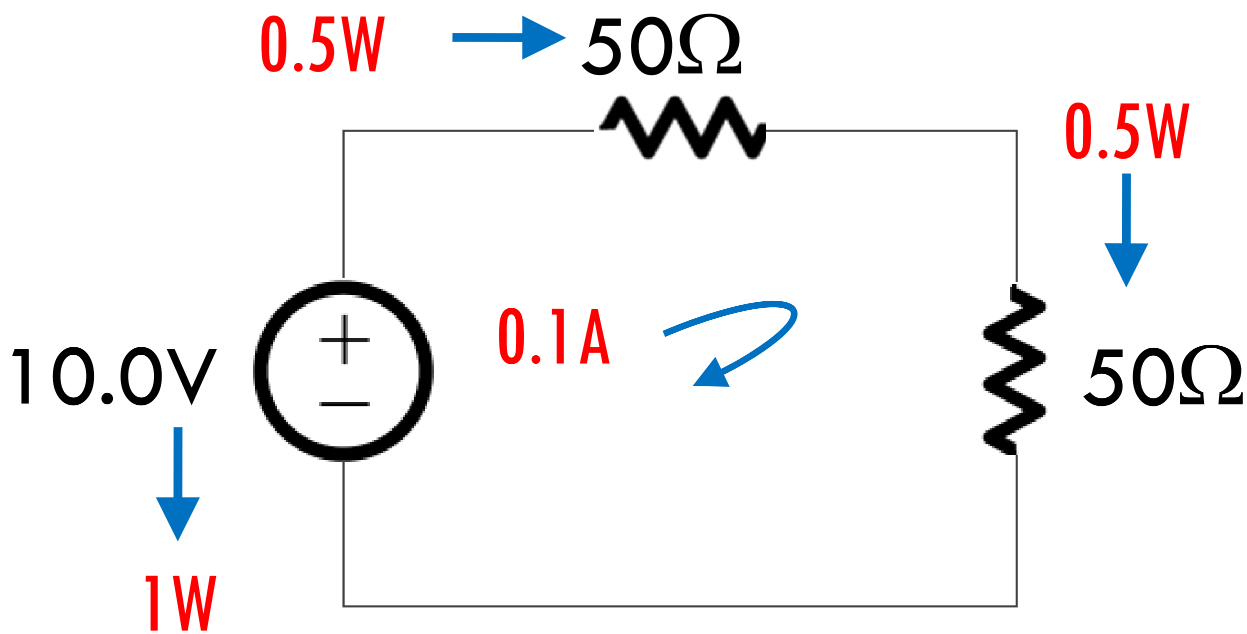

10.00
5.00
0.00
-5.00
-10.00
-15.00
-20.00
-25.00
-30.00
-35.00
-40.00
90°
80°
70°
60°
50°
40°
30°
20°
10°
0°

Field Regions Around An Antenna



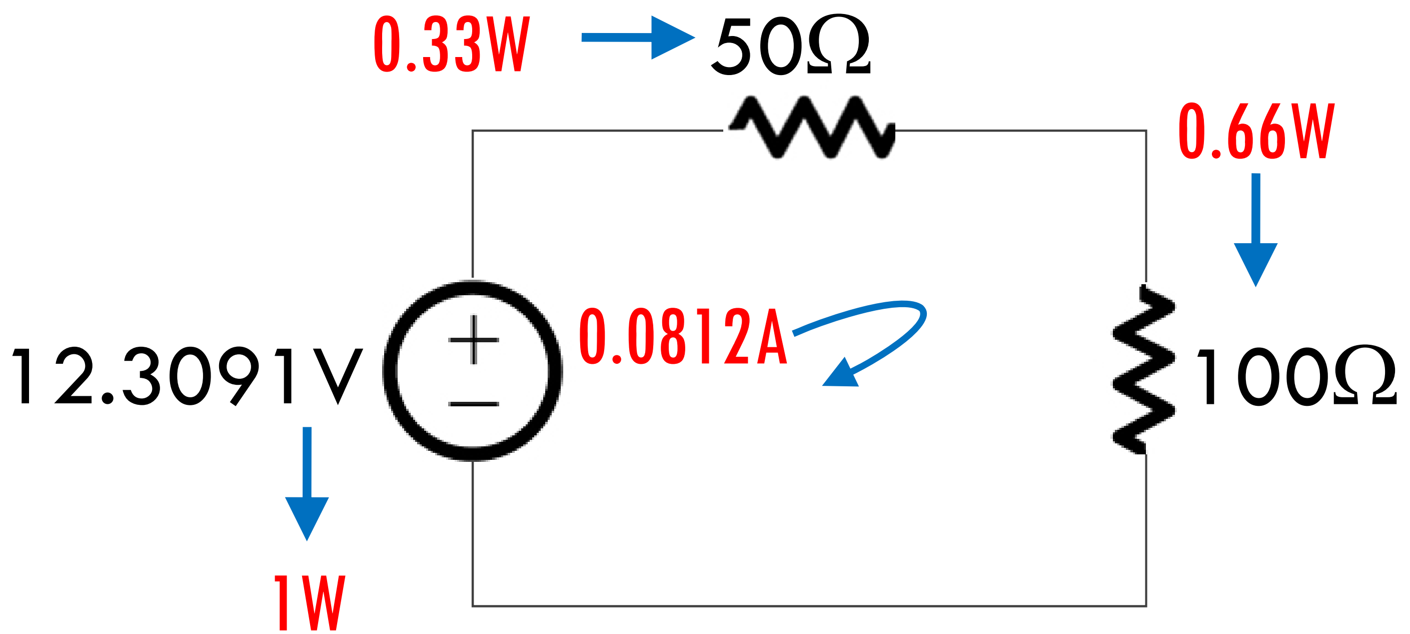
This is why SDMA has never worked in a near field scenario!

MAXIMUM POWER TRANSFER



Equal power dissipated in each load: As expected

WHAT HAPPENED TO MAXIMUM POWER TRANSFER??

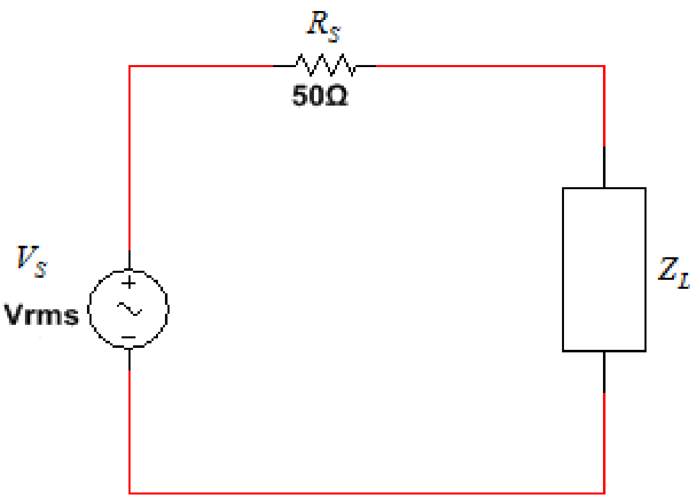


More power is dissipated in the load!!

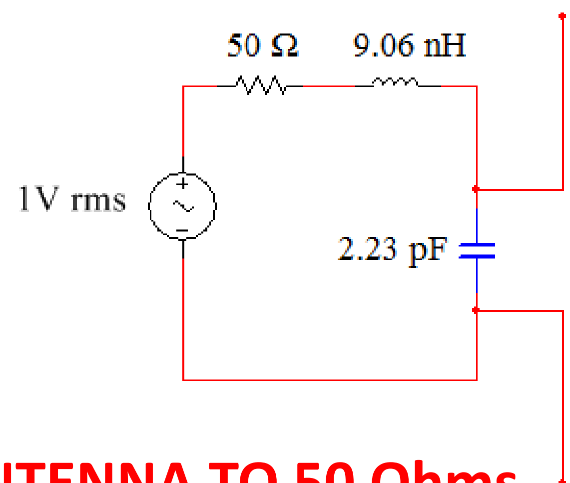
APPLICAITION RELEVANT TO ANTENNA MATCHING

Simulate a z-oriented half-wave dipole with 1 V rms excitation at the center point, the source internal resistance is 50 ohms, the frequency is 1 GHz, and the wire radius is 1 mm. **Different** matching network exists between the source and the antenna. The input impedance of the antenna is

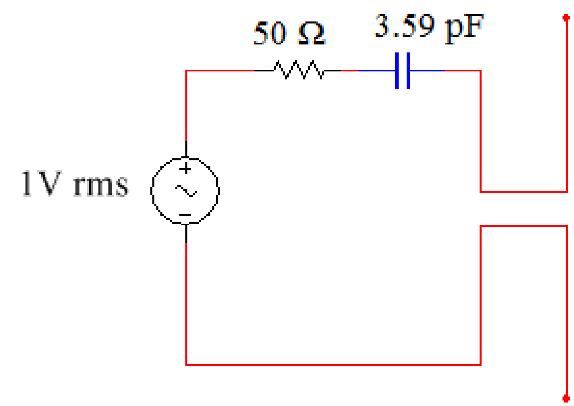
$$z_{in} = 93.84 + j44.35\ \Omega$$



CASE A: UNMATCHED



CASE B: MATCH THE ANTENNA TO 50 Ohms



CASE C: CANCEL THE EFFECT OF THE IMAGINARY PART

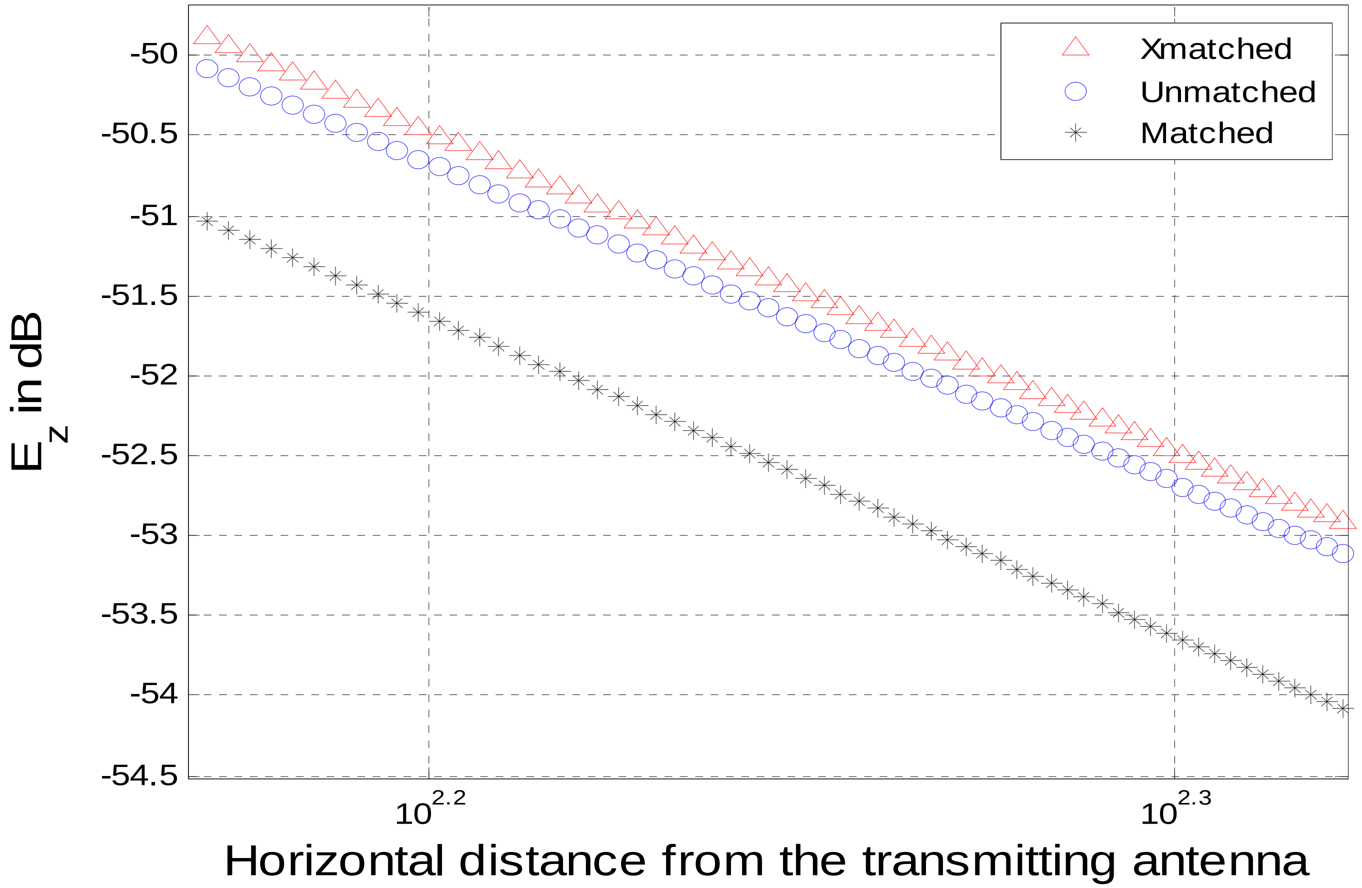
SAME INPUT POWER FOR ALL THREE CASES: 1W

Analysis using antennas

The simulation results are summarized below:

Case	$\lvert P_{in} \rvert$ in mVA	P_{out} in mW	$\eta : Efficiency$
Matched	10	5	0.5
Unmatched	6.644	4.142	0.623
Xmatched	6.952	4.536	0.652

Analysis using antennas



Analysis using antennas

Case	$\|P_{in}\|$ in mVA	P_{out} in mW	η
Unmatched 0.75λ	1.156	1.02	0.88
Unmatched 0.65λ	2.156	1.48	0.69

Unmatched Transmitting Scenario

- First, we consider the antenna to be in free space and not matched. We simulated a z-directed dipole antenna with four different lengths (1 GHz, 1V rms):

0.01λ, 0.1λ, 0.25λ, and 0.5λ

- Which of this dipole antenna will radiate the most electric field?

