LF/MF Antennas for Amateurs

Rudy Severns

N6LF

Preface

Vertical transmitting antennas have been used since the beginning of radio. With well over 100 years of experience, literally thousands of articles and hundreds of books you would think there would be very little left to discover. However, when I began researching the literature I found continuous repetition of the same basic information but not so much on fine details, especially over the last 40 years or so. Useful information current at earlier times has often been dropped or forgotten. In a few cases I felt some of the accepted "common knowledge", much of it derived from broadcast applications, was not correct or at least not appropriate for amateur applications. In particular HF ground system design and the significant differences between HF and LF-MF soil electrical characteristics has not been adequately aired. I admit to a passion for fine details bordering on obsession but in the case of vertical antennas at least some details can be of practical help.

Sixty plus years ago Tom Erdmann, W7DND (SK), gave me some advice on priorities for my first station: if I had a $100 I should spend $90 on the antenna, $9 on the receiver and $1 for the transmitter. Prices have gone up a bit in the past 65 years but I still keep Tom's priorities firmly in mind. I've always invested far more time, money and effort in my antennas than the rest of the station. Antennas are a lot fun and in retirement LF, MF and HF verticals have become an obsession. For the past 25 years I've been particularly interested in 160m operation, building a number of vertical and sloper arrays. For the last 10 years I've been part of the ARRL 600m experimental group (WD2XSH) transmitting at 465-510 kHz. At the 2012 WRC amateurs succeeded in obtaining worldwide allocations on 2200m and 630m and the FCC has now given U.S. amateurs access to these bands.

For many years amateurs have had only one MF band, 1.80-2.0 MHz (160m) but now 135.7-137.8 kHz (2200m) and 472-479 kHz (630m) has been added. Except for a few experimental licenses amateurs haven't been allowed on these frequencies for over 100 years. This lack of experience means there is a need for practical information on many subjects related to LF/MF operation including antennas. There are many "old" but potentially useful ideas which deserve renewed consideration.

Because of the vast literature on vertical antennas I've made no attempt to make this book a compendium, it is just "some notes", nothing more. I've focused on subjects of interest to me and have direct experience with. Some of the material in this book is original but most is drawn from the work of others which I acknowledge whenever I can identify the source. The references have been placed at the end of each chapter.

It's my personal philosophy that one needs both theory and experiment to understand a phenomenon. In particular theory and experiment must give the same answer, if not then you don't understand what's going on and need to keep working to resolve the differences! I have spent many hours wondering "what the h... is going on?"

Because I'm speaking to a audience with a very wide range of experience, from the non-technical newcomer to the graduate engineer, the arrangement of this book has to be a little different. I've divided the material into two levels. The six chapters in the body of the book are limited to basic information which, combined with a little modeling or simple calculations, can be applied directly using graphs for special problems like inductor design. The math has been minimized. For those more mathematically proficient and wanting more detailed justification for the advice given in the main text, an extensive body of technical information has been placed in a series of appendices which are available on-line (tbd).

While this book is far from perfect or complete I hope it's useful.

GL and 73, Rudy Severns N6LF, WD2XSH/20 April 2017

Acknowledgements

In preparing these notes I've had a great deal of help from other hams reviewing my scribbling which has been crucial, resulting in rewriting of the entire text several times. in short they've kept me reasonably honest! I would like to thank Fritz Raab, W1FR, Pat Hamel, W5THT, and Jim Miller, AB3CV, for their time and efforts which really helped. In the process these notes were placed on my web page and a number of amateurs made helpful comments. Many of the pictures in chapter 6 were supplied by others who are identified in the captions.

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Chapter 1

An Overview

1.0 Introduction

100 years ago amateurs were restricted to wavelengths below 200m (f >1.5 MHz). This has recently changed and we now have allocations at 2200m (135.7-137.8 kHz, LF) and 630m (472-479 kHz, MF). However, amateurs have very little experience at these frequencies and design and construction of antennas for the new bands is substantially different from HF. The intent of these notes is to provide practical advice on LFMF transmitting antennas. There is a perception that substantial acreage is required for the antennas on these bands. That is not the case! Those with small properties can be successful but we have to know how!

There are differences between LFMF and HF which impact antenna design:

1) Wavelengths are much longer so that any practical antenna will be electrically small.

2) Soil electrical characteristics change substantially going from HF down to LFMF.

3) Transmitting power limitations are in terms of power radiated from the antenna rather than maximum transmitter output power.

1.1 Long wavelengths

At 1.9 MHz the wavelength (λ) ≈518' so a $\lambda/4$ vertical will be ≈130' high. If you divide 1.9 MHz by four you get 475 kHz, right in the middle of the new 630m band. A $\lambda/4$ on160m will be only ≈$\lambda/16$ at 475 kHz. 2200m is another factor of 3.5 lower in frequency so a $\lambda/4$ vertical on 160m is only ≈0.018λ on 137 kHz. At 475 kHz λ≈2071' so a $\lambda/4$ vertical would be ≈500' high. At 137 kHz $\lambda/4$ ≈1800'! In any case, the FCC has limited the maximum height to 197' (60m), which is only 0.095λ at 137 kHz.

The focus of this book is on antennas with heights (H) practical for amateurs, i.e. H=20'→100' (H≈0.01→0.05λ at 475 kHz and H≈0.003→0.015λ at 137 kHz). In terms of electrical height these are certainly "short" antennas, with very low radiation resistance (Rr), narrow matched SWR bandwidth and low efficiency. A major part of the design effort for LFMF antennas is directed towards obtaining adequate efficiency.

1.2 Soil characteristics

Because soil electrical characteristics have a profound affect on transmitting antenna performance, basic information on soil electrical characteristics is useful.

σ = soil conductivity in Siemens/meter [S/m], Siemen=Mho

$\boldsymbol{\varepsilon_o}$ = permittivity of a vacuum = 8.854 X 10^{-12} [Farads/m]

$\boldsymbol{\varepsilon_r}$ = relative permittivity or relative dielectric constant

$\boldsymbol{\varepsilon=\varepsilon_o\varepsilon_r}$= effective permittivity or dielectric constant [Farads/m]

$\boldsymbol{\mu_o}$ = permeability of free space = 4π 10^{-7} H/m

ω=2πf

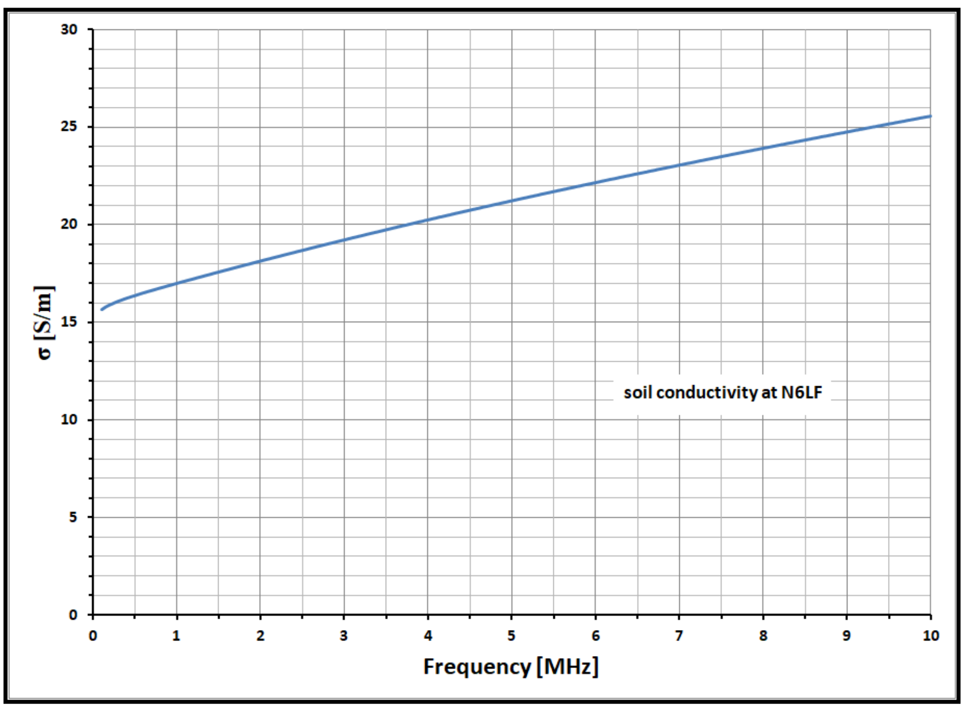

Figure 1.1 - Typical soil conductivity variation.

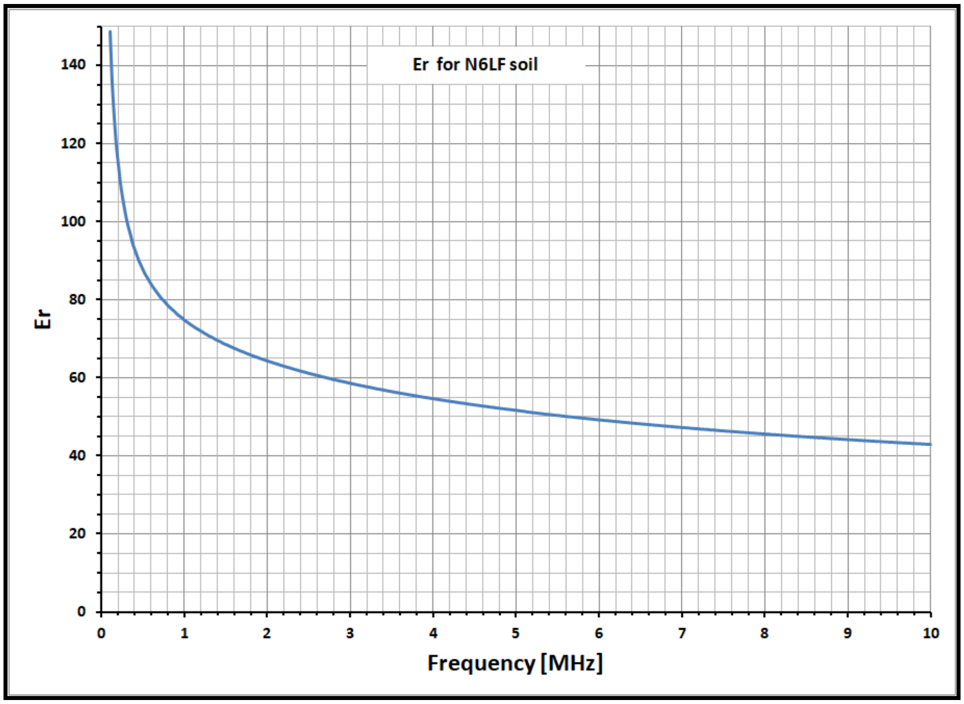

Figure 1.2 - Typical soil permittivity variation.

Figures 1.1 and 1.2 illustrate how soil electrical characteristics vary with frequency at a typical QTH. In this example at 100 kHz $\sigma \approx 0.15$ S/m and that value increases with frequency. εr behaves just the opposite, decreasing with frequency.

1.3 EIRP and radiated power

On the new bands power limits are stated in terms of "effective isotropic radiated power" or "EIRP". The "isotropic" in EIRP refers to an idealized antenna in free space which radiates power uniformly in all directions, i.e. if you measure the power density (S, in W/m^2) on the surface of a virtual sphere surrounding an isotropic radiator you'll find the power density is the same everywhere.

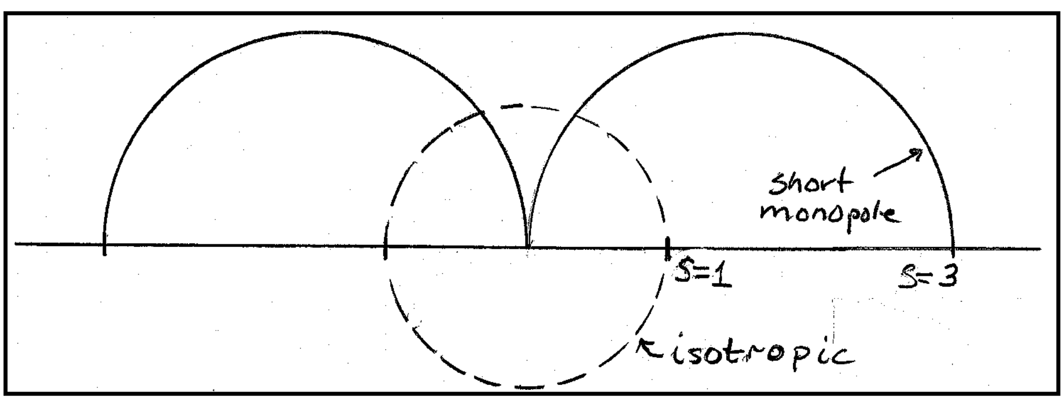

Figure 1.3 - Power density: isotropic radiator versus a short monopole.