Unmatched Transmitting Scenario

- In our simulations,

  the input complex power magnitude for the 0.01λ antenna was 35.083 μ VA,

  the input complex power magnitude for the 0.1λ antenna was 363.3 μVA,

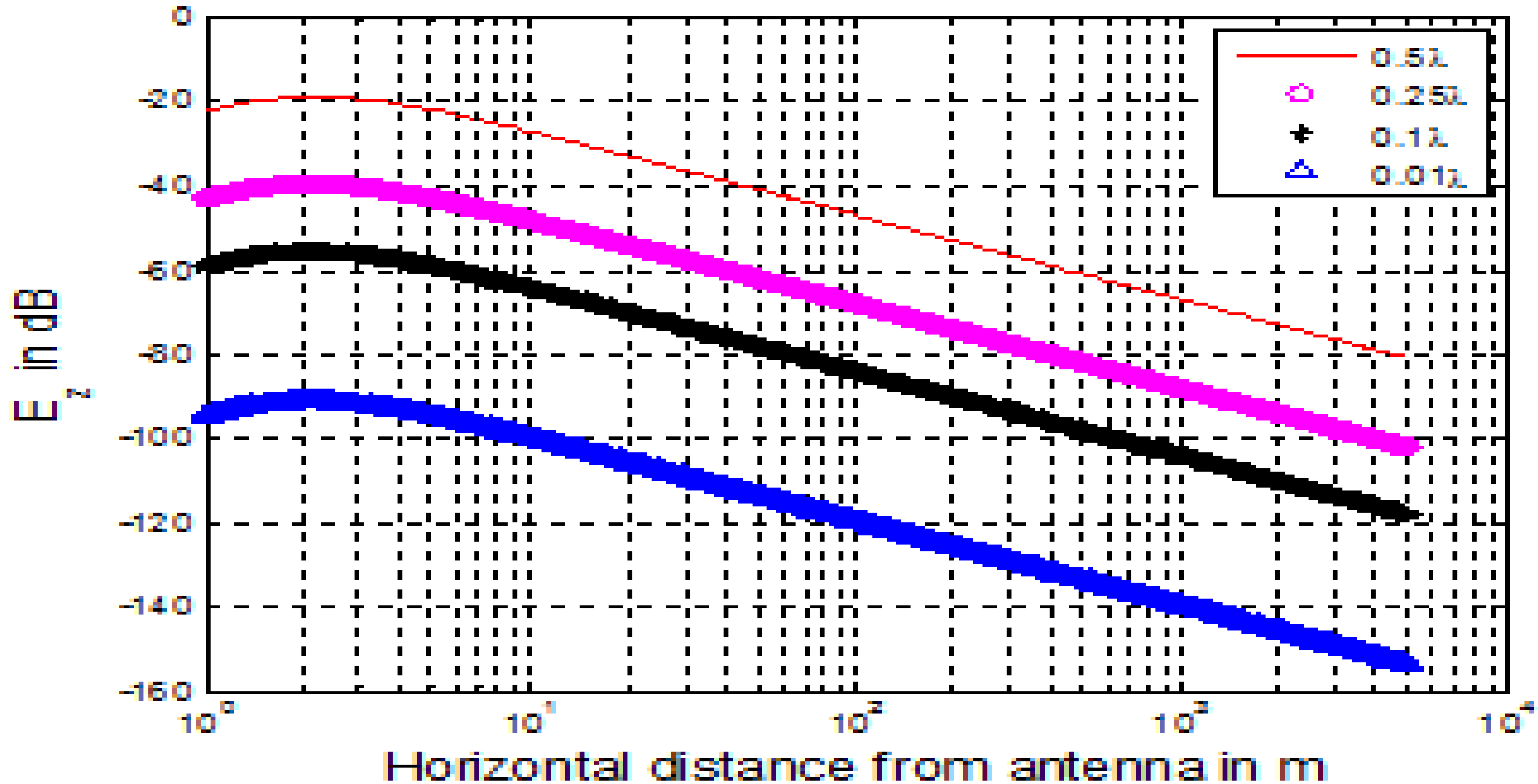
  the input complex power magnitude for the 0.25λ antenna was 1.131 mVA, and

  the input complex power magnitude for the 0.5λ antenna was 4.87 mVA.

- It makes sense that the input power for the half wave dipole is the largest as the antenna is almost self-matched compared to the other lengths. We need to scale the fields' strengths according to the square root of the input complex power magnitude ratios, basically we multiplied the magnitude of the z-component of the electric field for the 0.25λ by 2.075, for the 0.1λ by 3.66, for the 0.01λ by 11.782, and for the 0.5λ by 1.

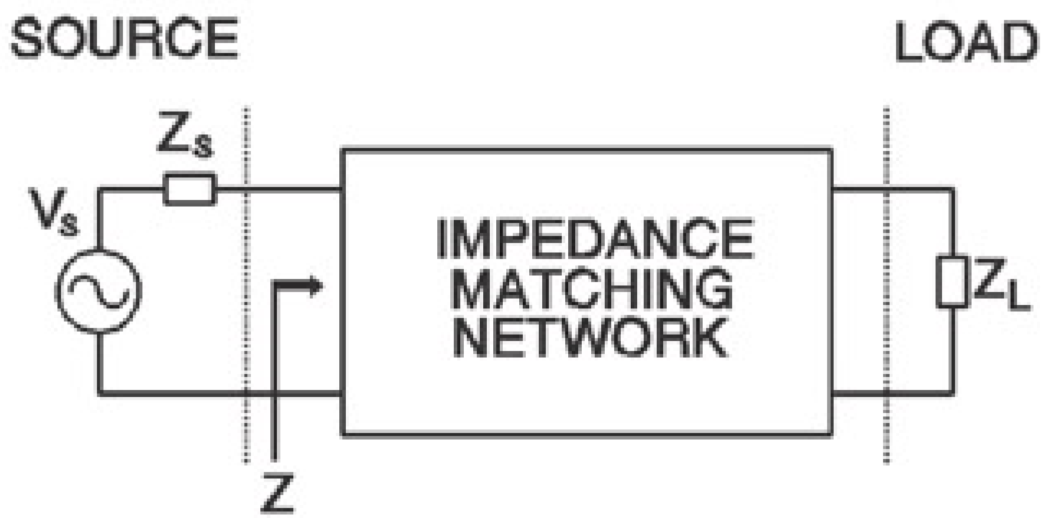
Unmatched Transmitting Scenario

The magnitude of the z-component of the electric field in dB for the various dipole lengths



Matched Transmitting Scenario

- The second scenario we study is when the antenna is matched to the real part of the input impedance using a lossless matching network (inductor or capacitor with zero resistance).

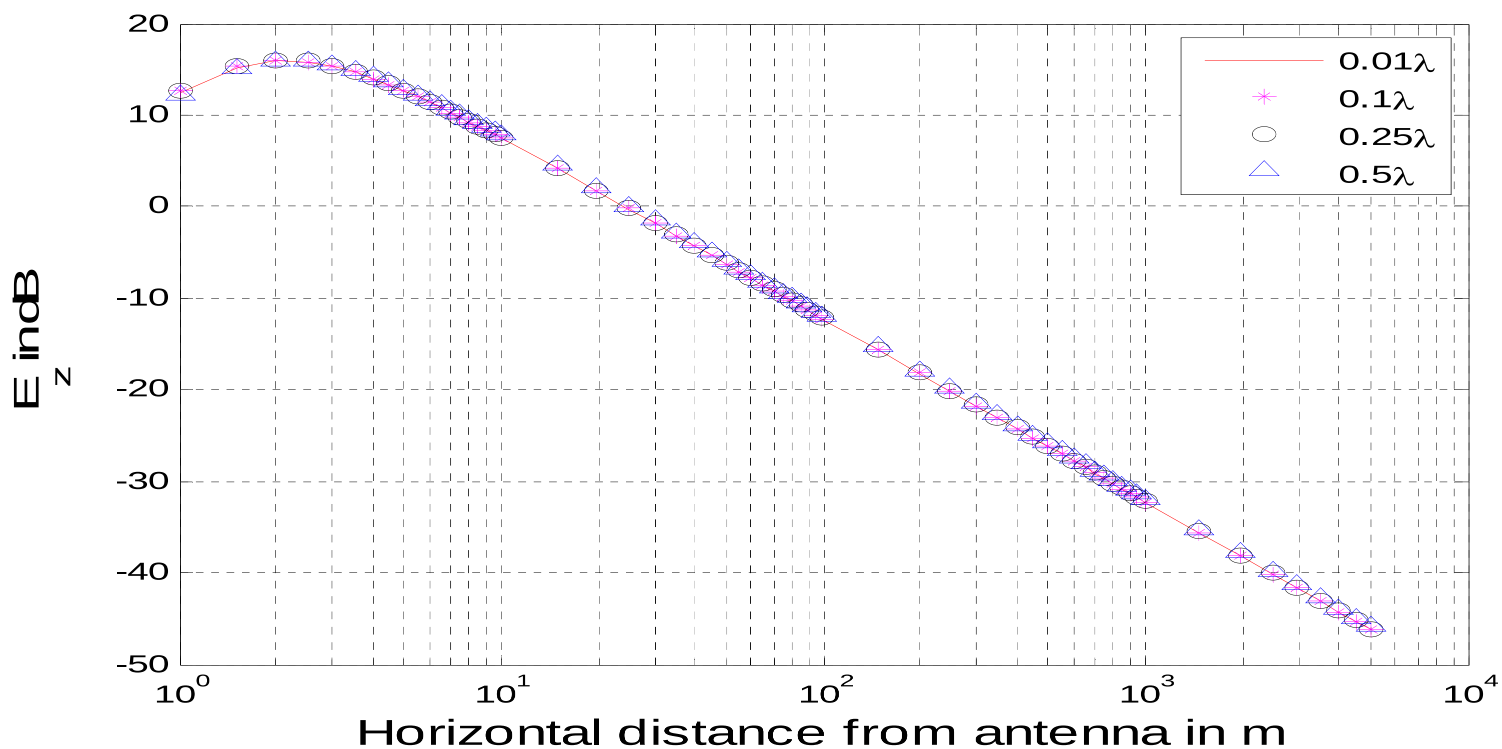


Matched Transmitting Scenario

- We consider z-directed dipole antenna in free space with four different lengths: 0.01λ, 0.1λ, 0.25λ, and 0.5λ.
- The input power, in this case, was real so we made sure that the real input power is the same for the different dipole lengths so as to be able to make a fair comparison. Instead, we scaled the electric field strengths by the square root of the ratio between the input powers as in the previous unmatched scenario.

Matched Transmitting Scenario

The magnitude of the z-component of the electric field in dB for the various dipole lengths (matched scenario).



Matched Transmitting Scenario

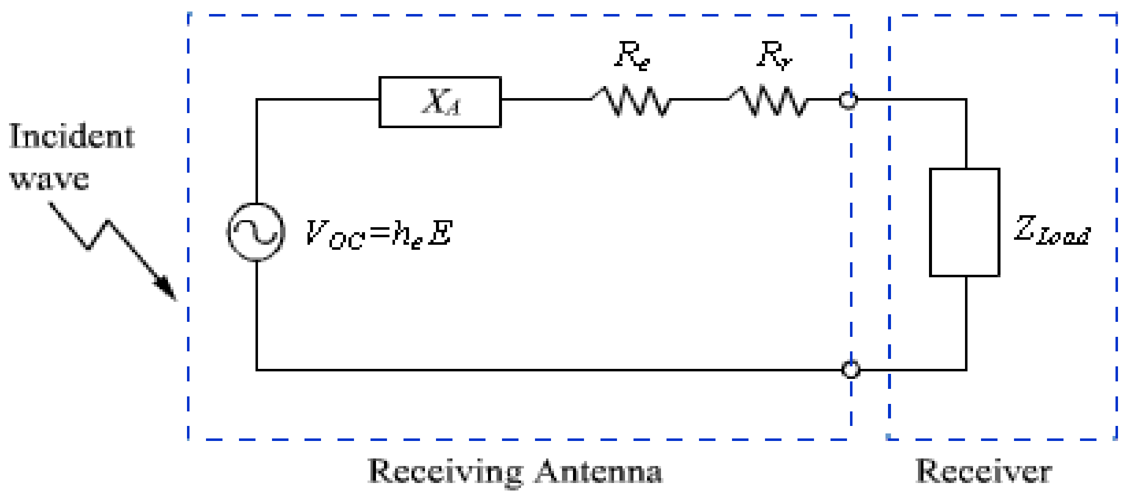
- We can clearly see that the field strength is almost the same for the various antenna lengths when the antenna is matched (the input power consists of real part only) and the difference between them does not exceed 0.3 dB.
- This means that the degradation in the antenna performance for an ESA is negligible, when operating in a matched condition.

Unmatched Receiving Scenario

- We also study the antennas in the receiving mode operating in free space.
- We consider z-directed dipole antenna with four different lengths:

$$0.01\lambda,\ 0.1\lambda,\ 0.25\lambda,\ \text{and}\ 0.5\lambda$$

- Antenna equivalent circuit (receiving mode)



Unmatched Receiving Scenario

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- We consider z-directed dipole antenna with four different lengths:

$$0.01\lambda,\ 0.1\lambda,\ 0.25\lambda,\ \text{and}\ 0.5\lambda$$

- We excited the antenna with a plane wave where we set the theta component of the electric field to be 1 V/m incident at theta=90^{o} and we set the phi component of the electric field to be 0 V/m, theta and phi are measured from the z-axis and x-axis, respectively.
- First, we study the scenario of the receiving mode of dipoles with various lengths while the antenna is connected to a 50Ω load (unmatched).

Unmatched Receiving Scenario

Length	0.01λ	0.1λ	0.25λ	0.5λ
P_r (mw)	$1.3*10^{-10}$	$1.43*10^{-6}$	$9.77*10^{-5}$	0.026
P_r(dBm)	-98.86	-58.45	-40.1	-15.85

Table 1. Received power by a 50 Ω load connected to dipole antenna with various lengths.

- Table 1 summarizes the received power in a 50 Ω load.
- From the results in Table 1, we conclude that the received power in a 50 Ω load is the largest in the case of the half wave dipole. This result is expected because the input impedance of the half wave dipole is best matched to 50 Ω compared to all other dipoles.

Matched Receiving Scenario

- For the same previous setup, we study the scenario of the receiving mode for dipoles of different lengths while the connected load at the antenna port is conjugately matched to the antenna input impedance, i.e.

$$Z_l = R_r - jX_A$$

Matched Receiving Scenario

Length	0.01λ	0.1λ	0.25λ	0.5λ
P_r (mw)	0.0279	0.0286	0.0291	0.0312
P_r (dBm)	-15.54	-15.43	-15.36	-15.05

Table 2. Received power by a load, cojugately matched to the input impedance of the antenna, connected to dipole antenna with various lengths.