Figure 1.3 compares the radiation patterns of an isotropic radiator in free space to a short vertical over ideal ground. The directivity of the isotropic radiator is 1 (0 dBi). When a short monopole is placed over a perfect ground-plane, for the same total radiated power (Pr) the power density, at the same distance horizontally from the base, will be greater by a factor of 3 (+4.77 dB). This increase comes from two sources, Pr is being radiated into a hemisphere rather than a sphere because of reflection from the ideal ground which doubles S and there is a further increase of 1.5X (+1.77 dB) due to the directivity of a short monopole. There is a direct relationship between the power density at a given distance and the magnitude of the electric field intensity (|E|) at that point:

$$|\mathbf{E}| = \sqrt{\mathbf{377S}} \quad (1.2)$$

Because of it's directivity we must reduce the Pr of the short monopole by a factor of three to maintain the same power density as an isotropic. On 630m 5W EIRP is allowed and on 2200m the allowed EIRP is 1W, which means Pr is about 1.7W on 630m and 0.33W on 2200m. The key word is "radiated" power.

At HF, antenna efficiencies are typically >90% and the focus is on gain. On LFMF our goal is to achieve sufficient efficiency that we can radiated the allowed power with the available transmitter power. This is a fundamentally different mindset! We have the choice of a large efficient antenna with small input power (Pi) or a small inefficient antenna with a large input power. Most installations will be a balance between the two extremes. Running very high power is an option in theory but, as shown in chapter 6, section 6.10 and chapter 2, section 2.10, very high power into a small antenna results in very high voltages (tens of kV!) and currents (kA). The high power approach is self limiting! A transmitter output power of 100W is generally pretty easy to obtain and 100W is frequently assumed in later chapters unless stated otherwise. In addition to the EIRP power limit, the FCC has also limited the input power to the antenna to 500W pep on 630m and 1.5 kW pep on 2200m, however, given limitations due to the high voltages associated with these power levels, from a practical point of view these limits are moot.

How can we determine the radiated power (Pr) for a particular antenna? The pros do it by measuring the electric field intensity at a given distance from which Pr can be calculated. For most amateurs that's not very practical. If the value for the antenna's radiation resistance (Rr) is known we can calculate Pr in a couple of ways. Given Rr and a measurement of Io, the current at the base of the antenna, then:

$$\mathbf{Pr = Io^2 Rr} \qquad (1.3)$$

As an alternative we can measure the input power (Pi) and the resistive component of the feedpoint impedance (Ri) to give:

$$\mathbf{Pr = Pi\left(\frac{Rr}{Ri}\right)} \qquad (1.4)$$

Where do we get Rr from? As will be shown in chapters 2 and 3, Rr can be found using either modeling or manual calculations. Using the value for Rr from a model over perfect ground is in general not valid at HF where the dielectric properties of soil have a direct influence on Rr. However, at frequencies below ≈1 MHz the soil electrical characteristics are dominated by conductivity and Rr can be approximated by the perfect ground value.

1.4 Some fundamental advice

A very succinct summary of LFMF antenna design was given by Woodrow Smith[2] 75 years ago:

> *"The main object in the design of low frequency transmitting antenna systems can be summarized briefly by saying that the general idea is to get as much wire as possible as high in the air as possible and to use excellent insulation and an extensive ground system."*

This simple advice should be taken literally!

This advice can be organized in order of priority:

1) Make the vertical as tall as you can.

2) Use as much capacitive top-loading as practical (chapter 3).

3) Use carefully placed high Q loading coils (chapters 4 & 6).

4) Put substantial effort into the ground system, with the radial density high near the base of the vertical and under the top-loading hat (chapter 5).

5) Minimize conductor losses by using multiple wires and/or large diameter conductors (chapter 3).

6) Use high quality insulators, at the base and at wire ends.

1.5 Modeling and calculations

Antennas for these bands have to be customized for each installation to take advantage of available resources, space and supports. There are several ways to approach the design: use a combination of algebraic approximations and graphs or use antenna modeling CAD software or some combination of the two.

Much of the material in this book was derived using CAD modeling, EZNEC Pro4 v6[3] (with the NEC4.2 engine) and AutoEZ[4] an EXCEL spreadsheet which automates many modeling functions. These are very good tools but except for buried ground systems most design questions can be adequately addressed with NEC2 based software like 4NEC2[5] which is an excellent free program.

Computer modeling is not the only way. One of the consequences of the small electrical size of LFMF antennas is that the currents on the conductors tend have only small phase differences and relatively linear amplitude variation. As shown in chapters 2 and 3, it's possible to use simple algebraic expressions to estimate radiation resistance (Rr), effective capacitance of top-loading structures (Ct) and other quantities.

1.6 Loading inductors

A major part of the design effort for LFMF antennas is directed at obtaining adequate efficiency. Given practical height limitations, most LFMF antennas will require loading inductors for resonance and matching. In many cases the losses in this inductor may determine the efficiency of the antenna. Much of the design effort is directed towards first minimizing the required inductance (L) with height and top-loading (chapter 3) and then maximizing inductor "Q" (QL) (chapter 6).

1.7 Examples of early LF/MF antennas

Small antennas are not new. At the beginning of radio very long wavelengths were used so all antennas were "small" even those hundreds of feet high. A lot of effort was directed towards improving these antennas, work that continued into the 1960's for VLF applications[6]. The low efficiency and narrow bandwidth associated with small antennas arises from fundamental physics [7,8]. Like the perpetual motion machine, 100% efficiency in small antenna is not in the cards but adequate efficiencies are not hopeless. Interestingly, short antennas are still a hot topic today among professionals where the interest is in very small antennas[9] for wireless mobile devices, RFID, etc. Despite 120 years of work there's still a lot to learn! A rich source of ideas for LFMF antennas are old radio books. Often these books are seen at ham flea markets or used book stores for a few dollars. The 1920's and especially the 1930's were a time when most amateurs did not have a lot of money and improvisation was the order of the day. Much of that work is still useful today.

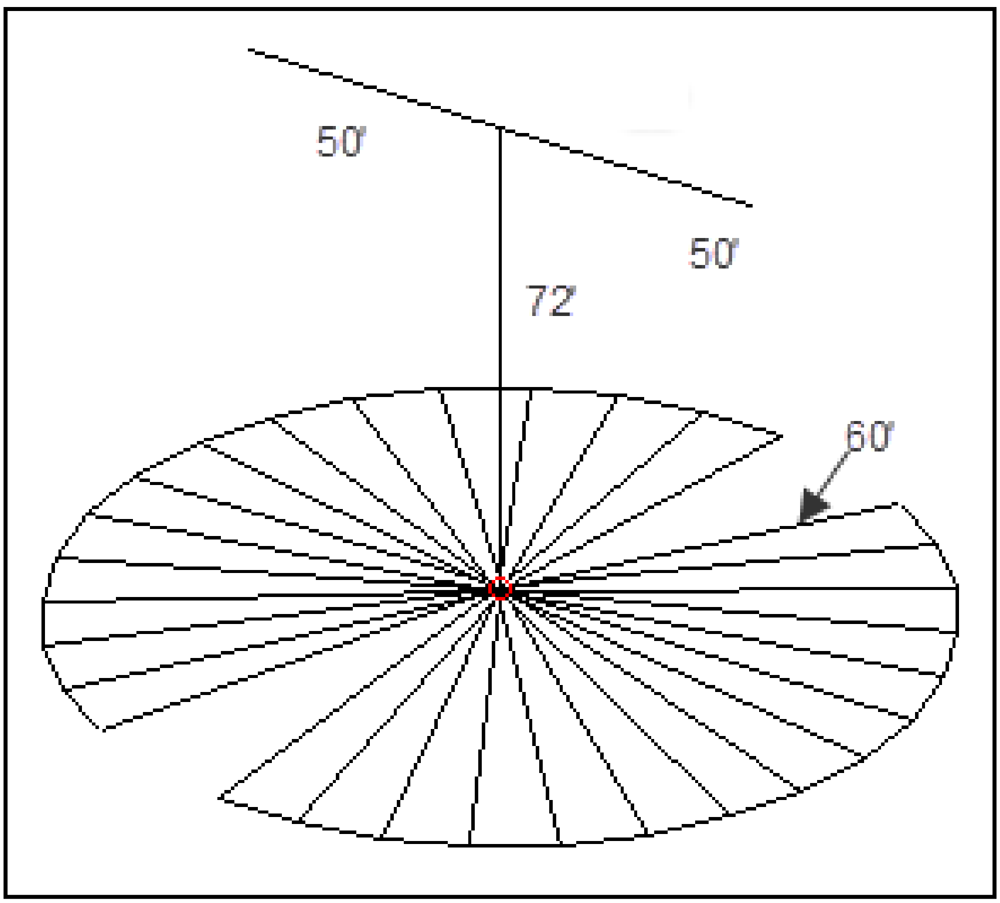

Figure 1.4 - EZNEC model of the 1BCG antenna.

Figure 1.4 is a sketch of the antenna used for the initial transatlantic tests by amateurs (1BCG) in 1921-22[10, 11]. The operating frequency was ≈1.3 MHz (λ≈230m). At 1.3 MHz λ/4 = 189' so the 60' radius of the counterpoise corresponds to ≈0.08λ. Figure 1.5 (taken from the Moyer & Wostrel[12]) shows a variety of possibilities, including inverted L's, T's, fans and umbrellas. Some of the simplest top-loaded antennas are the "inverted-L" and the "T" which can be just a single wire suspended between two supports with a wire (the "down lead" or "lead-in") down to the shack as shown in figure 1.6 or it can use a multi-wire top-hat and down-lead as shown in figure 1.7. Figure 1.7 also shows a very large elevated ground system or counterpoise. Very effective but few amateurs would build something on that scale!

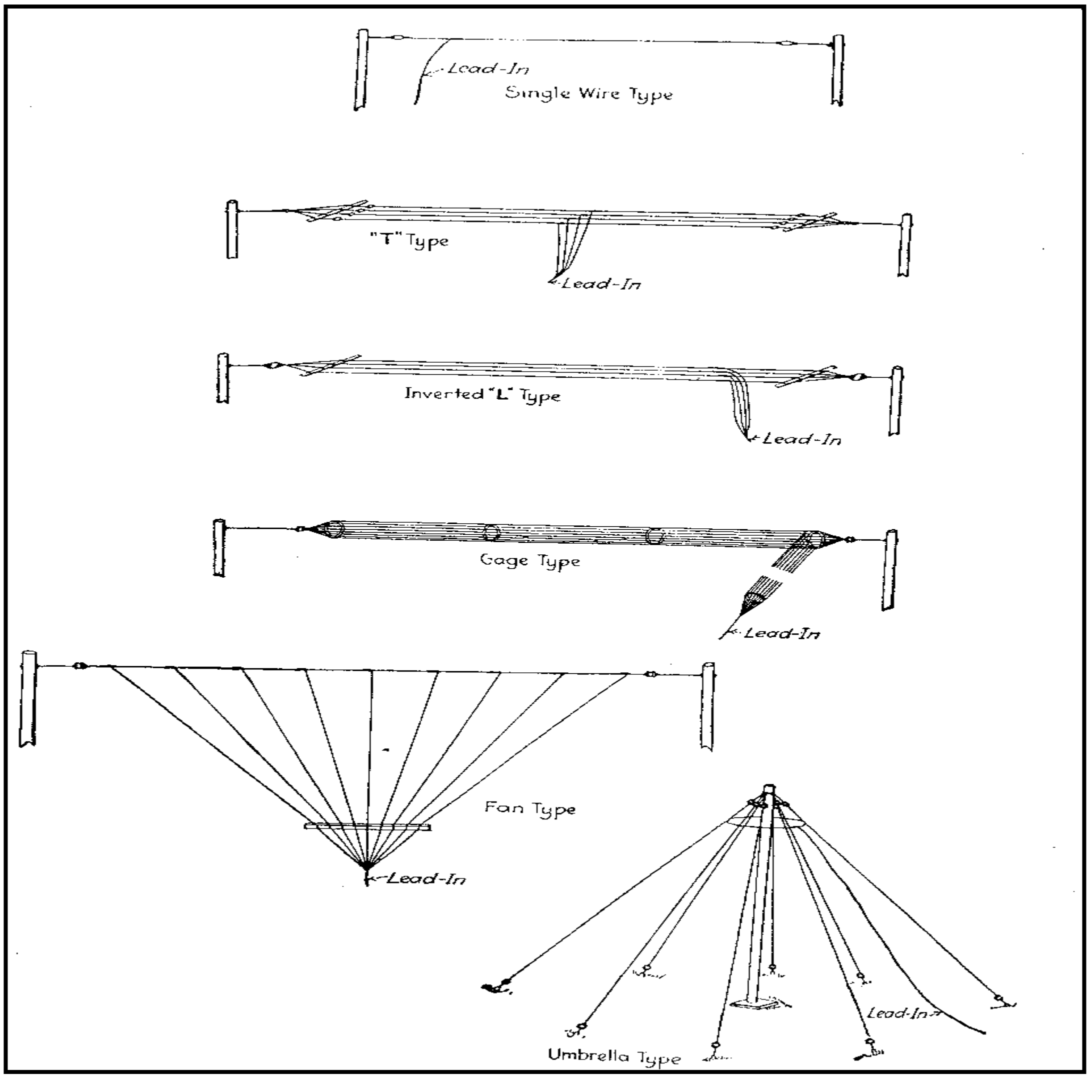

Figure 1.5 - Examples of early antennas [12].