- Table 2 summarizes the received power by a conjugately matched load.
- From the results in Table 2, we conclude that received power by a conjugately matched load is almost the same for all different dipole lengths. The difference does not exceed 0.5 dB in the worst case.
- This means that the degradation in the antenna performance for an ESA is negligible, when operating in a matched condition.

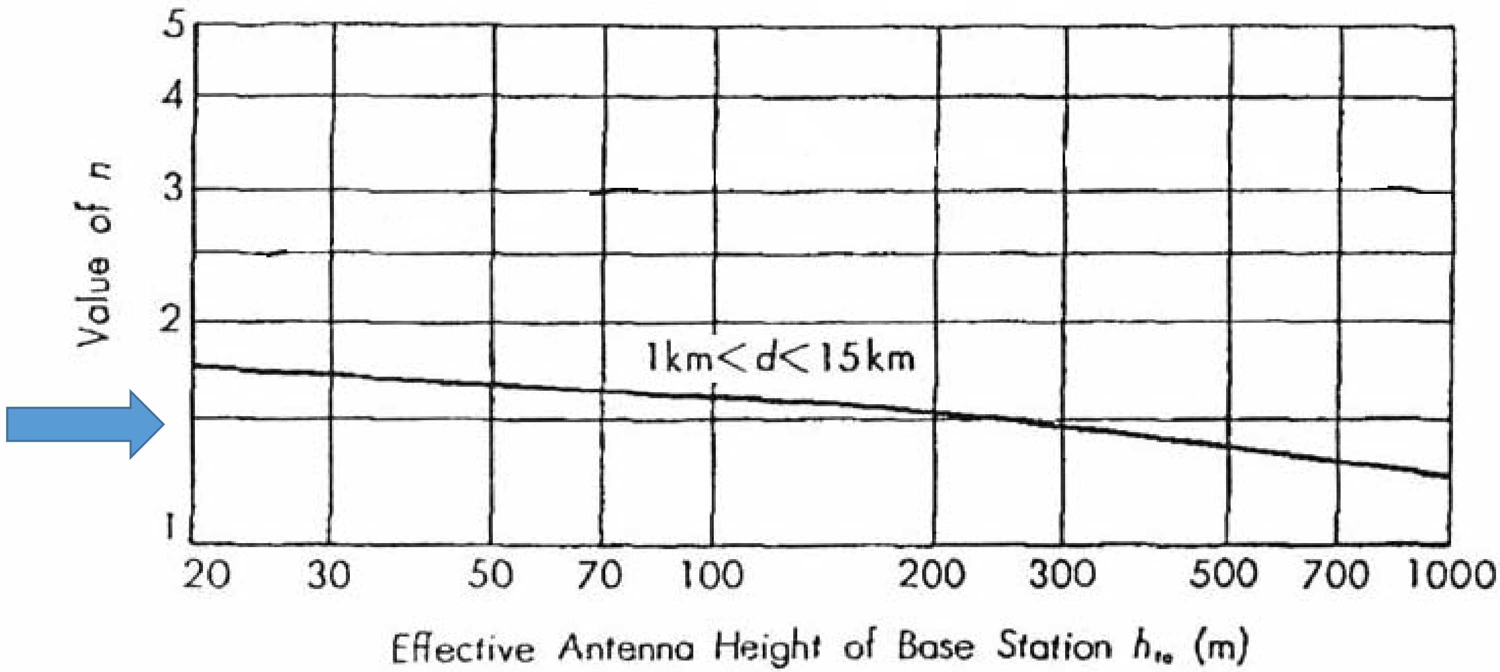
Antenna inside a prehistoric transistor radio


TP-LINK
PWR
SYS
WLAN
4
3
2
1
WAN
QSS
Advanced Wireless N Router
QSS

SURPRISES IN PROPAGATION MODELING IN WIRELESS COMMUNICATION

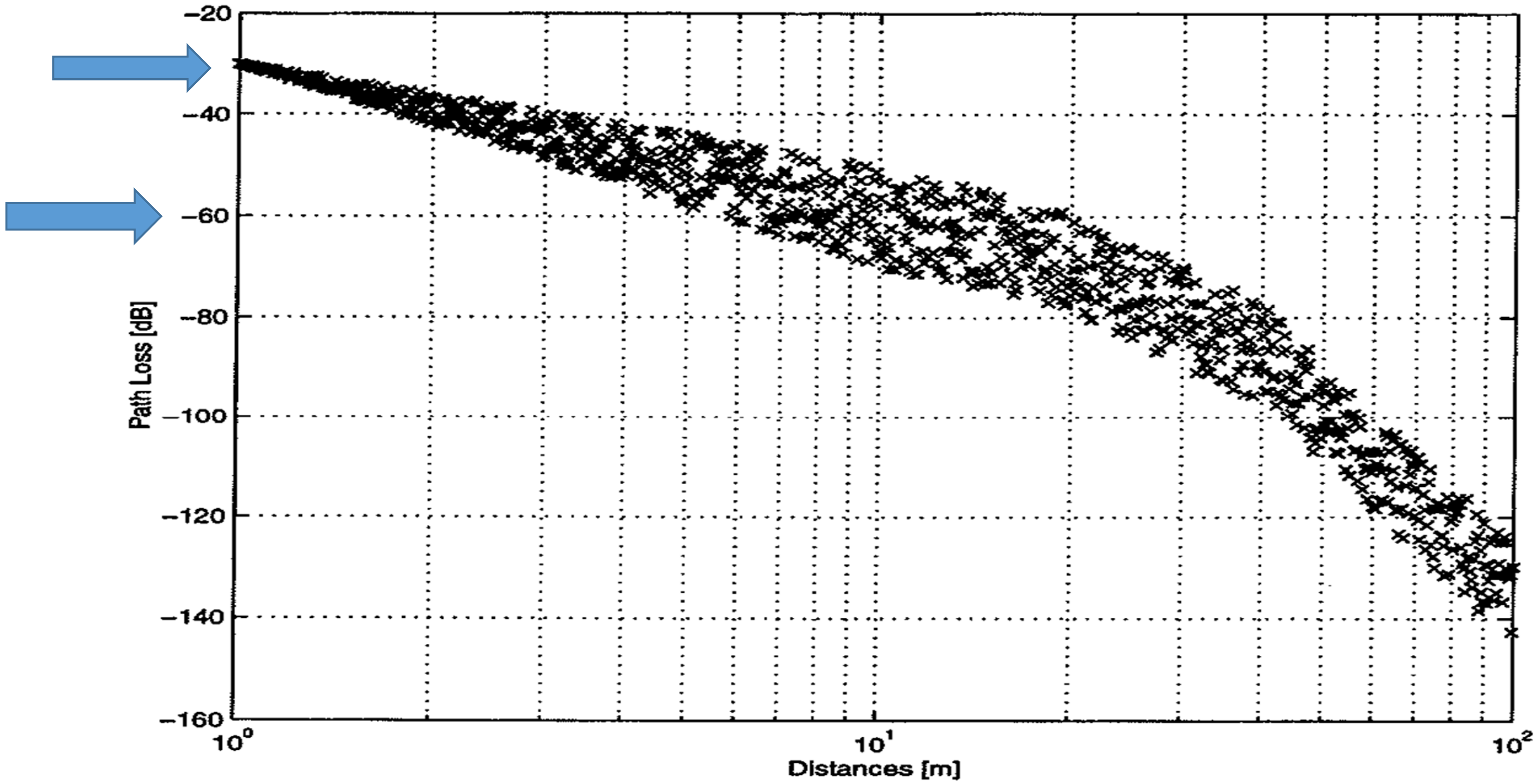
Electric field $\propto \rho^{-1.5}$; $Power \propto \rho^{-3}$ ► $10\log(\rho^{-3})$ ► $-30\frac{dB}{decade}$ $of\ distance$

n=1.5

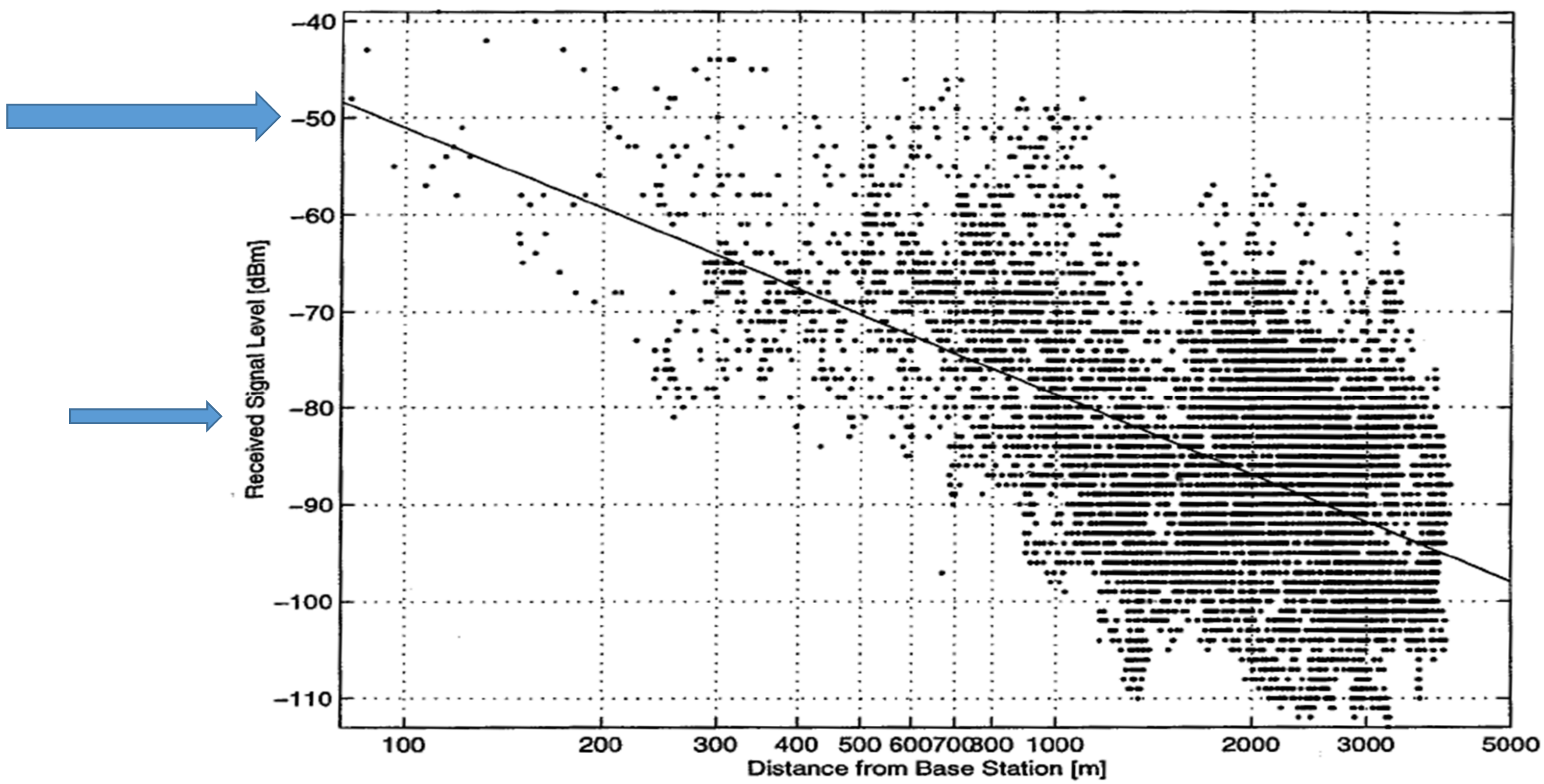


Distance dependence of median field strength in urban area ($E_m \propto d^{-n}$).

From Okumura's et al.'s historical paper



Prediction from Ericsson in-building path loss model. Reproduced by permission from Simon R. Saunders, Advances in mobile propagation prediction methods, Chapter 3 of *Mobile Antenna Systems Handbook*, Edited by: Kyohei Fujimoto, Artech House, 2008.



Empirical model of macrocell propagation at 900 MHz, the dots are measurements taken in suburban area, where as the solid line represents a best fit empirical model. Reproduced by permission from Simon R. Saunders, Advances in mobile propagation prediction methods, Chapter 3 of *Mobile Antenna Systems Handbook, Third edition*, Edited by: Kyohei Fujimoto, 2008, Artech House, Inc.

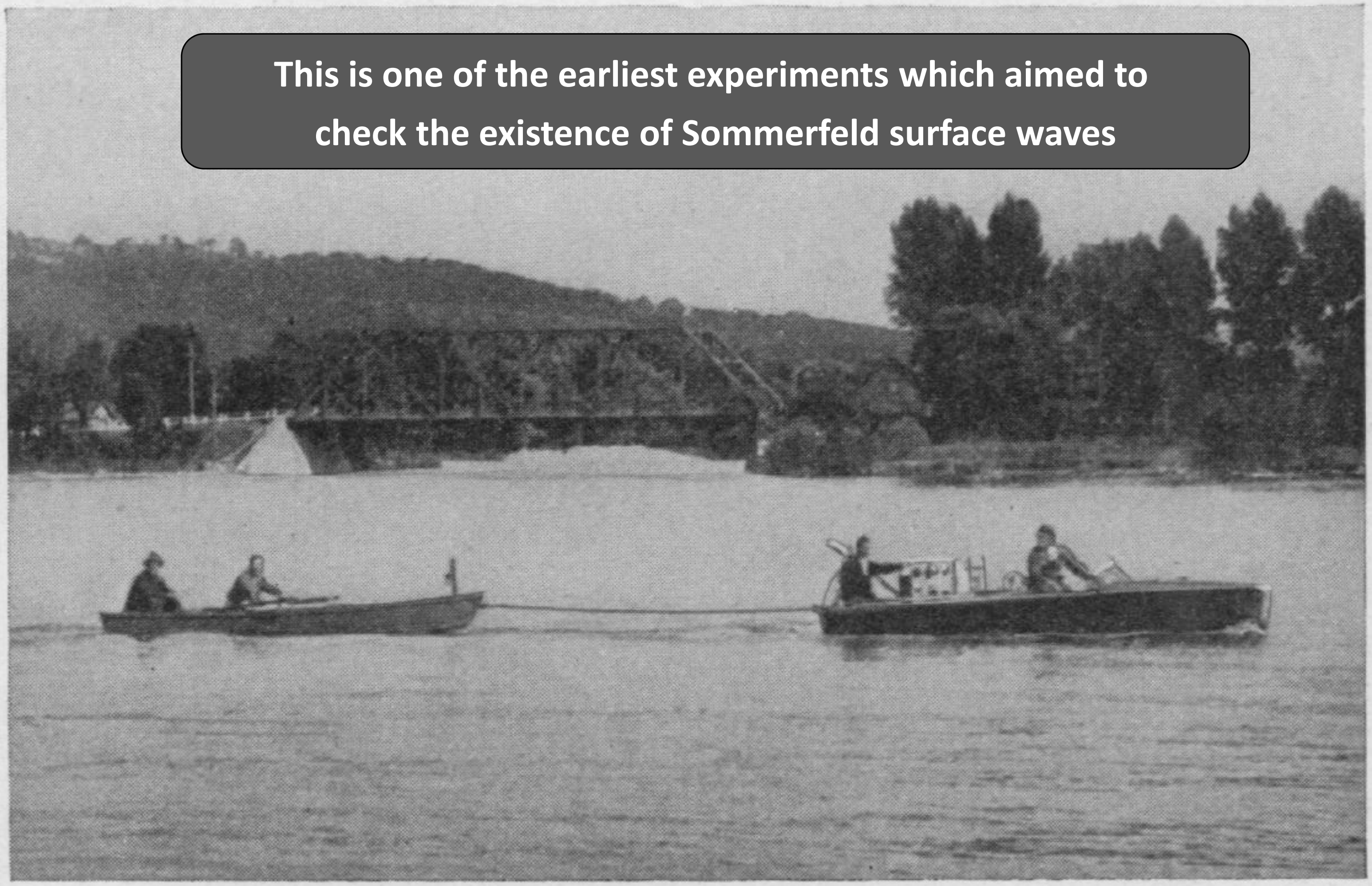
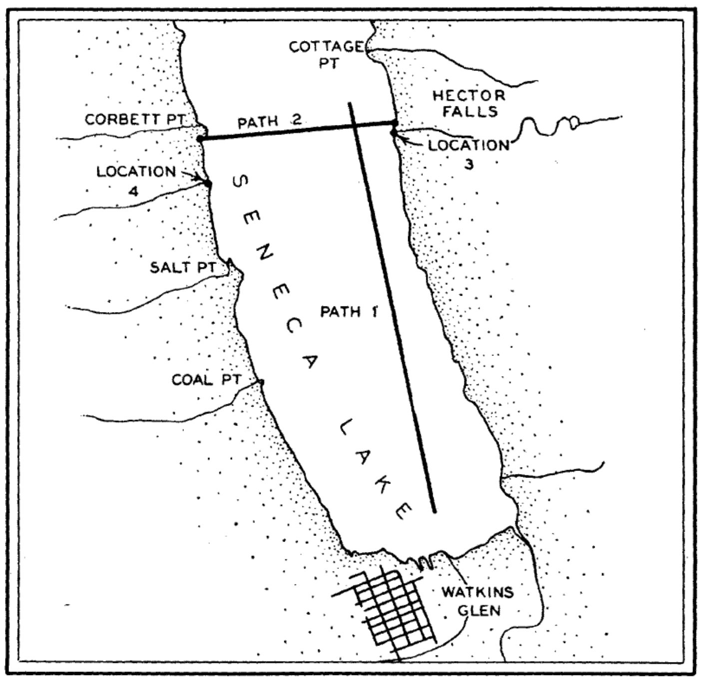


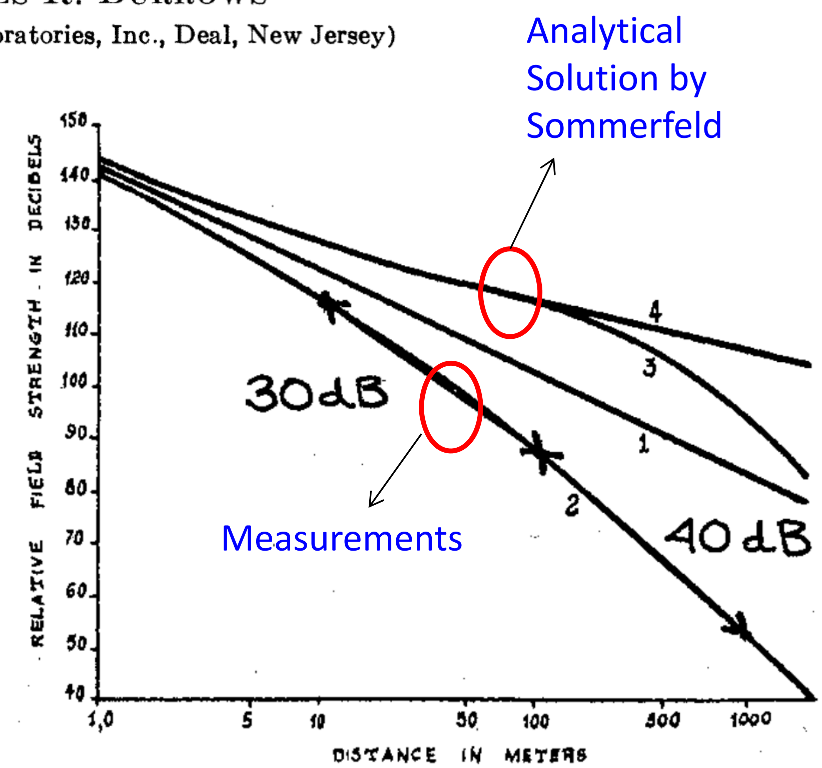
Fig. 1—Experimental arrangement for determining the variation of the received field strength with distance.

Proceedings of the Institute of Radio Engineers
Volume 25, Number 2 *February, 1937*

THE SURFACE WAVE IN RADIO PROPAGATION OVER PLANE EARTH*

BY
CHARLES R. BURROWS
(Bell Telephone Laboratories, Inc., Deal, New Jersey)

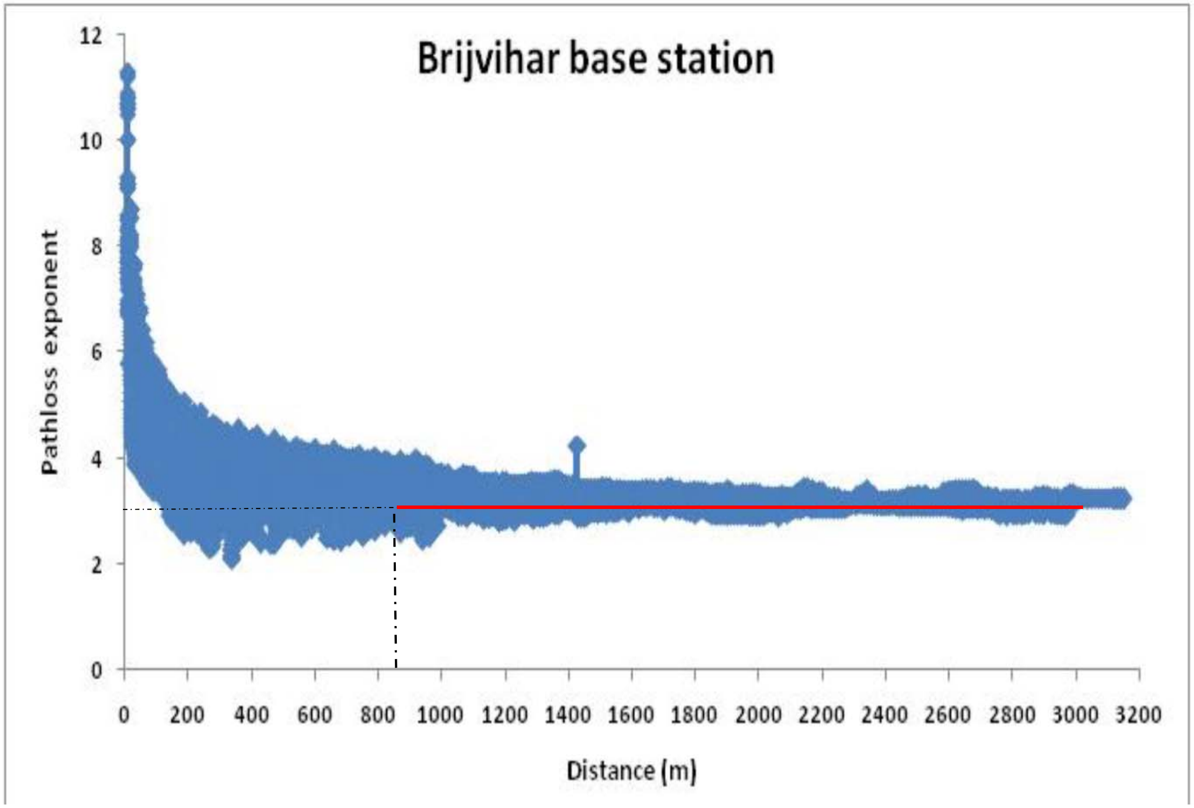




Experimental Data

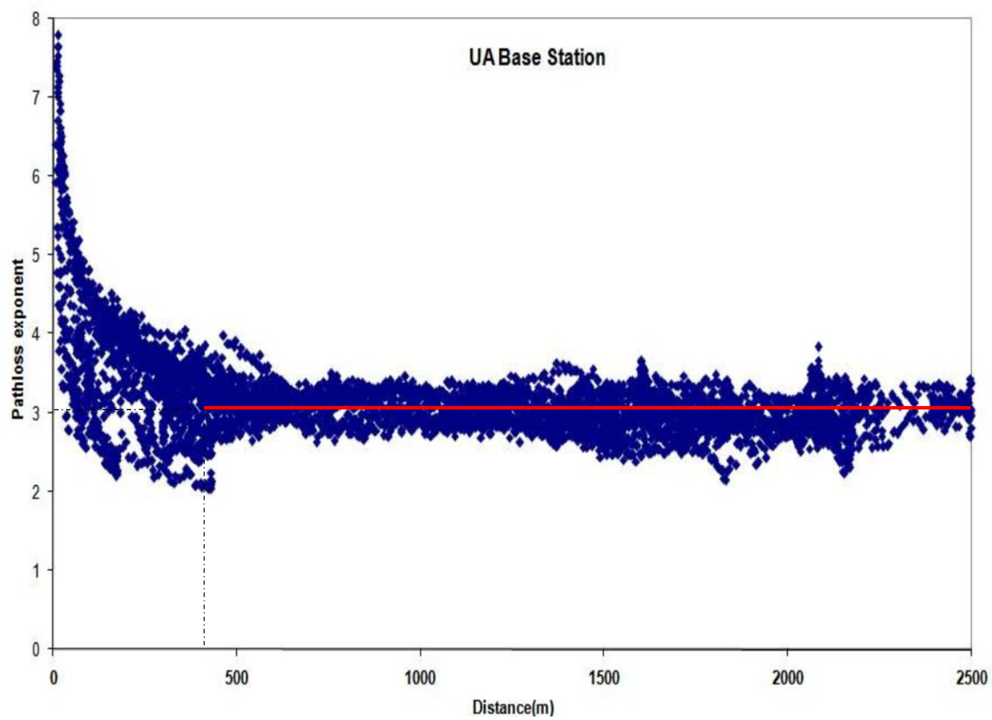
Photograph of a Delhi typical urban environment in this study.

Experimental Data

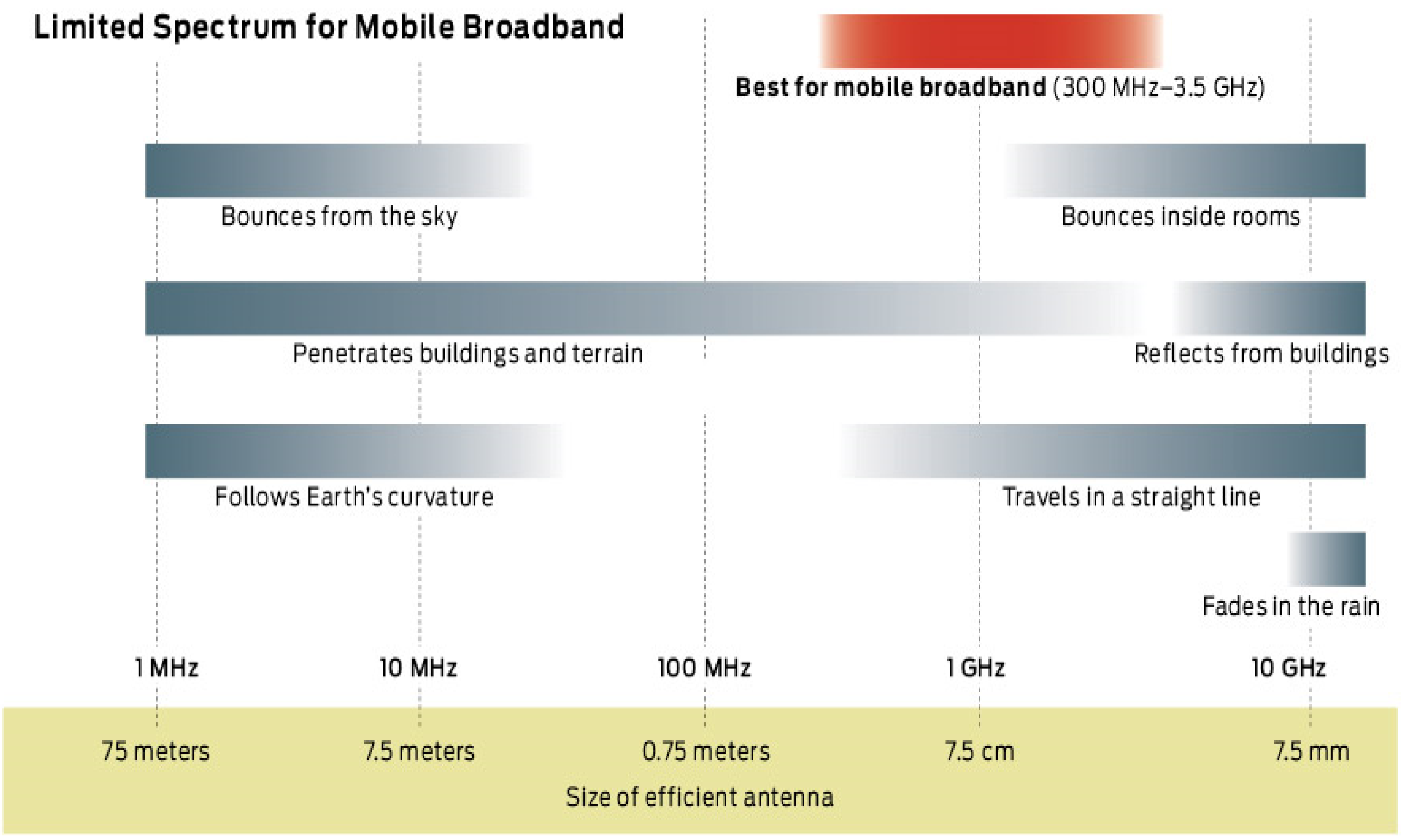


Variation of path loss exponent with distance for BJV base station (1800 MHz). Base station height: 24 m. Beginning of smooth region: 864 m.