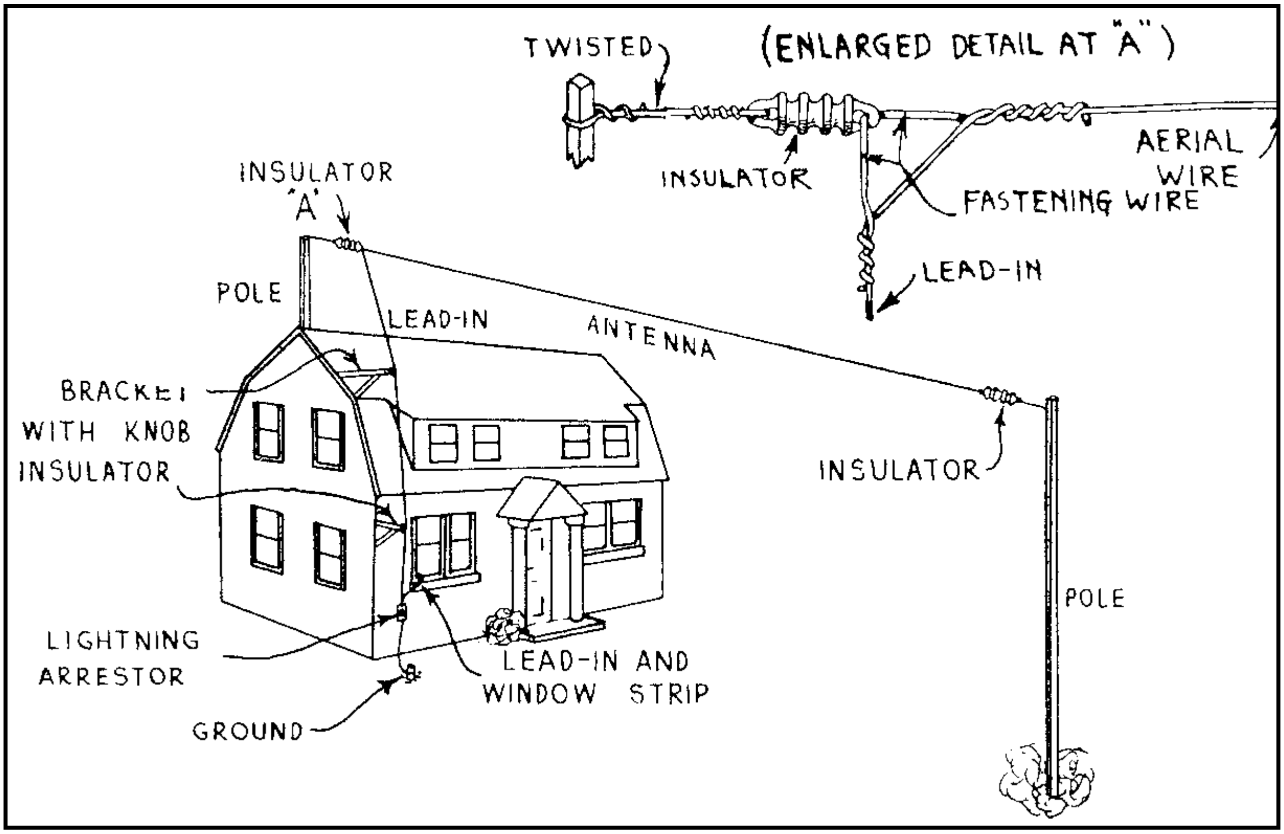

figure 1.6 - Example of an inverted L antenna. From Ghirardi[13].

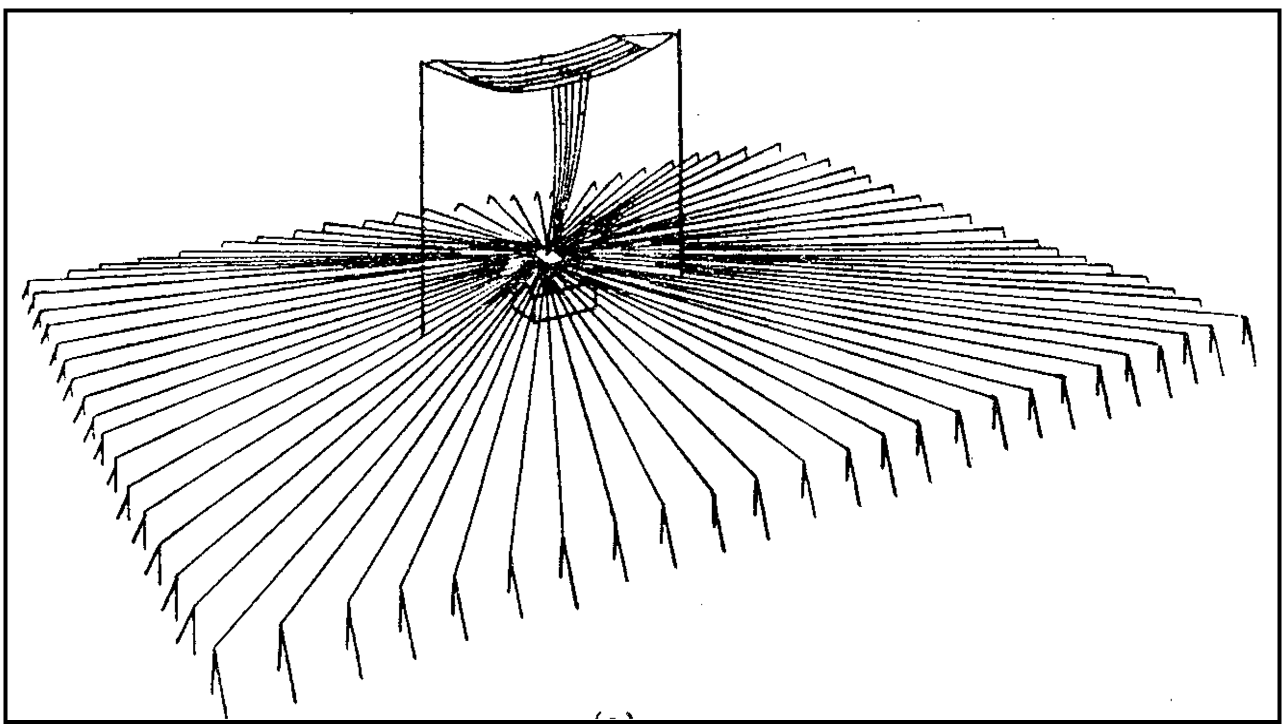

Figure 1.7 - A very large LF elevated ground system. From the Admiralty Handbook of Wireless Telegraphy, 1932 [14].

References

[1] King and Smith, Antennas in Matter, MIT press 1981, page 399, section 6.8

[2] Woodrow Smith, Antenna Manual, 1948

[3] Lewallen, Roy, W7EL, EZNEC Pro/4 v6, www.w7el.com

[4] Maguire, Dan, AC6LA, AutoEZ, http://ac6la.com/autoez.html

[5] 4NEC2, http://home.ict.nl/~arivoors/

[6] Watt, Arthur D., VLF Radio Engineering, Pergamon Press, 1967

[7] Wheeler, Harold A., Fundamental Limitations of Small Antennas, Proceedings of the I.R.E., vol. 35, December 1947, PP. 1479-1484

[8] Chu, Lan J., Physical Limitations of Omni-Directional Antennas, Journal of Applied Physics, vol. 19, December 1948, pp. 1163-1175

[9] Yaghjian and Best, Impedance, Bandwidth and Q of Antennas, IEEE transactions on Antennas and Propagation , Vol. 53, No. 4, April 2005, pp. 1298-1324

[10] Burghard, George, "Station 1BCG", QST, February 1922, pp. 29-33

[11] Kelley and Hudson, "Hams Span the Atlantic on Shortwave!", QST, December 1996, pp. 28-30

[12] Moyer and Wostrel, The Radio Handbook, McGraw-Hill, 1931

[13] Ghirardi, Alfred A., Radio Physics Course, Radio & Technical Publishing, 1933

[14] Admiralty Handbook of Wireless Telegraphy, His Majesty's Stationary Office, 1932, pg. 799, Figure 452

Chapter 2
Short Verticals

2.0 Introduction

The purpose of this chapter is to introduce the reader to the basic limitations of antennas using only a single vertical conductor. Useful terms like radiation resistance (Rr), ground loss resistance (Rg), power lost in soil (Pg), equivalent height (h), etc, will be introduced and defined. Simple methods for estimating Rr and the reactive parts of the feedpoint impedance are given and at the end there is a discussion of the very high voltages and currents which can be present even with relatively low input powers (Pi). This is intended to serve as an introduction to the more antennas shown in later chapters.

2.1 Equivalent circuit for a short vertical

A equivalent circuit for an electrically short vertical is shown in figure 2.1.

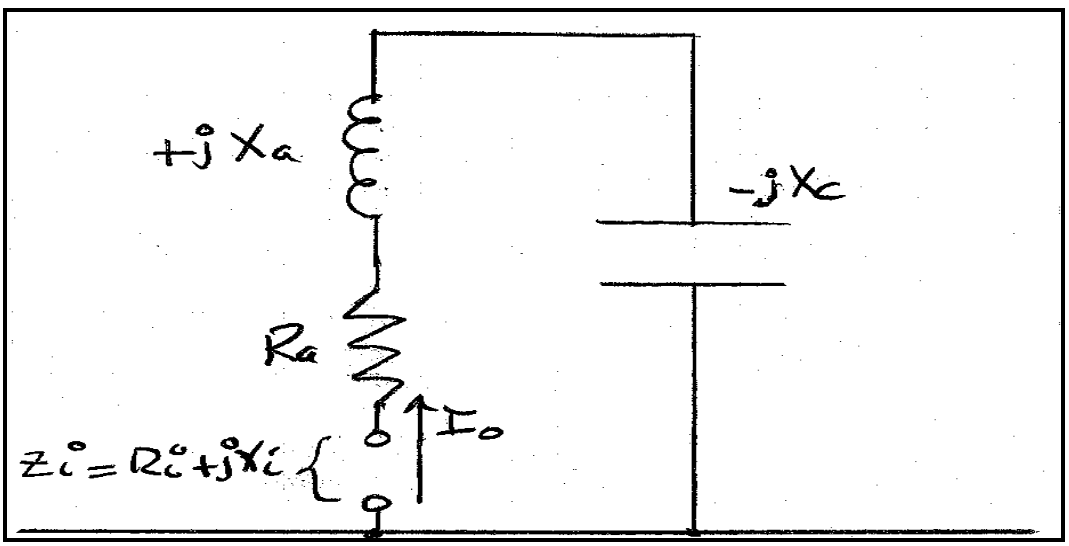

Figure 2.1 - Equivalent circuit for Zin.

Ra=Rr+Rg+Rloss represents the sum of radiation and loss resistances:

- Pi=Ra·Io^2
- Rr represents the radiated power
- Rg represents the loss in the soil close to the base (r<λ/2) of the antenna
- Rloss is the sum of conductor resistance (Rc), Losses due to leakage across insulators (Rin), and corona loss at wire ends (Rcor).