Experimental Data

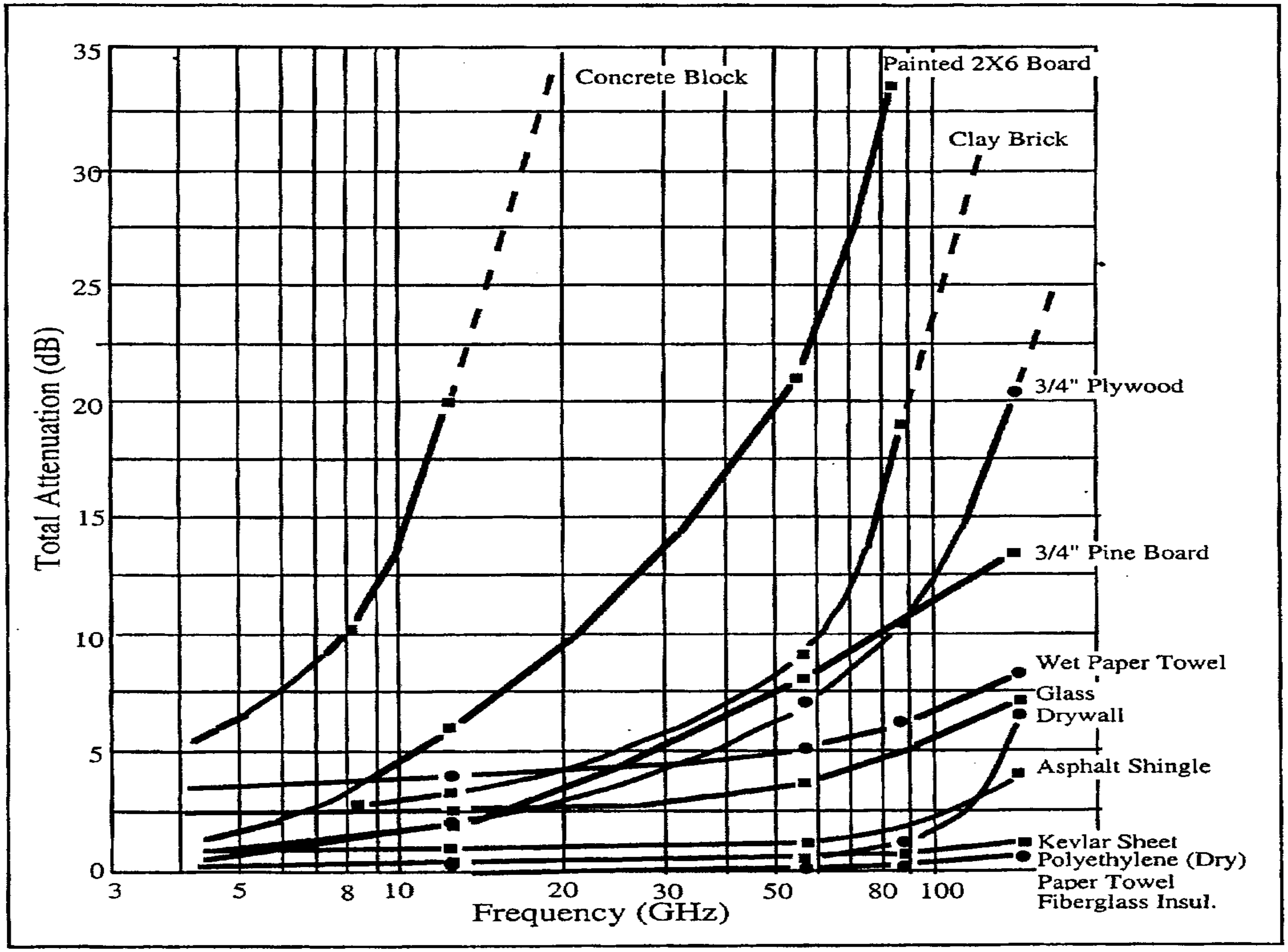


Variation of path loss exponent with distance for UA base station (900 MHz). Base station height: 24 m. Beginning of smooth region: 432 m.



OPPORTUNITY WINDOW: The best frequencies for mobile broadband are high enough that the antenna can be made conveniently compact, yet not so high that signals will fail to penetrate buildings. This leaves a relatively narrow range of frequencies available for use [red band].

IEEE SPECTRUM Magazine, October 2010, pp. 29



Concrete Block
Painted 2X6 Board
Clay Brick
3/4" Plywood
3/4" Pine Board
Wet Paper Towel
Glass
Drywall
Asphalt Shingle
Kevlar Sheet
Polyethylene (Dry)
Paper Towel
Fiberglass Insul.
Total Attenuation (dB)
0
5
10
15
20
25
30
35
3
5
8
10
20
30
50
80
100
Frequency (GHz)

SUMMARY FOR ALL THE EXPERIMENTAL DATA

- Path loss is 30 dB per decade **IRRESPECIVE** of the nature of the ground
- Path loss is **INDEPENDENT of FREQUENCY**
- Wireless Signals do penetrate well through buildings and the like and they do not generate the first order effects for the loss

An Observation from all the experimental data

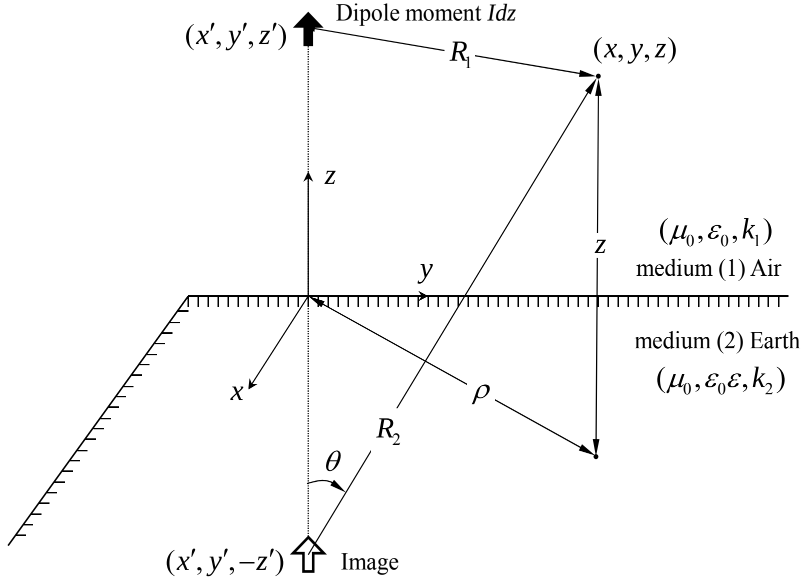
Electric field $\propto \rho^{-1.5}$; $Power \propto \rho^{-3}$ ► $10\log(\rho^{-3})$ ► $-30\frac{dB}{decade}$ $of\ distance$

At a distance of 1 km the path loss is	$10\log_{10}((1000)^{3})$ =-90 dB
At a distance of 2 km the path loss is	$10\log_{10}((2000)^{3})$ =-99 dB
At a distance of 3 km the path loss is	$10\log_{10}((3000)^{3})$ =-104 dB

The buildings, trees, and the like provide a 40-50 dB loss as shown by the variation of the measured data but compared to the free space Earth loss of 90 dB they are negligle! Hence a macro model based on physics can simulate the propagation path loss very accurately without going into expensive measurements

Theory – Sommerfeld Formulation

- Dipole of moment *Idz* oriented along the *z*-direction and located at (x', y', z') over an imperfect ground plane of complex relative dielectric constant ε.



$$k_1^2 = \omega^2 \mu_0 \varepsilon_0$$

$$\varepsilon = \varepsilon_r - \frac{j\sigma}{\omega \varepsilon_0}$$

$$k_2^2 = \omega^2 \mu_0 \varepsilon_0 \varepsilon$$

Theory – Sommerfeld Formulation

- Solution:

$$\Pi_{1z} = P\left[\left[\frac{\exp(-jk_1R_1)}{R_1} + \int_0^{\infty}\frac{J_0(\lambda\rho)}{\sqrt{\lambda^2-k_1{}^2}}\frac{\varepsilon\sqrt{\lambda^2-k_1{}^2}-\sqrt{\lambda^2-k_2{}^2}}{\varepsilon\sqrt{\lambda^2-k_1{}^2}+\sqrt{\lambda^2-k_2{}^2}}\exp\left(-\sqrt{\lambda^2-k_1{}^2}\,(z+z')\right)\lambda\,d\lambda\right]\right]$$

$$\Pi_{2z} = 2P\int_0^{\infty}\frac{J_0(\lambda\rho)\exp\left(\sqrt{\lambda^2-k_2{}^2}\,z-\sqrt{\lambda^2-k_1{}^2}\,z'\right)}{\varepsilon\sqrt{\lambda^2-k_1{}^2}+\sqrt{\lambda^2-k_2{}^2}}\lambda\,d\lambda$$

for $Real\left[\sqrt{\lambda^2-k_{1,2}^2}\right] > 0$, where

$$P = \frac{I\,dz}{j\omega 4\pi\varepsilon_0} \qquad \rho = \sqrt{(x-x')^2+(y-y')^2} \qquad R_1 = \sqrt{\rho^2+(z-z')^2}$$

Theory – Field Near the Interface

- For $|\varepsilon| \to \infty$ $\theta \approx \pi/2$, and ***W* small**,

$$G_{sV} \approx -\sqrt{\frac{2\pi k_1 j}{R_2}} \exp[-jk_1R_2] \frac{(z+z')}{R_2} \frac{\varepsilon}{\sqrt{\varepsilon^2 - 1}} \approx -\sqrt{2\pi k_1 j} \frac{(z+z')\exp[-jk_1R_2]}{R_2^{1.5}}$$

and three facts can be recognized:

- A variation of the Hertz potential as $\Pi_{1z} \propto \frac{1}{R_2^{1.5}}$ which will give a field variation of $\frac{1}{\rho^{1.5}}$ if $z+z' \ll \rho$, or equivalently, a path loss exponent of 3.
- A height gain effect: see the $(z+z')$ term.
- **No dependence with the ground parameters.**

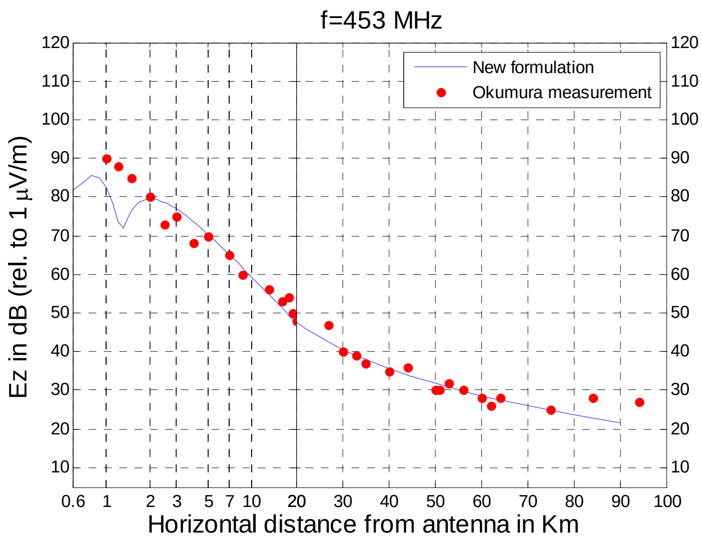
Theory – Field Near the Interface

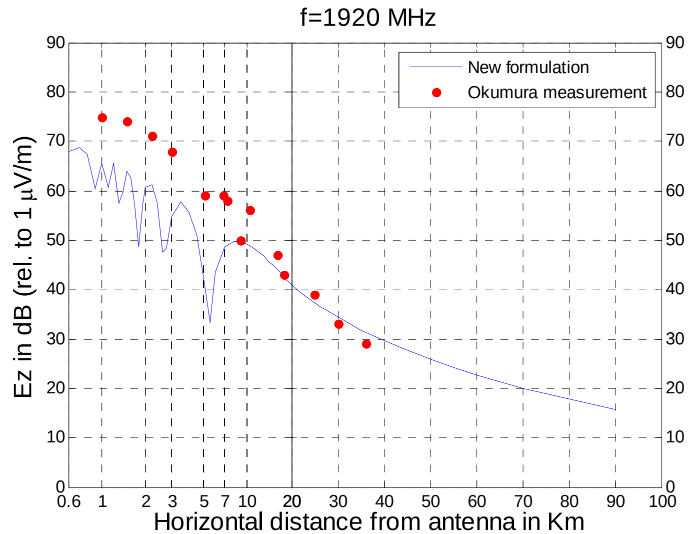
- For $|\varepsilon| \to \infty$, $\theta \approx \pi/2$, and ***W* large**,

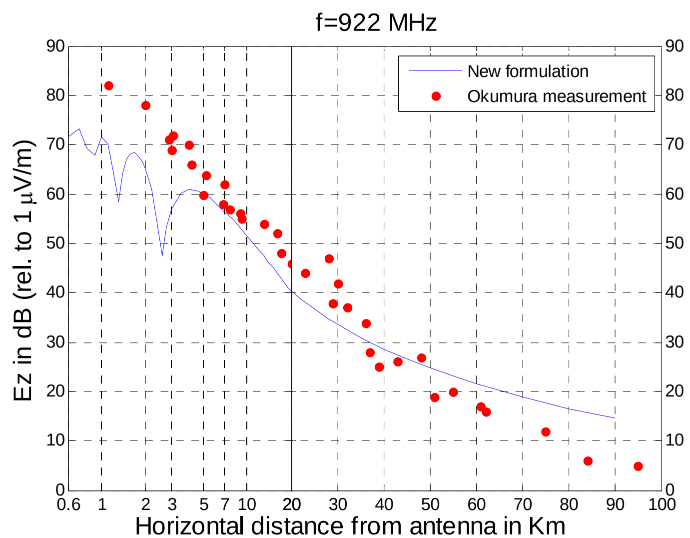
$$G_{sV} \approx 2\sqrt{\varepsilon}\,\exp[-jk_1 R_2]\frac{(z+z')}{R_2^2}\left[1-\frac{\varepsilon}{j\,k_1 R_2}\right]$$

and four facts can be recognized:

  - A term with a variation of the Hertz potential as $1/R^2$, which is recognized as a **Norton surface wave**, and will give a path loss exponent of 4.
  - A higher order term with a variation as $1/R^3$.
  - A height gain effect for both terms.
  - **A dependence with the ground parameters for both terms**.

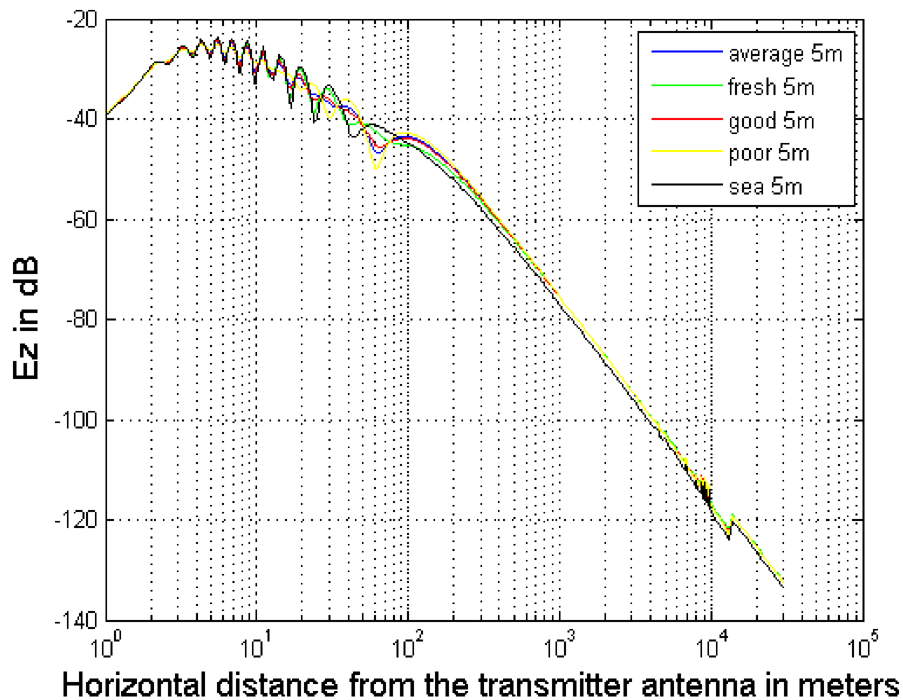






Comparison between Okumura's drive test measurements and the numerical analysis done using Schelkunoff Integrals

Numerical Analysis – Field Near an Earth-Air Interface

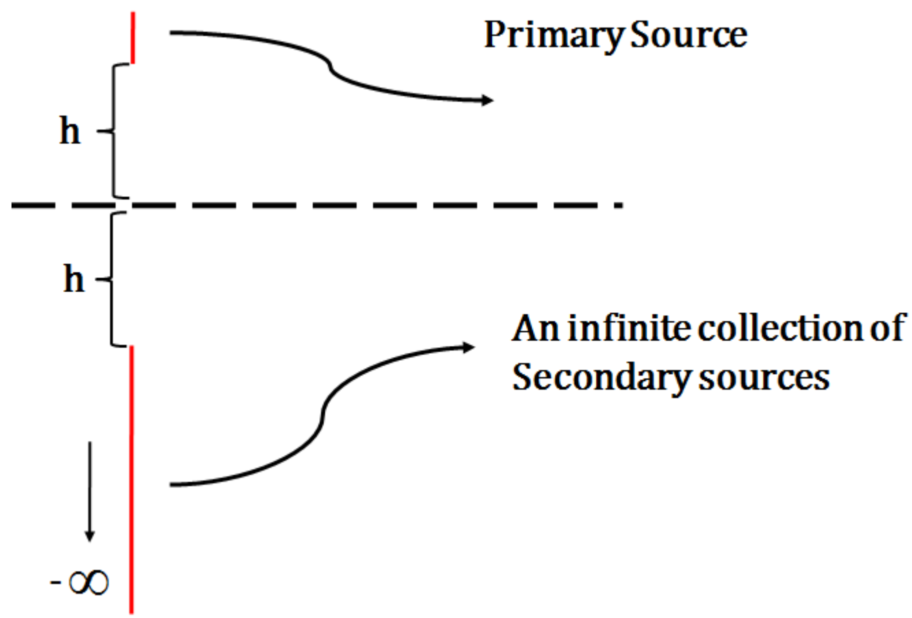


Variation of magnitude of E_z from a half-wavelength dipole as a function of distance, at an operating frequency of 900 MHz. The height of the observation point was 2 m. Five different types of ground have been used, with different parameters.

What Type of Wave Is It?

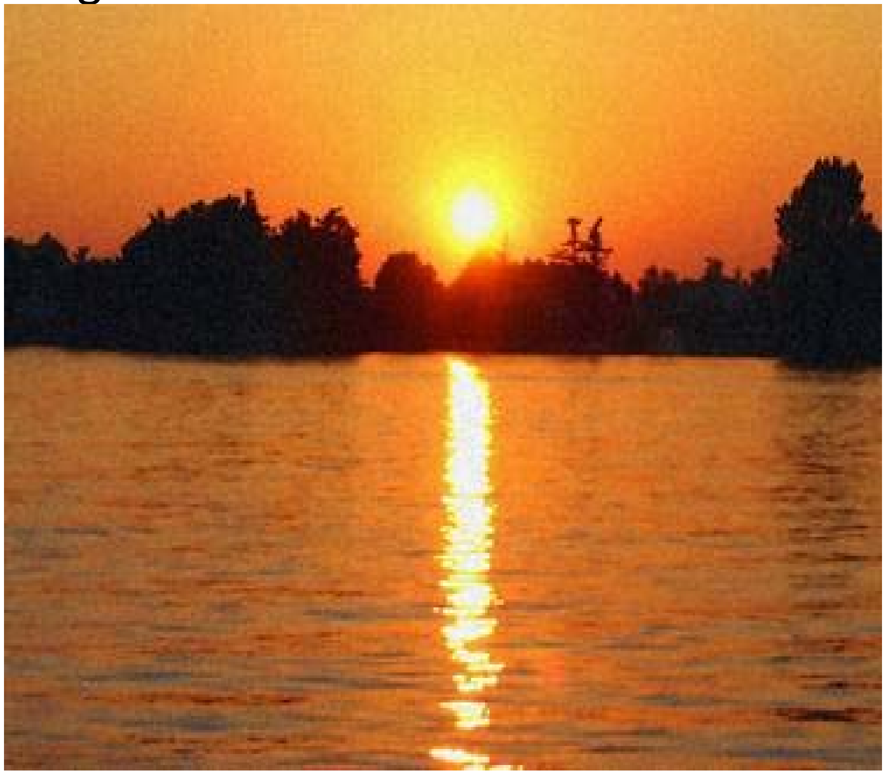
(Point Source:1/(r*r), Line source (1/r), Planar Source ©

- Following Van der Pol and other researchers, it can be concluded that such wave is a surface wave originated by a 2D infinite source as the secondary sources shown in the following diagram, which is precisely the situation in this problem.



What Type of Wave Is It?

- An optical analog situation:



Range extension due to height gain


Surface wave continuous propagation after a big obstruction


10

Ultrawideband: Working Methodology

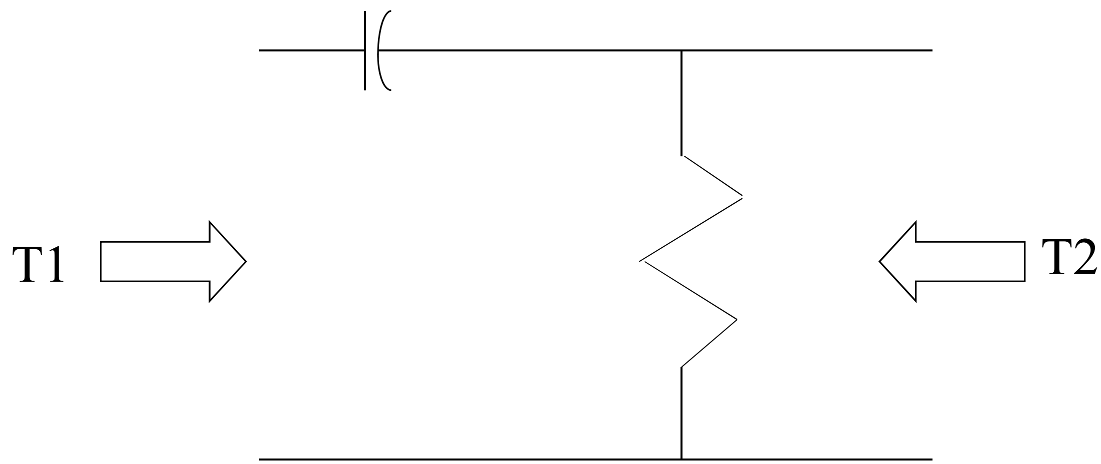
Old Paradigm: The transmit antenna pattern = Receive Antenna pattern

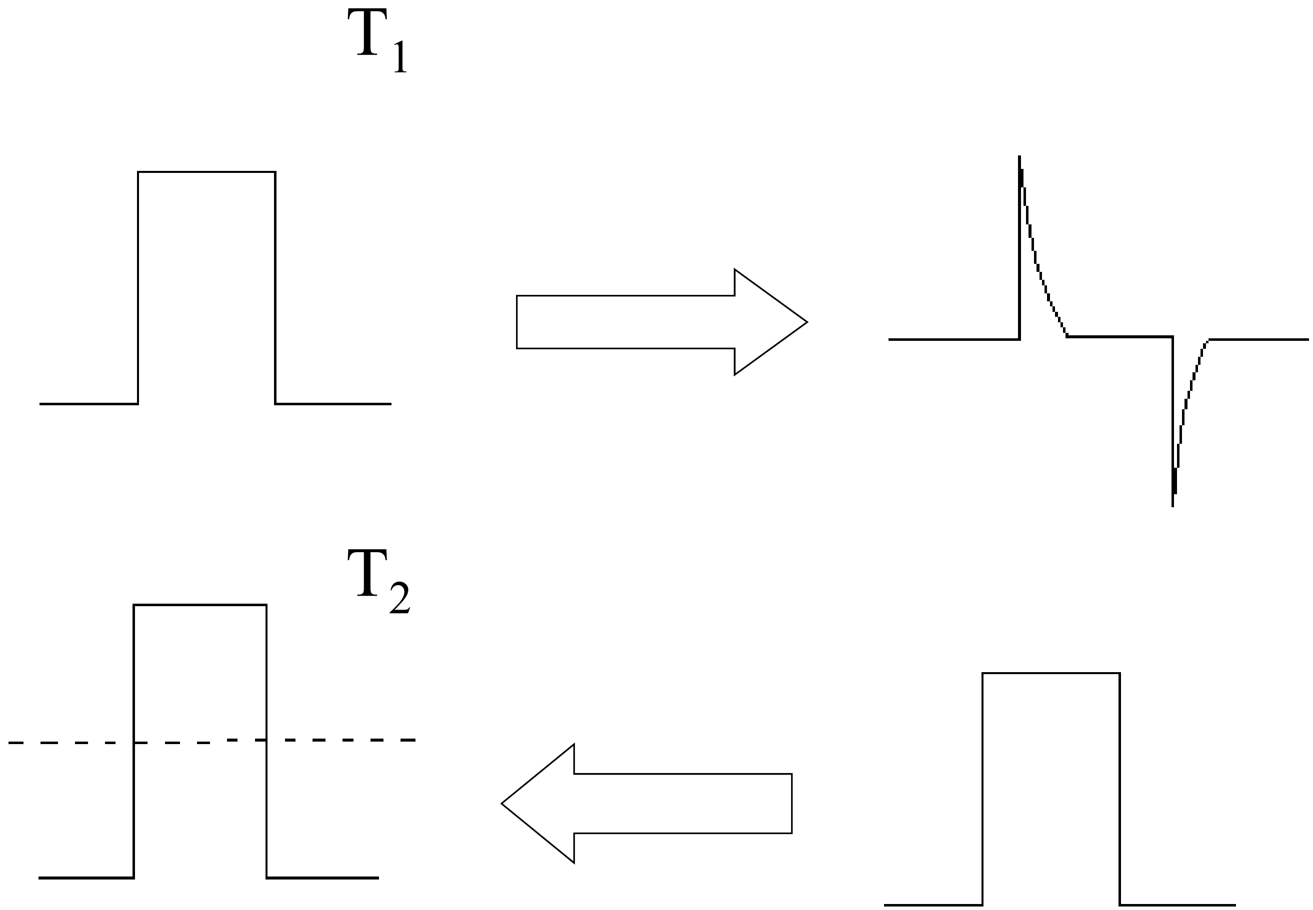
New Paradigm: For Broadband applications

In time domain the transmit impulse response of an antenna **IS NOT EQUAL TO** the impulse response of the antenna on receive

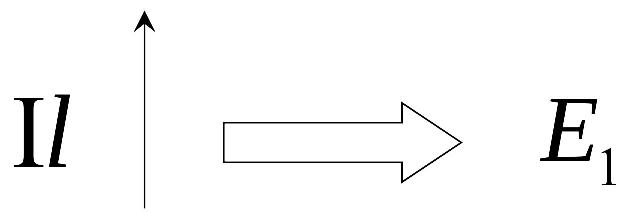
RECIPROCITY IN TIME DOMAIN IS NOT A SIMPLE PRODUCT IT INVOLVES A CONVOLUTION OVER TIME AND **HENCE WAVESHAPES ARE NOT RECIPROCAL**

Reciprocity does not apply to waveforms in the way we were use to understand it!!!