The inductor represents the energy stored in the magnetic component of the reactive near-field:

$$\mathbf{La} = \frac{\mathbf{Xa}}{\mathbf{2\pi f}} \qquad (2.1)$$

The capacitor represents the energy stored in the electric component of the reactive near-field:

$$\mathbf{Cc} = \frac{\mathbf{1}}{\mathbf{2\pi fXc}} \qquad (2.2)$$

The feedpoint impedance is Zi=Ra+jXi=Ra+j(Xa-Xc). In a short vertical operating well below resonance, Xc>>Xa so that Xi≈Xc with sufficient accuracy for most cases and in most cases Ra <<Xc.

Basically, a short vertical is a very small resistance in series with a large capacitive reactance!

2.2 Definition of Rr in a lossless antenna

The term "radiation resistance" (Rr) is used frequently so we need to be careful with our definition. A definition of Rr associated with a lossless antenna, can be found in most antenna books. A typical example is given in Terman[1]:

"The radiation resistance referred to a certain point in an antenna system is the resistance which, inserted at that point with the assumed current I_o flowing, would dissipate the same energy as is actually radiated from the antenna system. Thus

$$\boldsymbol{Radiation\ resistance = \frac{radiated\ power}{I_o^2}}$$

Although this radiation resistance is a purely fictitious quantity, the antenna acts as though such a resistance were present, because the loss of energy by radiation is equivalent to a like amount of energy dissipated in a resistance. It is necessary in defining radiation resistance to refer it to some particular point in the antenna system, since the resistance must be such that the square of the current times radiation resistance will equal the radiated power, and the current will be different at different points in the antenna. This point of reference is ordinarily taken as a current loop, although in the case of a vertical antenna with the lower end grounded, the grounded end is often used as a reference point."

2.3 Definitions for Rr, Pr, Pg and Rg over real ground

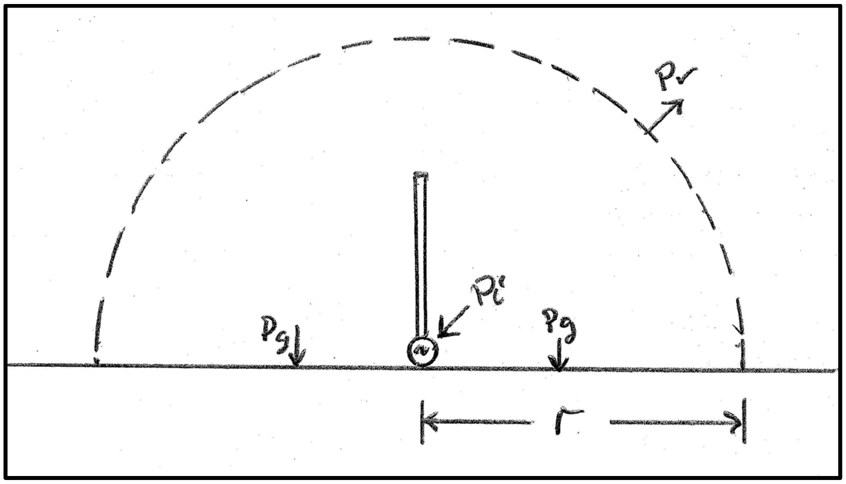

Figure 2.2 - Pr and Pg.

Figure 2.2 illustrates how the radiated power "Pr" and ground loss power "Pg" are determined for a monopole over real ground. The dashed line represents a hypothetical hemispherical surface enclosing the antenna. The hemisphere has a radius r: r=λ/2 is usually chosen because it is approximately the outer boundary of the reactive near-field. Pr is defined as the total power radiated through the hemisphere. Pg is defined as the power passing into the ground surface and dissipated in the soil.

For our purposes Rr and Rg are defined in terms of Pr and Pg:

$$\mathbf{Rr} \equiv \frac{\mathbf{P_r}}{\mathbf{I_o^2}}\ \boldsymbol{\Omega} \quad (2.3) \qquad \mathbf{Rg} \equiv \frac{\mathbf{P_g}}{\mathbf{I_o^2}}\ \boldsymbol{\Omega} \quad (2.4)$$

2.4 Rr from NEC modeling

Why do we care about Rr? The efficiency ($\boldsymbol{\eta}$) of the antenna will be:

$$\boldsymbol{\eta} \equiv \frac{\boldsymbol{P_r}}{\boldsymbol{Pi}} = \frac{\boldsymbol{Rr}}{\boldsymbol{Rr+Rg+Rloss}} \quad (2.5)$$

If we want an estimate of efficiency we need to have values for Rr, Rg and Rloss. We also need a value for Rr to calculate Pr from Io. Values for Rr are shown here, values for Rg and Rloss will be derived in later chapters. The vertical can be modeled over perfect ground to create a graph from which Rr can be read directly.

Figure 2.3 graphs Rr for a lossless #12 wire vertical for H=20'→100' at 137 and 475 kHz. We can see that Rr is very small even for heights of 100'. A λ/4 vertical would have Rr≈36Ω but in LFMF antennas Rr is typically smaller by a factor of 100 to 1000!

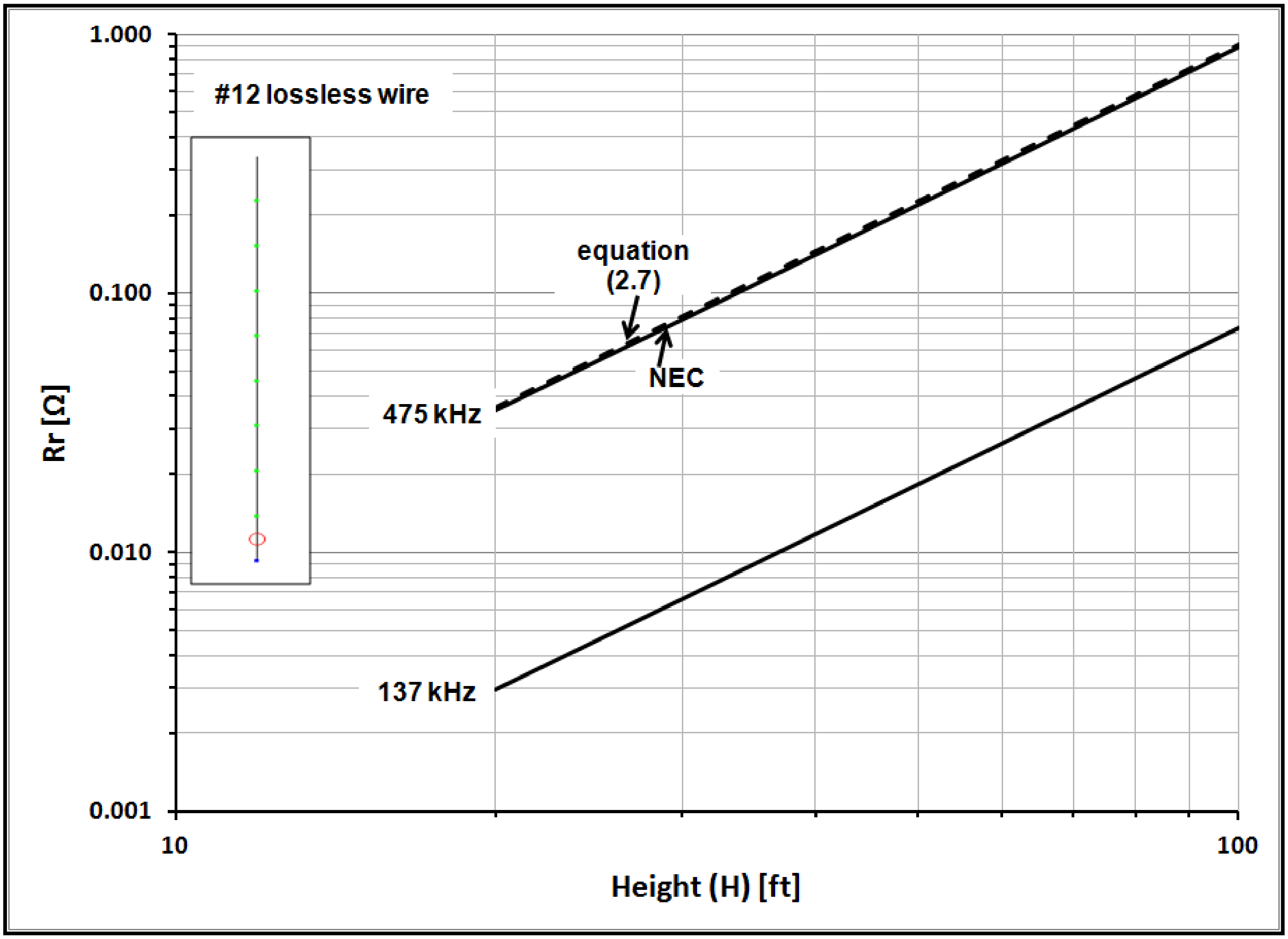

Figure 2.3 - Example of Rr variation with vertical height (H).

Conductors larger than #12 wire are often employed. To explore this two models were used, the first was simply a vertical wire where the diameter was varied from 0.081" (#12) to 6" but to simulate larger diameters and to reflect how larger diameters are actually implemented in practice, the model shown in figure 2.4 was used.

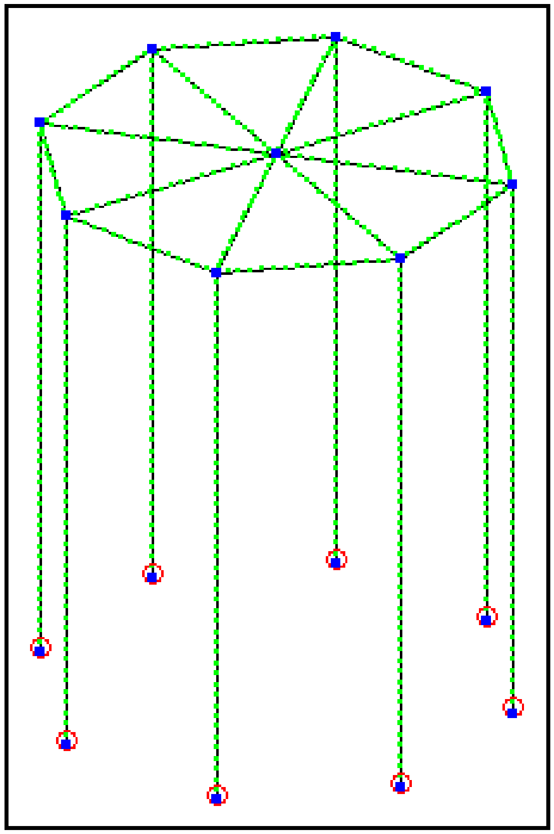

Figure 2.4 - A cage vertical.

For diameters up to a few feet, eight wires are more than adequate but for very large diameters, say 10'-40', adding more wires to the cage may be worth doing. Using a larger diameter conductor or more wires in a cage has the immediate benefit of reducing conductor loss (Rc). To simplify modeling of the cage vertical a source was placed at the ground end of each wire. In a real antenna the bottom ends of the vertical wires would be connected together with a skirt wire like that at the top. The bottom skirt wire is then driven against ground or, as shown in chapter 4, inductors are placed in each downlead and only one or two are driven. Figures 2.5 and 2.6 show the variation in Rr at 475 and 137 kHz as the conductor diameter (d) is varied from 0.080" (#12 wire) to 40' over a range of heights from 20' to 100'. It's interesting to note that for 0.08"<d<4' Rr goes down slightly as d is increased! Note that the contour for d=0.08" is for a solid #12 wire. The other contours are for a cage of #12 wires as shown in figure 2.4.