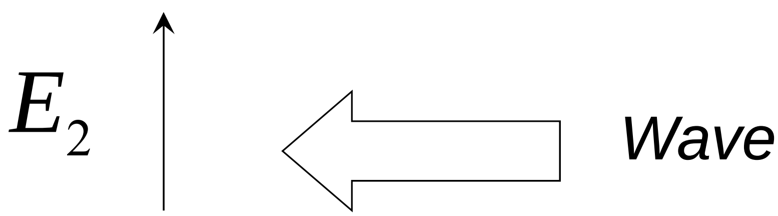

T_1
T_2

Transmit



$$E_1 = -\frac{j\omega\mu}{4\pi} \mathrm{I}l \frac{e^{-jkR}}{R}$$

Receive



$$\int E_2 I dl = \mathrm{Id}l \frac{e^{-jkR}}{4\pi R}$$

Far Field in Frequency Domain of a Point Radiator

$$\mathbf{A}(x,y,z) = \frac{\mu}{4\pi}\frac{e^{-jkR}}{R}\mathbf{J}_i$$

$$\mathbf{E}_{far}(x,y,z) = -j\,\omega\mathbf{A} = -j\,\frac{\omega\mu}{4\pi}\,\mathbf{J}_i\,\frac{e^{-jkR}}{R}$$

Far Field in Time Domain of a Point Radiator

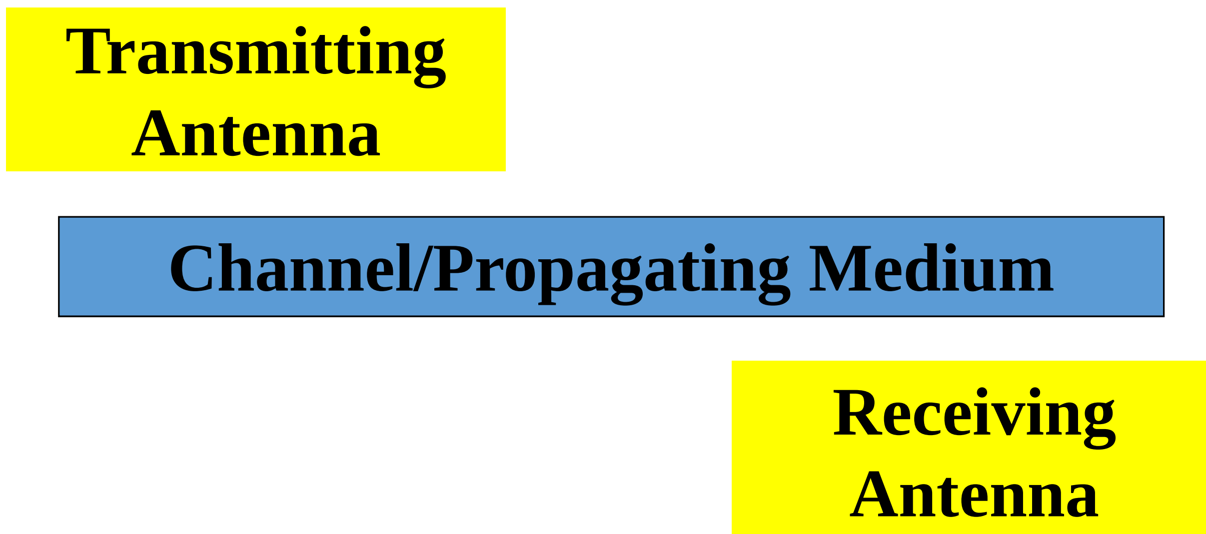
$$\boldsymbol{J}_i\,\delta(0,0,0,t) = \hat{z}\,\delta(0,0,0)\,f(t)$$

$$\boldsymbol{A}(\mathbf{r},t) = \frac{\mu}{4\pi}\frac{\hat{z}\,f(t-|R|/c)}{R}$$

$$\boldsymbol{E}(\mathrm{r},t) = -\frac{\partial\boldsymbol{A}(\mathrm{r},t)}{\partial t} = -\frac{\mu\,\hat{z}}{4\pi R}\frac{\partial f(t-|R|/c)}{\partial t}$$

A Point Radiator therefore **DIFFERENTIATES** the input waveshape

Model of a Wireless/RADAR System for BROADBAND Applications



Transmitting and Receiving Antenna systems must be part of the equalizers. The Impulse response of the antennas are a function of the azimuth and elevation angles

FRIIS Equation for PATH LOSS

$$P_{RX} = P_{TX} \; G_{RX} \; G_{TX} \; \frac{\lambda^2}{(4\pi R)^2}$$

The Gain of the Transmit and the Receive antenna increase with frequency and hence inversely with the wavelength: so what is the net effect?

PATH LOSS INDEPENDENT OF FREQUENCY!!

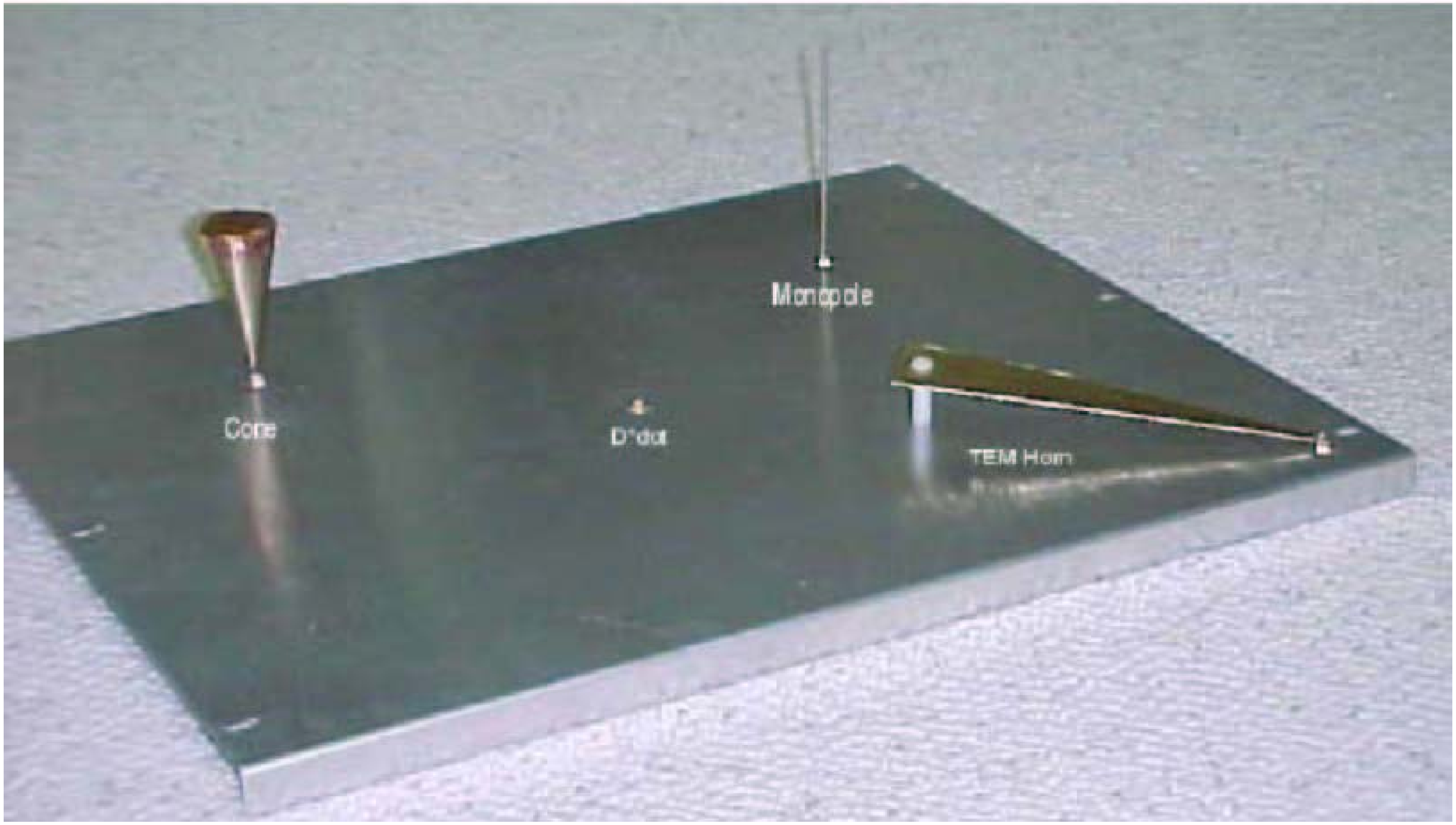
FRIIS Equation for PATH LOSS

$$P_r = P_t G_r G_t \left(\frac{\lambda}{4\pi R} \right)^2 = \frac{P_t G_t A_R}{4\pi R^2}$$

In this form if the transmitting antenna gain and the receiving aperture area are independent of frequency wideband operation is possible: so what is the net effect?

PATH LOSS INDEPENDENT OF FREQUENCY!!

Reference: Jim Andrews, *UWB Signal Sources, Antennas and Propagation, Picosecond Pulse Lab, Boulder, Colorado*



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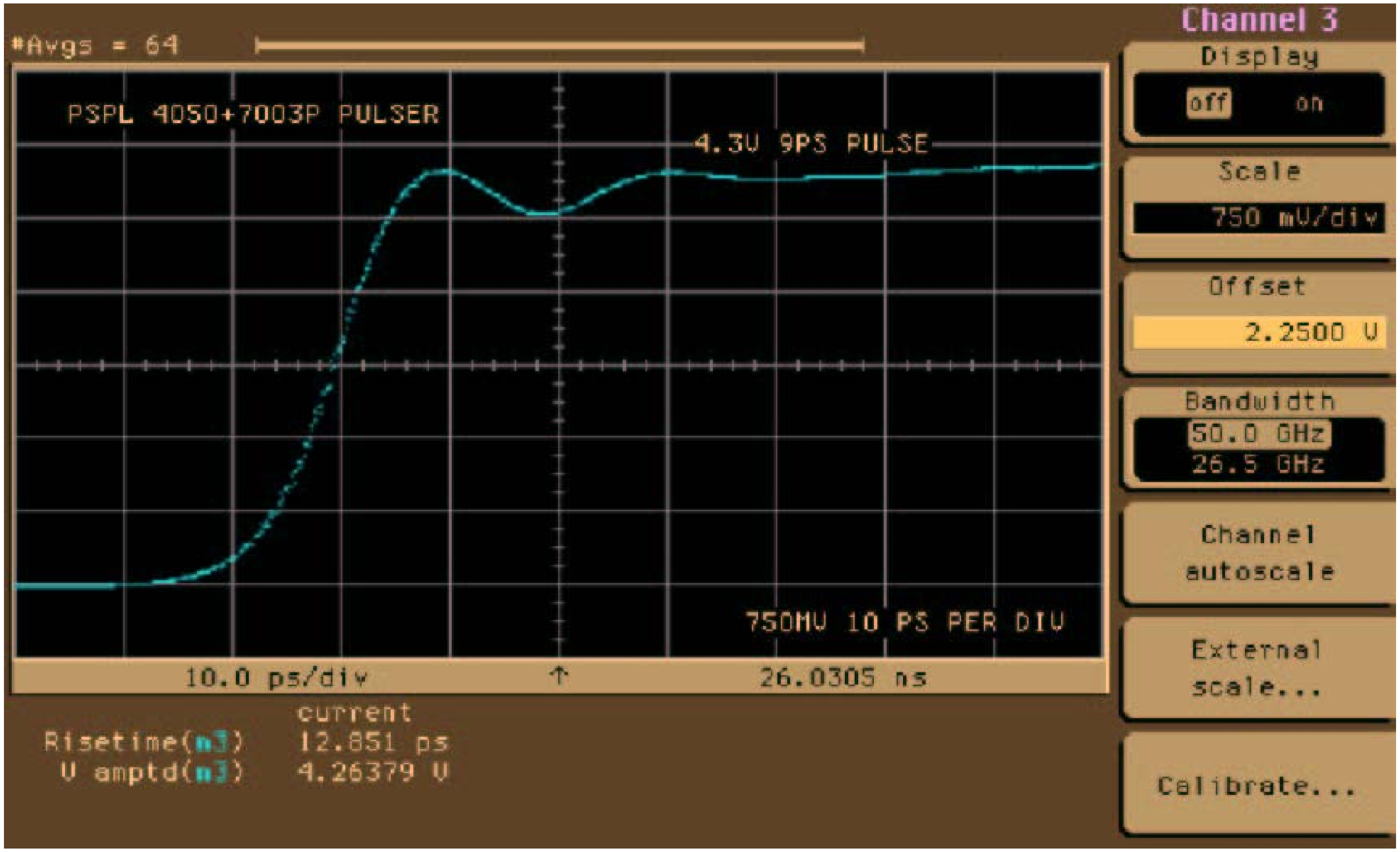
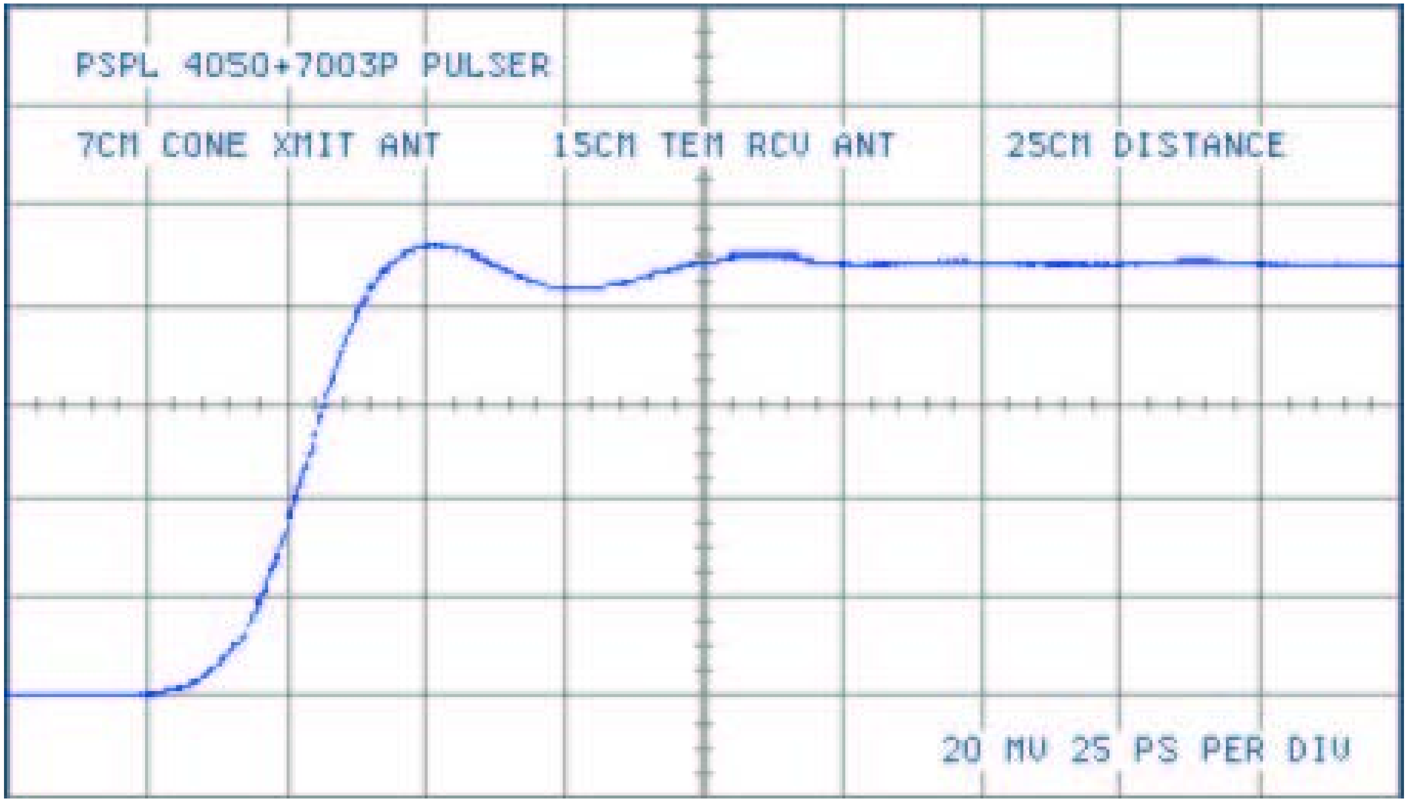