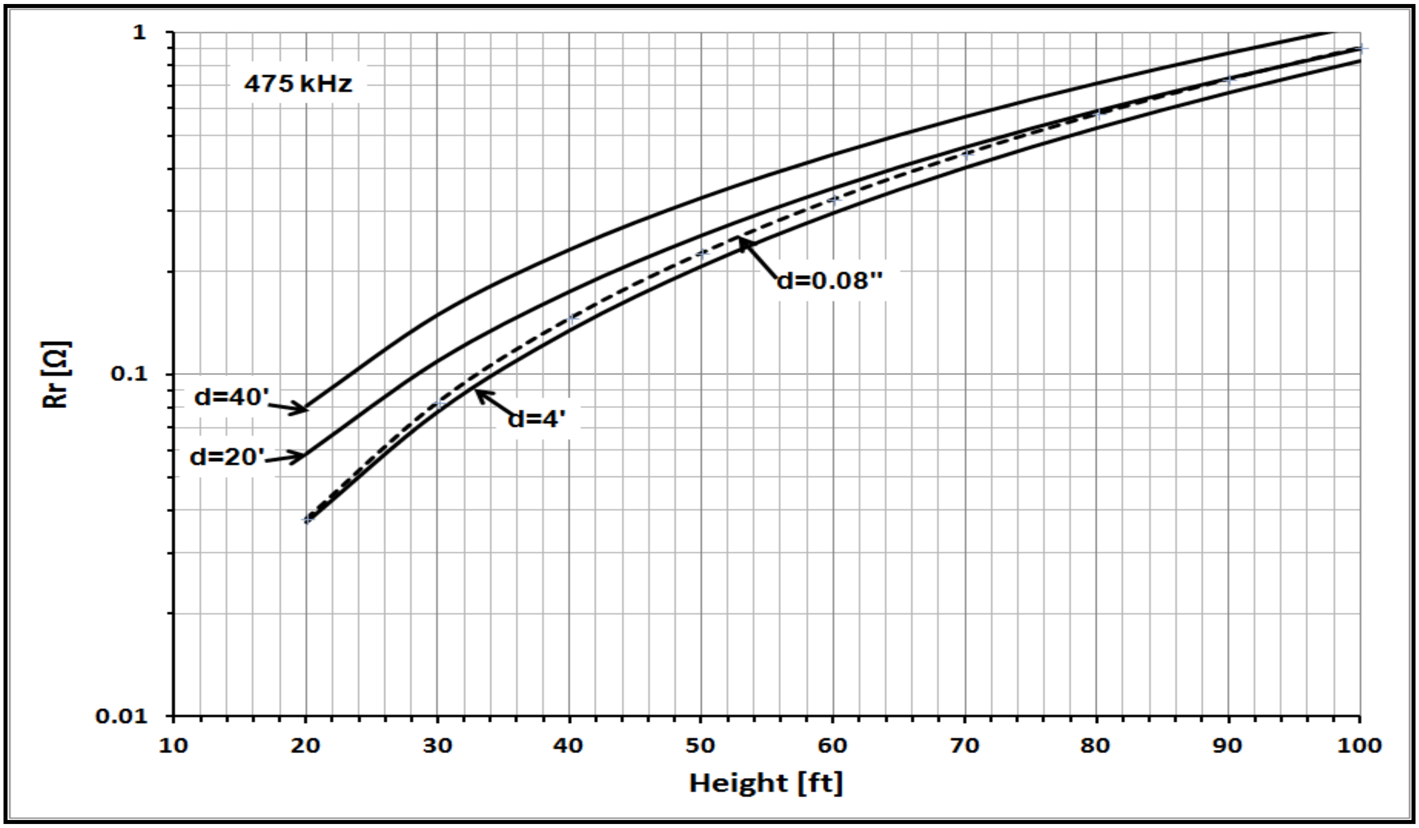

Figure 2.5 - Effect of conductor diameter on Rr at 475 kHz.

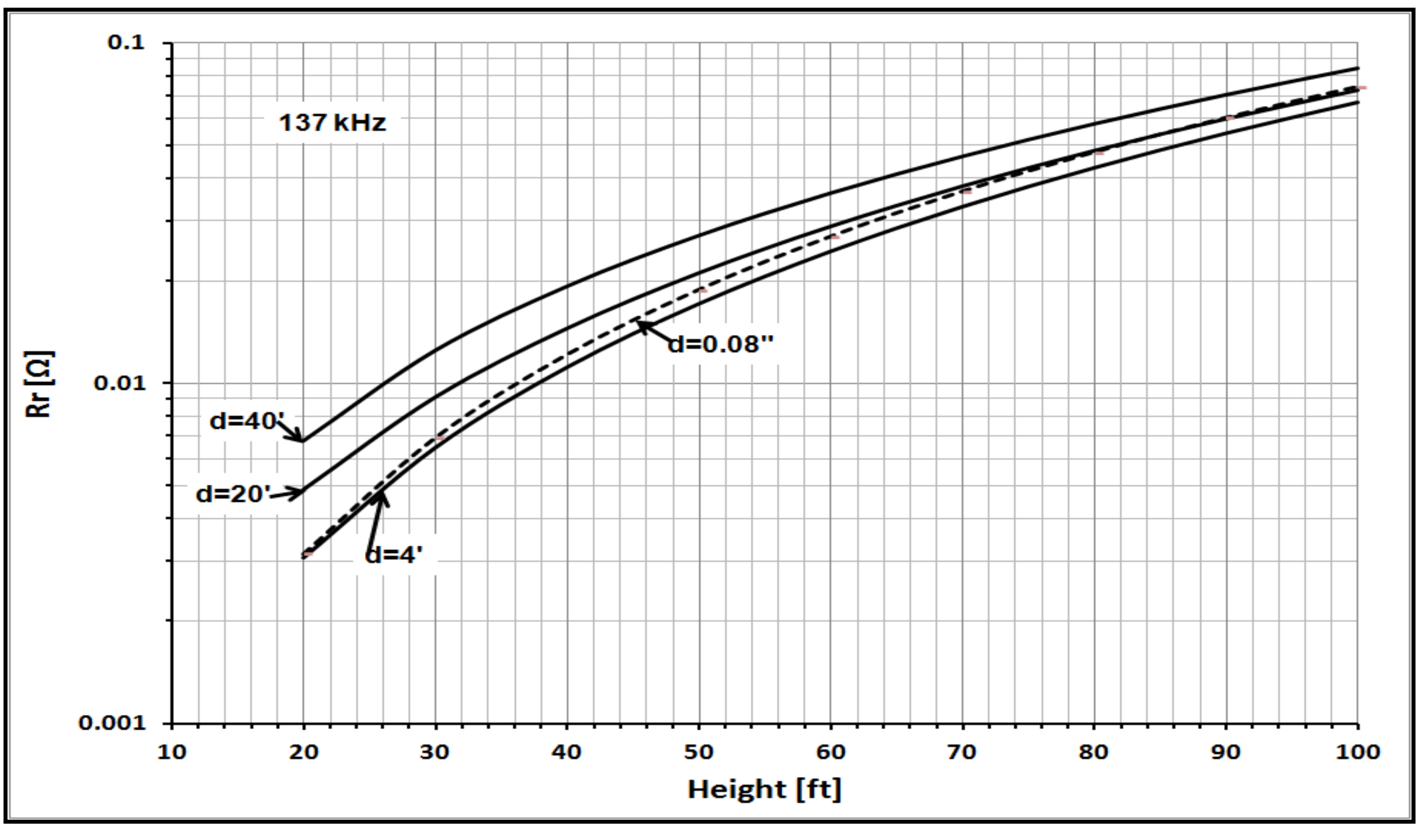

Figure 2.6 - Effect of conductor diameter on Rr at 137 kHz.

2.5 Calculating Rr

Rr can be calculated directly from the current distribution on the vertical. The solid line in figure 2.7 represents the current amplitude on a short vertical. The height can be expressed in a variety of units: feet, meters, fraction of a wavelength (0.1λ for example) or electrical degrees Gv. For antennas shorter than Gv=30° (H<0.083λ) the straight line in figure 2.7 is a very good approximation for the current distribution. If we sum (integrate) the product of the current and height we get an area A' (A'1 in figure 2.7) . If we state the height in electrical degrees (Gv) A' will have units of Ampere-degrees.

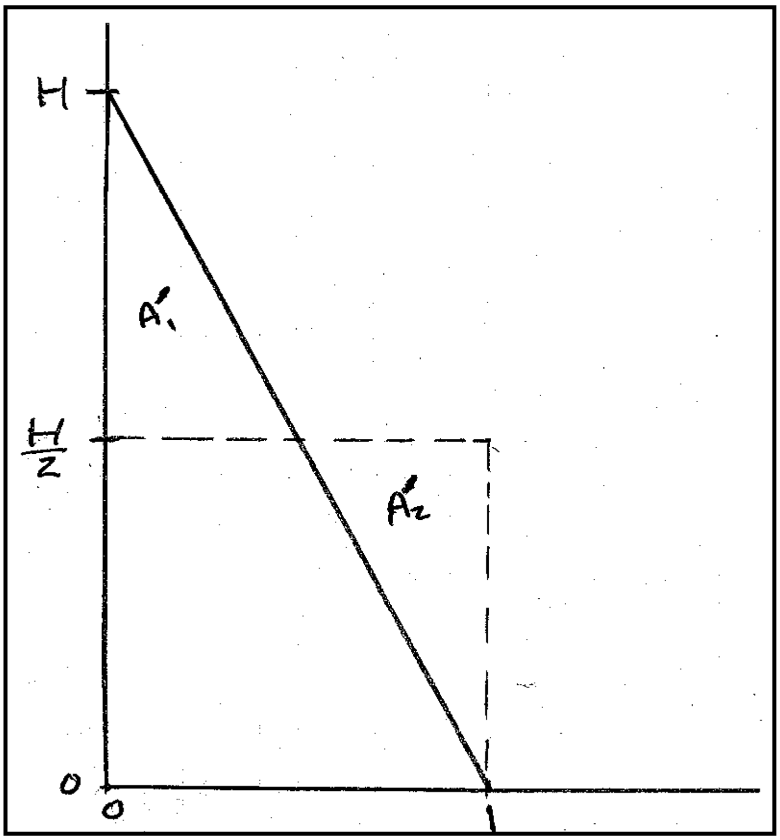

Figure 2.7 - current distribution on a short vertical, A'1=A'2

Laport[2] shows how A' can be used to compute the E-field strength (E) at a given distance (1 km):

$$\mathbf{E=kA' \quad [V/m]} \qquad (2.5)$$

When A' is in Ampere-degrees, k=0.00104 and E is the field strength in volts/meter at 1 km with Io=1A. The interesting thing about equation (2.5) is that it tells us our signal strength (for a given base current Io) will be a direct function of A'. If we can increase A' for the same base current the signal strength increases. Since

A' is a function of both H and the current distribution, if we increase the height and/or the amplitude of the current as we go up the antenna then E, at a given distance, for a given Io, will increase. As is shown in chapters 3 and 4, inductive loading and/or capacitive top-loading can be used to increase A'. Note that in figures 2.5 and 2.6 Rr is affected by the conductor diameter. If we look at the current distribution near the top of the vertical we find that the current is very close to zero for a thin wire but is not zero for very thick ones. This represents an increase in A' resulting in higher Rr.

Rr can be expressed in terms of A' [Ampere-degrees]:

$$\mathbf{Rr=0.01215A'^2\ [\Omega]} \quad (2.6)$$

For a thin wire vertical with a triangular current distribution when Io=1A, A'=Gv/2 and we can express Rr as:

$$\mathbf{Rr \approx 0.003Gv^2\ [\Omega]} \quad (2.7)$$

Equation (2.7) provides a quick estimate of Rr for short unloaded verticals. The dashed line in figure 2.3 shows the comparison between NEC and equation (2.7). The correspondence is close.

2.6 Effective height h

The concept of "effective height" (h) is closely related to A'. The following definition of is taken from Terman[2]:

"*The effective height of a grounded vertical-wire antenna is the height that a vertical wire would be required to have to radiate the same field along the horizontal as is actually present if the wire carries a current that is constant along its entire length and of the same value as at the base of the actual antenna."*

The solid line (A'1) in figure 2.7 shows the typical current distribution. The dashed line (A'2) represents the same area as A'1 with constant current over H/2. We say that the antenna has an "equivalent height" h=H/2. More generally we can find the equivalent height by computing A' for an arbitrary distribution and then substituting a height which has the same A' with constant current along the vertical. For example, in resonant λ/4 vertical h=(2/π)H≈0.64H. Equivalent height is also used for verticals in a receiving array where the open circuit voltage at the feedpoint (Vo) is: Vo=Eh. Where E is the electric field vector parallel to the conductor in V/m and h is the equivalent height in meters.

2.7 Xi and Xc from modeling

Why do we care about Xi or Xc? As shown in figure 2.8, a inductor is needed at the feedpoint to resonate the antenna . For resonance XL=Xi=Xc-Xa . We need to know the value of that inductor but it's value is derived from Xi, so we also need to estimate Xi!