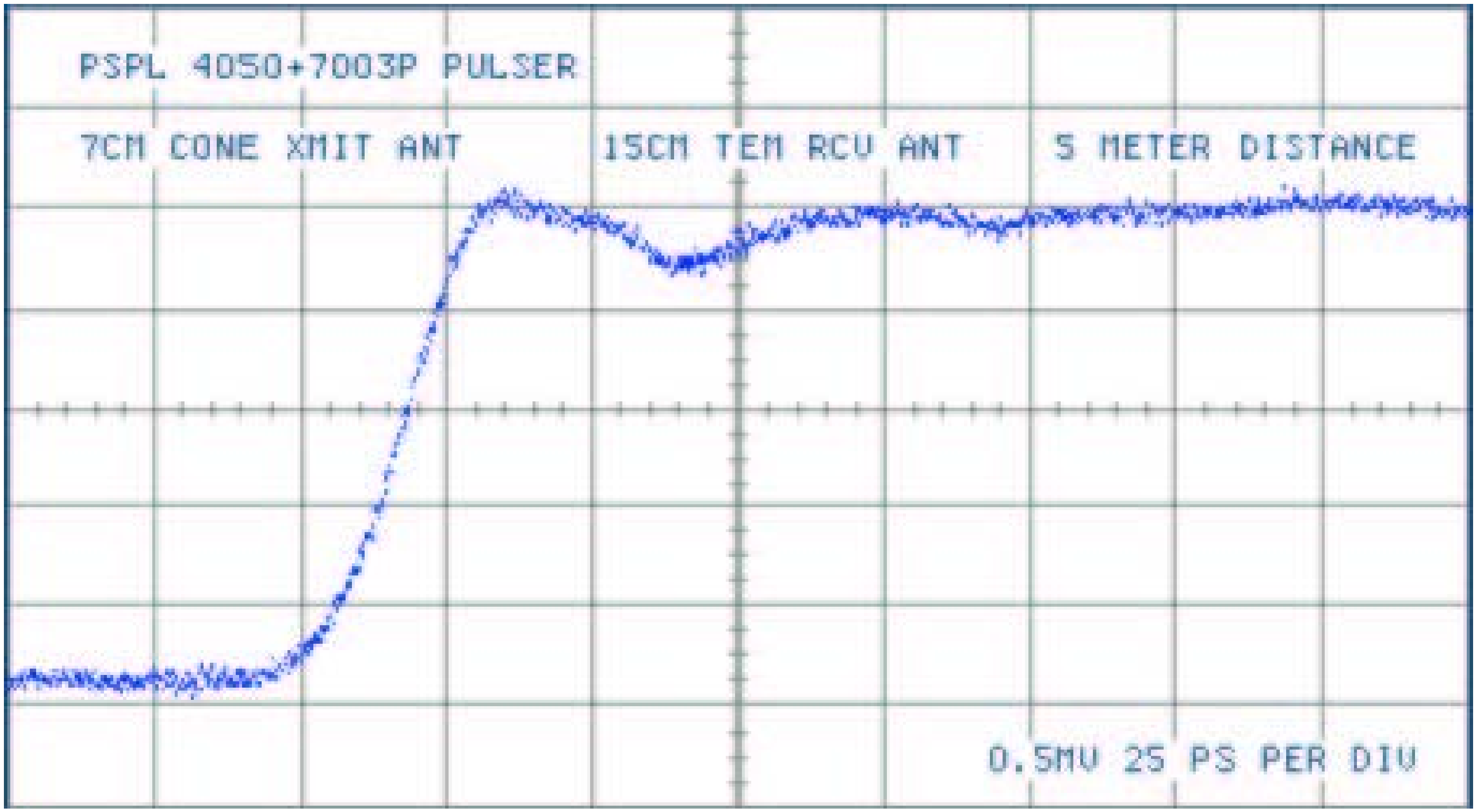
4V 9ps step pulse used for testing the theory. Measured by a HP 50 GHz 9 ps risetime sampling scope. (Transmitter = Bicone; Receiver = TEM Horn)

Reference: Jim Andrews, *UWB Signal Sources, Antennas and Propagation, Picosecond Pulse Lab, Boulder, Colorado*



Bicone radiates the same waveform and TEM receives the same waveform without distortion

Reference: Jim Andrews, *UWB Signal Sources, Antennas and Propagation, Picosecond Pulse Lab, Boulder, Colorado*



PATH LOSS IS INDEPENDENT OF FREQUENCY!!

A Note on Channel Capacity

Shannon (white noise limited)
Gabor (interference limited)
Tuller (system design –
for near field applications)

THE ZERO ERROR CAPACITY OF A NOISY CHANNEL

Claude E. Shannon
Bell Telephone Laboratories, Murray Hill, New Jersey
Massachusetts Institute of Technology, Cambridge, Mass.

The ordinary capacity C of a noisy channel may be thought of as follows. There exists a sequence of codes for the channel of increasing block length such that the input rate of transmission approaches C and the probability of error in decoding at the receiving point approaches zero. Furthermore, this is not true for any value higher than C. In some situations it may be of interest to consider, rather than codes with probability of error <u>approaching</u> zero, codes for which the probability <u>is</u> zero and to investigate the highest possible rate of transmission (or the least upper bound of these rates) for such codes. This rate, C_o, is the main object of investigation of the present paper.

2. A *transmitter* which operates on the message in some way to produce a signal suitable for transmission over the channel. In telephony this operation consists merely of changing sound pressure into a proportional electrical current. In telegraphy we have an encoding operation which produces a sequence of dots, dashes and spaces on the channel corresponding to the message. In a multiplex PCM system the different speech functions must be sampled, compressed, quantized and encoded, and finally interleaved

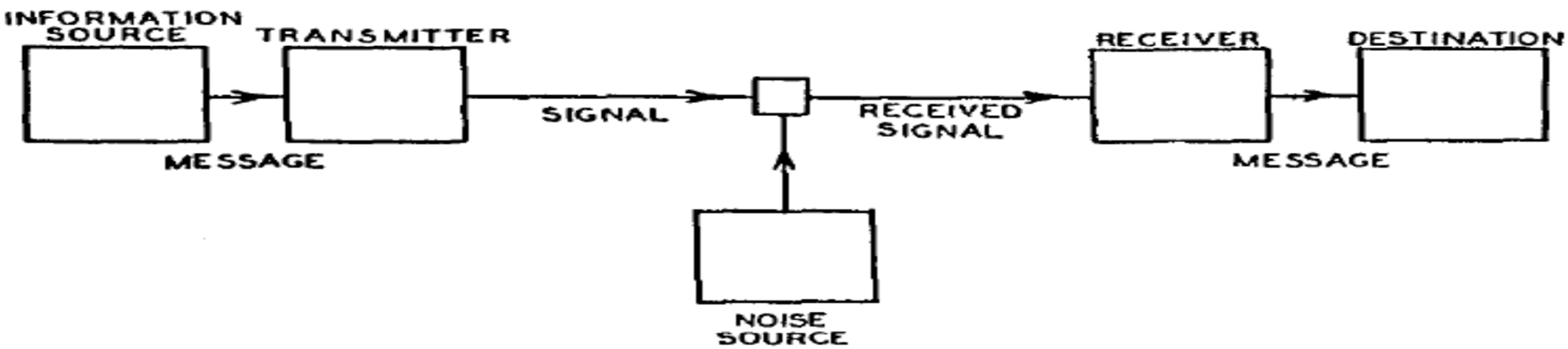


Fig. 1—Schematic diagram of a general communication system.

properly to construct the signal. Vocoder systems, television, and frequency modulation are other examples of complex operations applied to the message to obtain the signal.

3. The *channel* is merely the medium used to transmit the signal from transmitter to receiver. It may be a pair of wires, a coaxial cable, a band of radio frequencies, a beam of light, etc.

4. The *receiver* ordinarily performs the inverse operation of that done by the transmitter, reconstructing the message from the signal.

In a noisy system, the problem is mathematically considerably more difficult. Nevertheless, a definite channel capacity C exists in the following sense. It is possible by proper encoding of binary digits into allowable signal functions to transmit as closely as desired to the rate C binary digits per second with arbitrarily small frequency of errors. There is no method of encoding which transmits a larger number. In general, the ideal rate C can only be approached by using more and more complex encoding systems and longer and longer delays at both transmitter and receiver.

Since the output of any information source can be encoded into binary digits using, statistically, R binary digits per symbol, the problem of defining a channel capacity can be reduced to the problem of determining the maximum number of binary digits that can be transmitted per second over the channel.

Shannon's Contribution according to Viterbi

makes sense to apply the Shannon channel-coding and capacity theorems to design an appropriate code that will introduce the necessary redundancy at the transmitter for GPS systems to combat the noisy characteristics of the channel, so that the received signal provides reliable transmission of information. The launch of the first artificial satellite generated a need for conserving transmitted power over a real channel that is approximated by this model. Shannon's theorems thus served as a powerful stimulus to further theoretical work, particularly toward the goal of minimum-complexity decoding. By the late 1960s, it was feasible to code this

additive white Gaussian noise (AWGN) channel with an error probability of 10^{-5}, so that the channel could conserve 5-6 dB of power relative to the uncoded operation at data rates in megabits per second. To gauge the economic advantage, consider that an alternative method of gaining 5 dB of power requires more than triple the area of the receiving antenna [31]. When the signal pow-

The Bandwagon

CLAUDE E. SHANNON

INFORMATION theory has, in the last few years, become something of a scientific bandwagon. Starting as a technical tool for the communication engineer, it has received an extraordinary amount of publicity in the popular as well as the scientific press. In part, this has been due to connections with such fashionable fields as computing machines, cybernetics, and automation; and in part, to the novelty of its subject matter. As a consequence, it has perhaps been ballooned to an importance beyond its actual accomplishments. Our fellow scientists in many different fields, attracted by the fanfare and by the new avenues opened to scientific analysis, are using these ideas in their own problems. Applications are being made to biology, psychology, linguistics, fundamental physics, economics, the theory of organization, and many others. In short, information theory is currently partaking of a somewhat heady draught of general popularity.

Toward a Circuit Theory of Communication

Michel T. Ivrlač and Josef A. Nossek, *Fellow, IEEE*

Abstract—Electromagnetic field theory provides the *physics* of radio communications, while information theory approaches the problem from a purely *mathematical* point of view. While there is a law of conservation of energy in physics, there is no such law in information theory. Consequently, when, in information theory, reference is made (as it frequently is) to terms like energy, power, noise, or antennas, it is by no means guaranteed that their use is consistent with the physics of the communication system. Circuit theoretic multiport concepts can help in bridging the gap between the physics of electromagnetic fields and the mathematical world of information theory, so that important terms like energy or antenna are indeed used consistently through all layers of abstrac-

Dennis Gabor wrote:[1972, IEEE Trans on Information Theory, First Issue]

The wireless communication systems are due to the generation, reception and transmission of electro-magnetic signals. Therefore all wireless systems are subject to the general laws of radiation.

Communication theory has up to now been developed mainly along mathematical lines, taking for granted the physical significance of the quantities which are fundamental in its formalism.

But communication is the transmission of physical effects from one system to another. Hence communication theory should be considered as a branch of physics.

Thus it is necessary to embody in its foundation such physical data.

Gabor contd…..

WE OBSERVE FIRST THAT ALL ELECTRIC SIGNALS ARE CONVEYED BY RADIATION. EVEN IF LINES OR CABLES ARE USED IN THE TRANSMISSION, BY THE MAXWELL-POYNTING THEORY THE ENERGY CAN BE LOCATED IN EMPTY SPACE. HENCE WE CAN APPLY TO OUR PROBLEM THE WELL KNOWN RESULTS OF THE THEORY OF RADIATION.

Communication Theory:

Provides an approximate analysis & has nothing to do with the realizability of the system

Electromagnetic Theory:

Provide system design and provides the basis of a practical system

Q&A