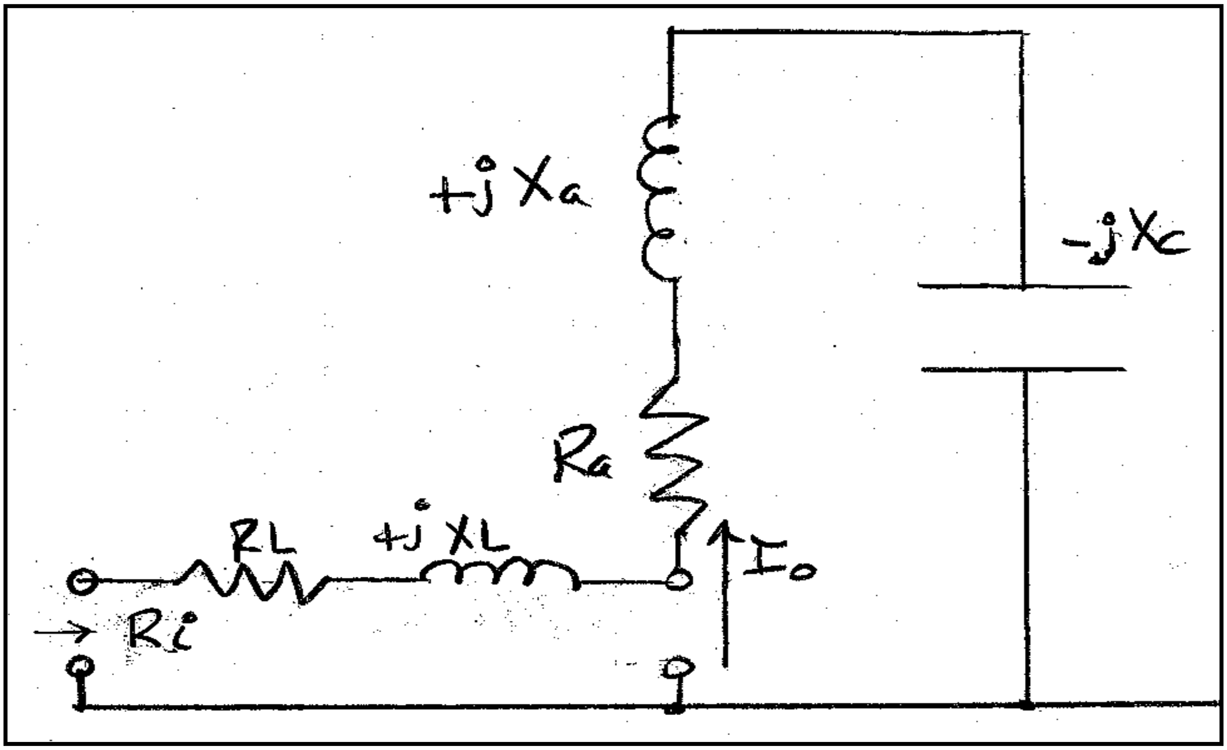

Figure 2.8 equivalent circuit of a short vertical with a resonating inductor.

Any practical inductor will have a series loss resistance (RL) and RL=XL/QL, where QL is the inductor Q. In many amateur installations the efficiency of the antenna will be dominated by inductor losses so from a practical point of view very early in the design process we need to know how large an inductance will be needed. Values for Xi (Xi=Xc-Xa) for a vertical with height H and diameters from 0.081" to 6" are shown in figures 2.9 and 2.10 for 630m and 2200m.

Unlike Rr, Xi is very sensitive to conductor diameter. At a given height, a larger diameter conductor will have less conduction loss but more importantly the size of the tuning inductor and it's associated losses is reduced.

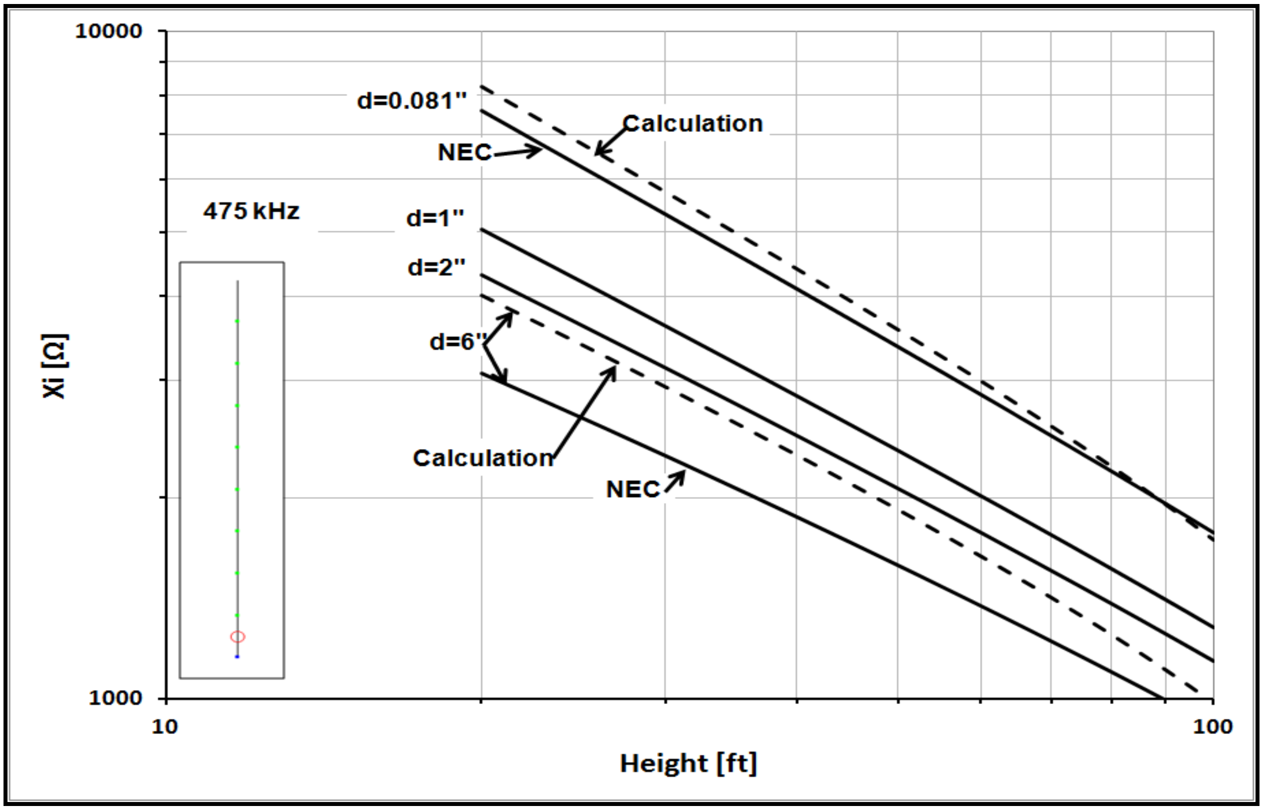

Figure 2.9 -Variation in Xi with diameter at 475 kHz.

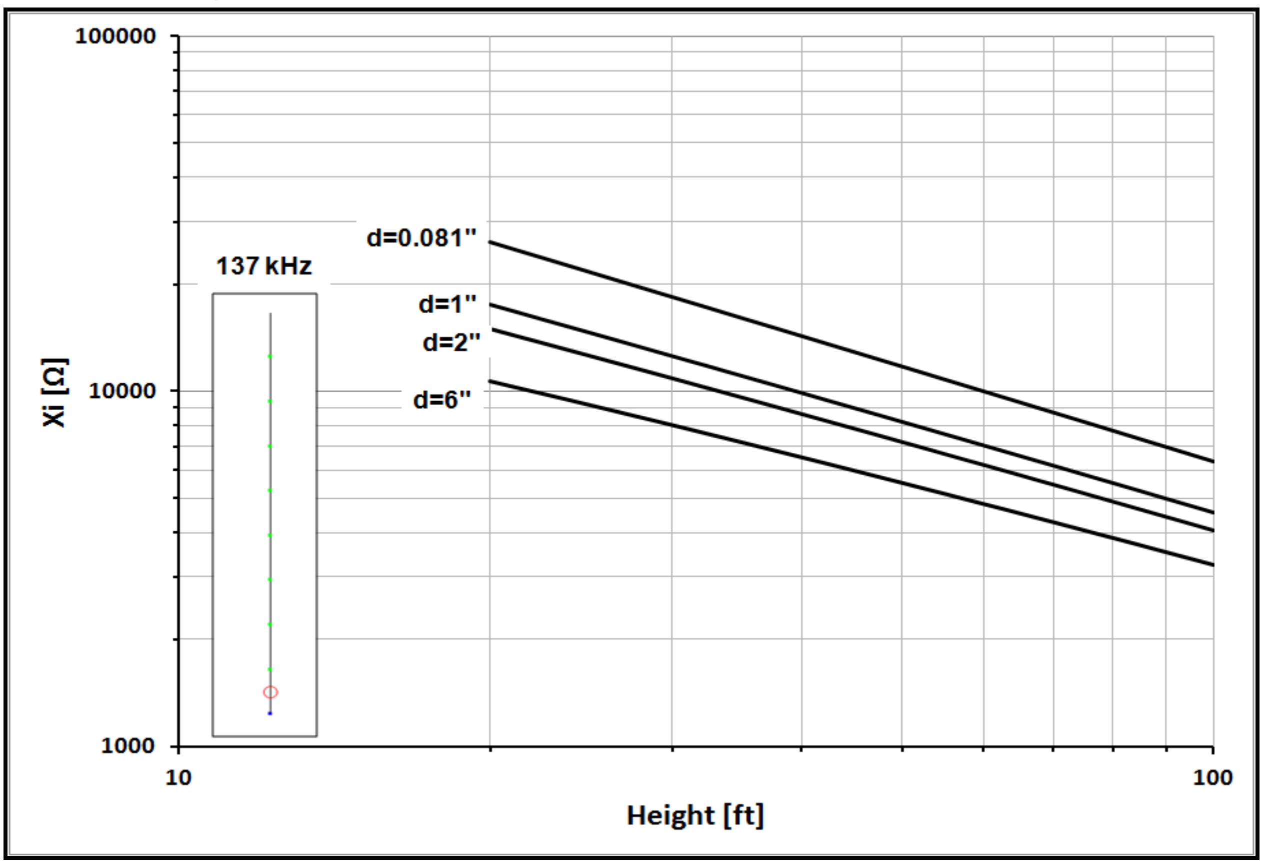

Figure 2.10 - Variation in Xi with diameter at 137 kHz.

2.8 Calculating Xc and Xa

If modeling is not available we can calculate Xc and Xa. It's possible to view a vertical as a single wire non-uniform transmission line[3] with an average characteristic impedance of Za and use expressions for the input impedance of either short or open-circuited transmission lines as suggested in figure 2.11.

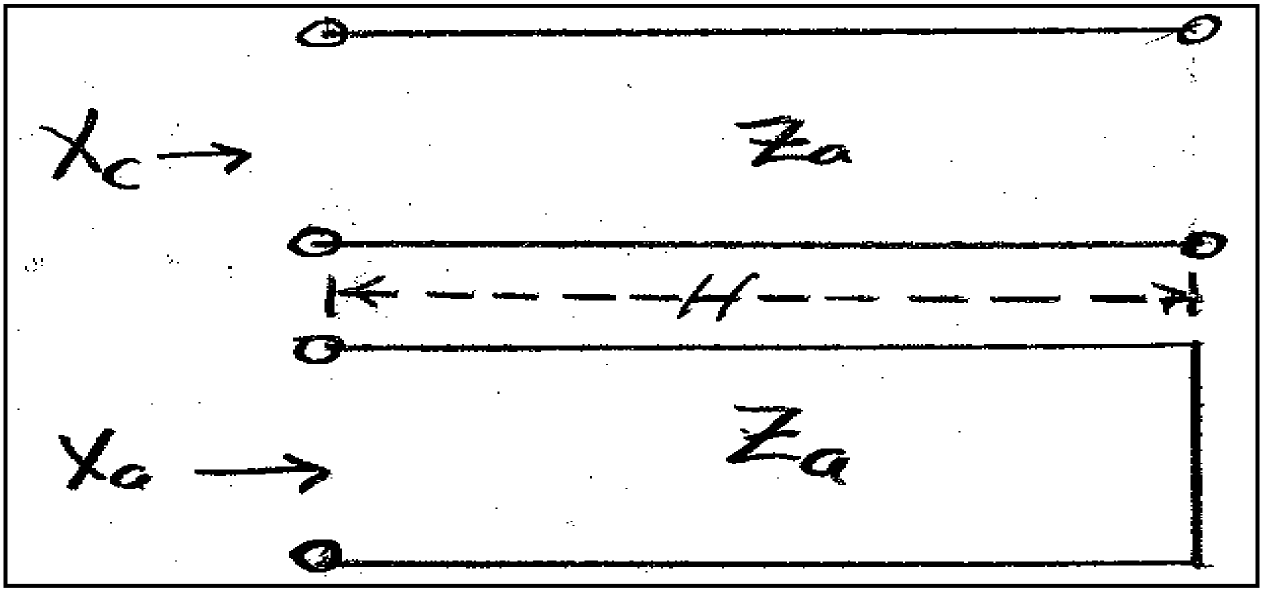

Figure 2.11 - O/C and S/C transmission lines with Zo=Za and length H.

Za can be calculated from:

$$\mathbf{Za = 60\left[\ln\left(\frac{4H}{d}\right) - 1\right]} \qquad (2.8)$$

Where: d is the conductor diameter and H is the height <u>in the same units</u>.

With Za we can calculate Xc and Xa from:

$$\mathbf{Xc = \frac{Za}{tanH}} \qquad (2.9)$$

$$\mathbf{Xa = Za \cdot tanH} \qquad (2.10)$$

Where H is the height in degrees or radians. How good is this approximation? The dashed lines in figure 2.9 provide a comparison. For the #12 wire (d=0.081") the agreement is very good but for large diameters the calculation over-estimates Xi so the calculation has to be viewed as an approximation.

2.9 XL, RL and efficiency

Adding a tuning inductor (RL+jXL) as indicated in figure 2.8:

$$\mathbf{QL = \frac{XL}{RL}} \quad (2.11)$$

Equation (2.11) shows the relationship between QL, XL and RL. QL can range from 100 to >1000. In general, for a given inductor, QL at 137 kHz will be ≈0.54 QL at 475 kHz or a little less if the QL at 475 kHz is near it's peak value (see chapter 6). While very high QL inductors are possible most of this discussion will assume QL=200 at 137 kHz and 400 at 475 kHz because these values are practical with modest effort but keep in mind that higher values are possible as explained in chapter 6.

Antenna efficiency (**η**) is:

$$\boldsymbol{\eta} = \frac{\textbf{power radiated}}{\textbf{input power}} = \frac{\mathbf{R_r}}{\mathbf{R_i}} = \frac{\mathbf{R_r}}{\mathbf{R_r + RL + Rg + Rc + \cdots}} \quad (2.12)$$

We can get a good feeling for the effect of loading inductor losses (RL) on efficiency by assuming Ri= RL+Rr (i.e. ignoring other losses) and calculate the efficiency as shown in figures 2.12 and 2.13. QL=200 at 137 kHz and 400 at 475 kHz are assumed. Figure 2.12 is truly bad news. For example with H=20', at 137 kHz **η**=0.0024% and at 475 kHz **η**=0.20% and that doesn't consider any other losses! Increasing H to 100' makes a great difference. At 137 kHz **η**=0.24%, still very low but a factor of 100 improvement. With 100W output from the transmitter, to radiate the allowed maximum powers the antenna will have to have **η**> 2% at 475 kHz and **η**>0.33% at 137 kHz. There are horizontal dashed lines corresponding to these values in figure 2.12. We can see from the graph (for a simple vertical) a minimum height of 45' on 630m and >100' on 2200m is needed. Note, the efficiency scale is <u>logarithmic</u>, a small change in height means a large change in efficiency! As if we're not already depressed enough the y-axis in figure 2.12 can be converted to dB to better illustrate the effect of losses on our signals as shown in figure 2.13. The signal reduction for this range of heights is particularly severe on 2200m.

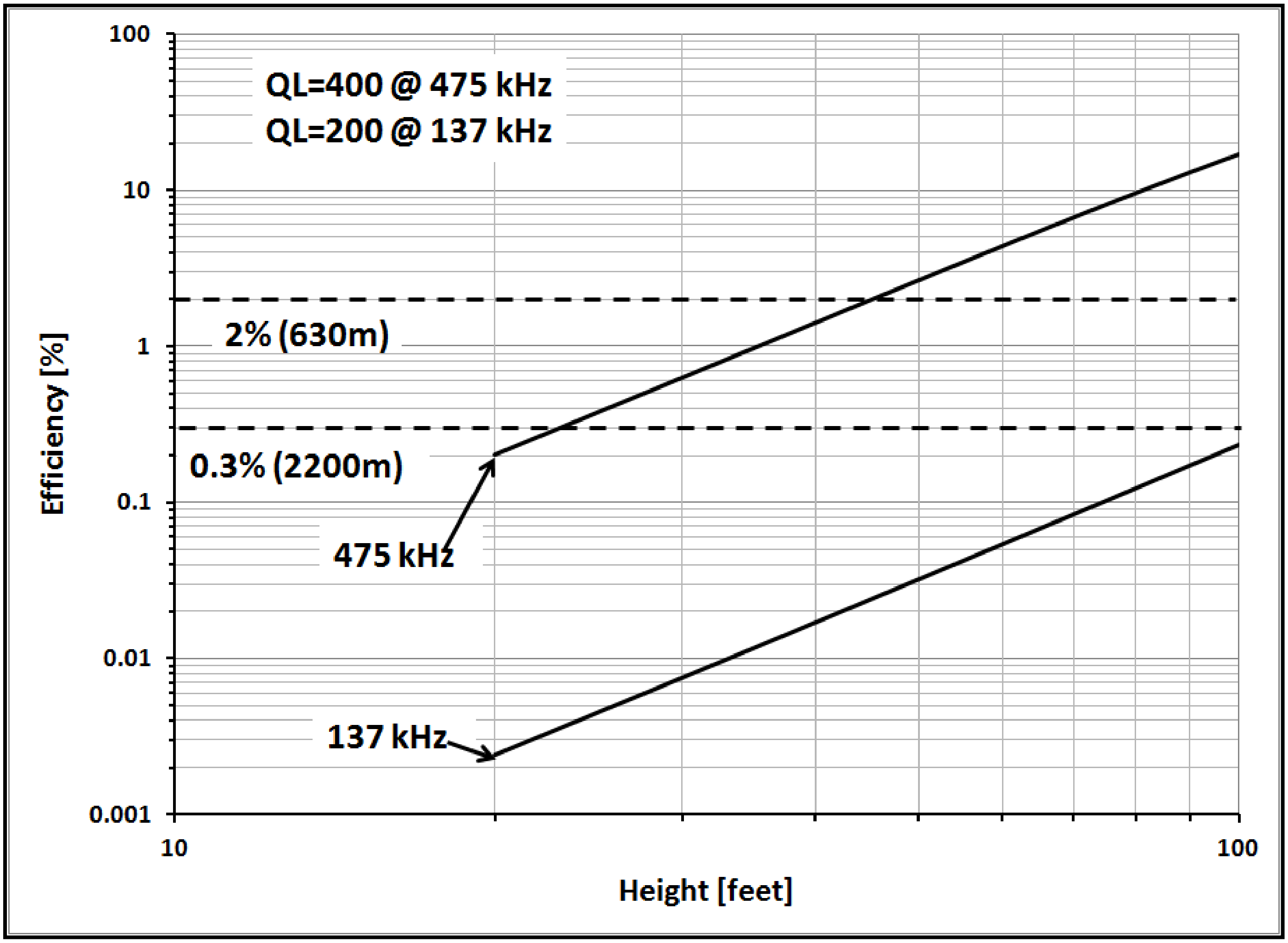

Figure 2.12 -Efficiency using a QL=200 and 400 loading inductors.

These graphs make an important point:

Maximizing height is a vital for improving efficiency!

RL is the dominant loss throughout this range of H, especially as we go lower in frequency. This observation is important because it tells us what our design priorities must be. The value of RL is tied directly to the value of XL (XL≈|Xc|) through QL. The message is very clear:

To reduce RL we must reduce Xc!

As will be shown in chapter 3, once height has been maximized, top-loading becomes the primary tool for reducing Xc.

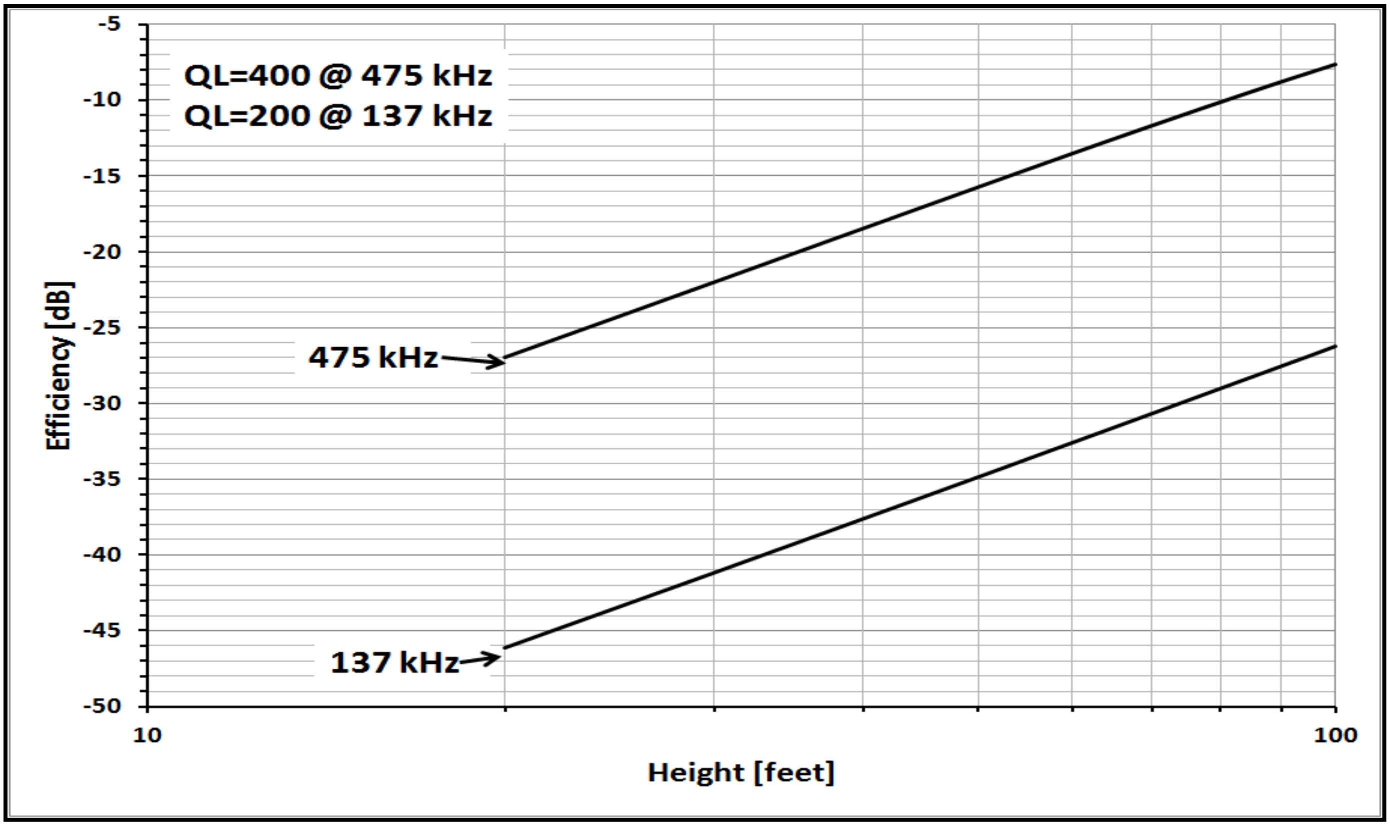

Figure 2.13 - Efficiency stated in dB=10 LOG(efficiency)

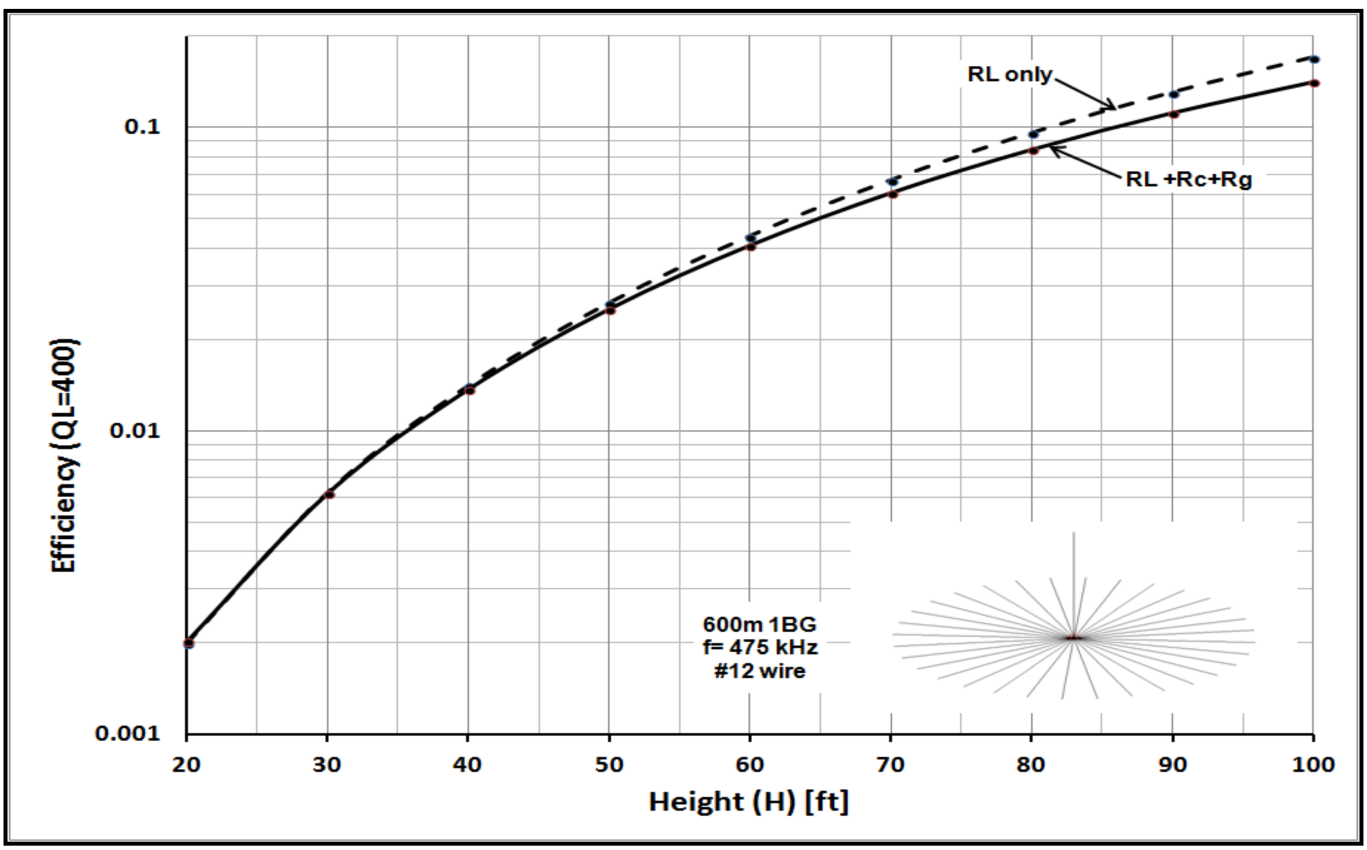

Figure 2.14 - Effect on efficiency from Rg and Rc.

To this point the effect of ground loss (Rg) and conductor loss (Rc) has not been included. A sample including Rg+Rc is shown in figure 2.14 for a vertical with 32 radials over average ground (0.005 S/m, Er=13). Note that at smaller values of H, where large values are needed for XL, the loss in RL dominates! This is treated in more detail in chapter 5.

2.10 Voltages and currents

Unfortunately low efficiency is not the only bad news. Base currents (Io) and feedpoint voltages (Vo) can be very high. The following discussion is for a simple vertical without top-loading. As shown in chapter 3, top-loading significantly reduces Io and Vo.

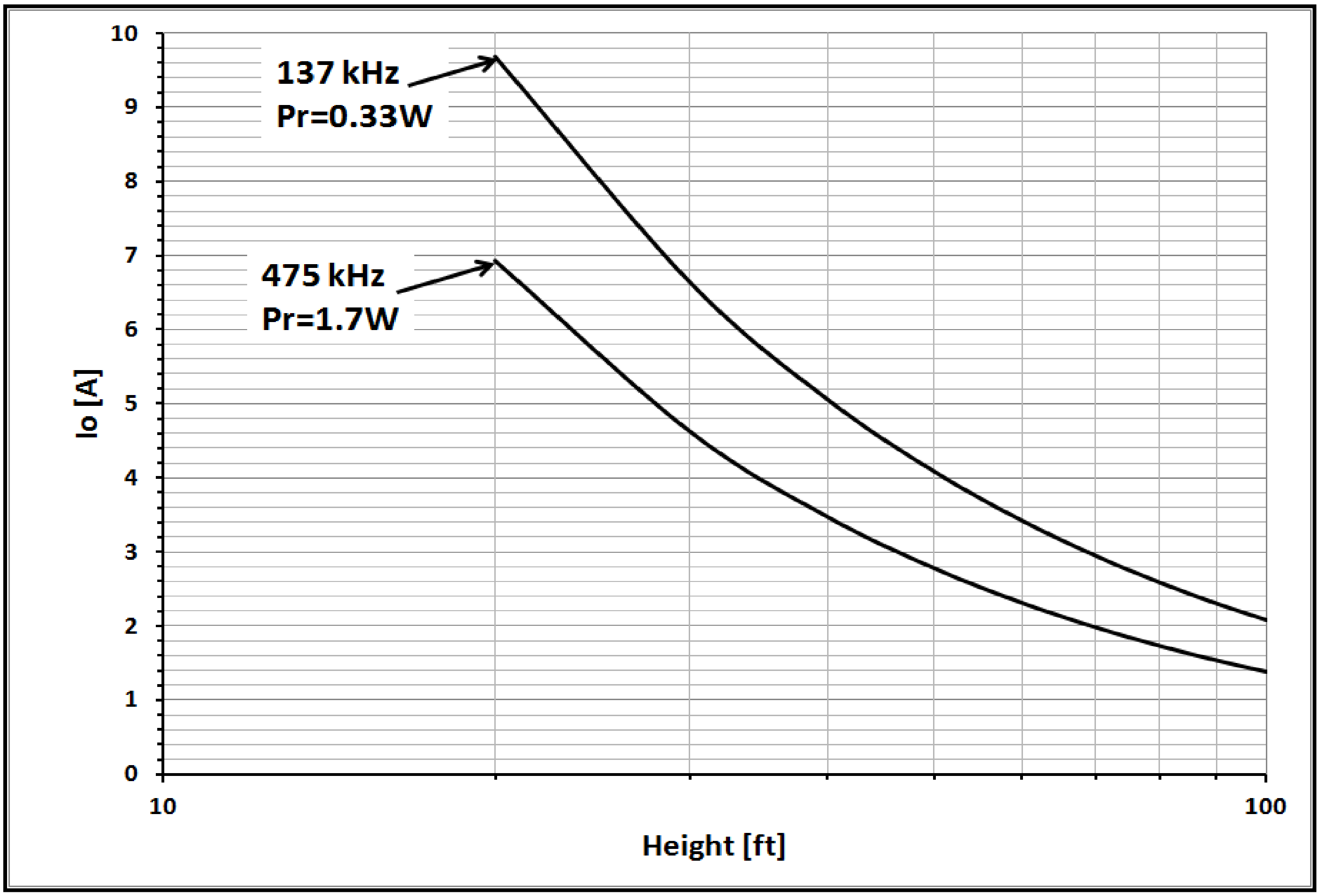

Figure 2.15 - Base current (Io).

Figure 2.15 shows base current (Io [Arms]) as a function of H. These are the rms currents required to produce the allowed radiated power on each band. Figure 2.16 shows the Pi required to produce the allowed Pr on each band for a given loading inductor Q. If you wish to use a simple 20' vertical on 137 kHz radiating the maximum allowed power you'll have to provide Pi≈9kW!

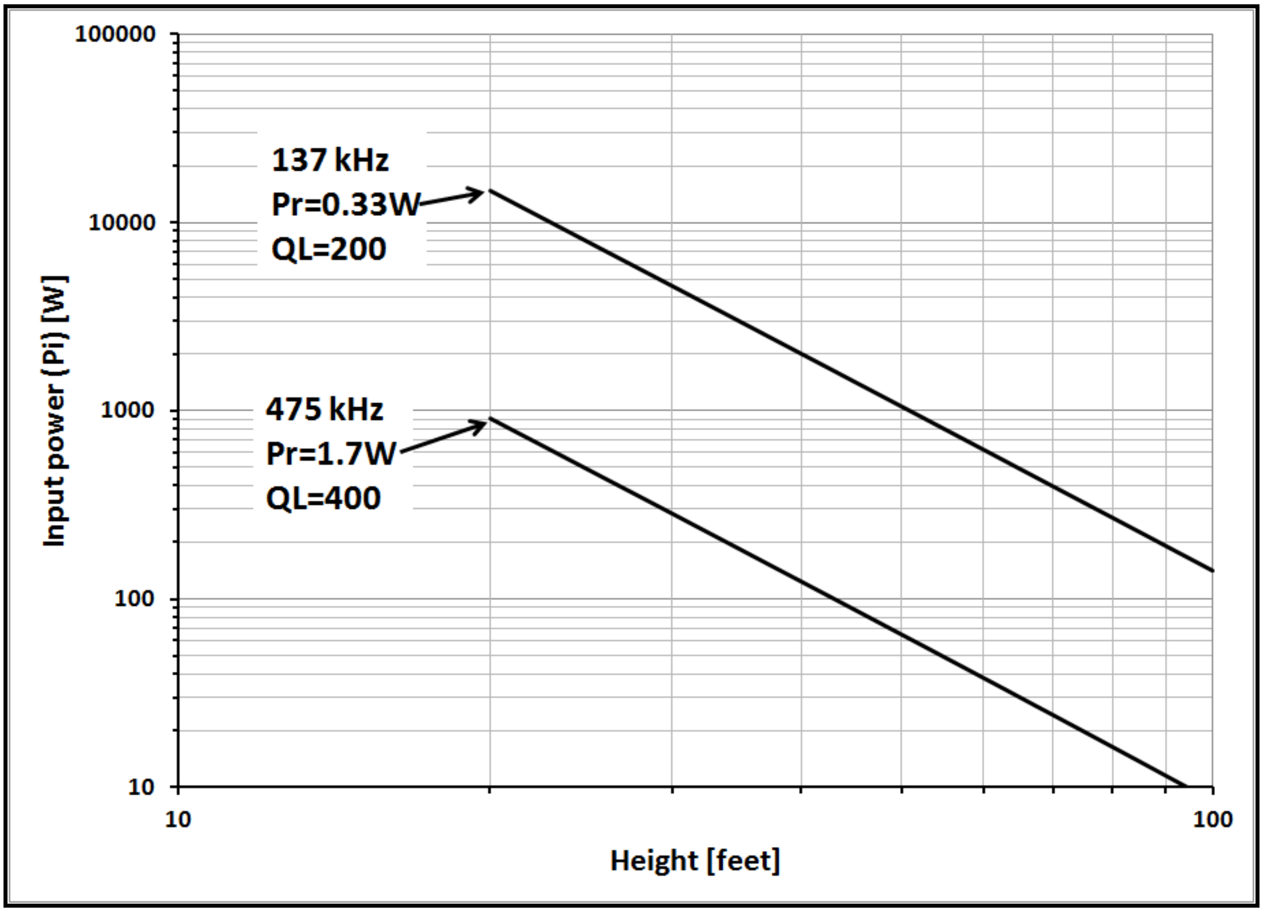

Figure 2.16 -Input power (Pi) needed to produce Pr.

As shown in figure 2.1 the input current (Io) flows through Ra, +Xa and -Xc. In short antennas Ra and Xa are very small compared to Xc, the capacitive reactance. As shown earlier Xi≈Xc and Xc will be very large:

$$\mathbf{Vo = IoXi} \quad (2.13)$$

The voltage across the feedpoint (Vo) will be very high as indicated in figure 2.17. A 20' vertical at 137 kHz with Pi≈9kW and Pr=0.33W will have Vo≈300kV! Which is of course absurd, we cannot work with these voltage levels.

Given the very modest radiating power allowed, these voltage levels can come as an unpleasant surprise when a hard won increase in transmitter power unexpectedly causes the loading coil to go up in flames or there is arcing across the base insulator or within tuning network components.

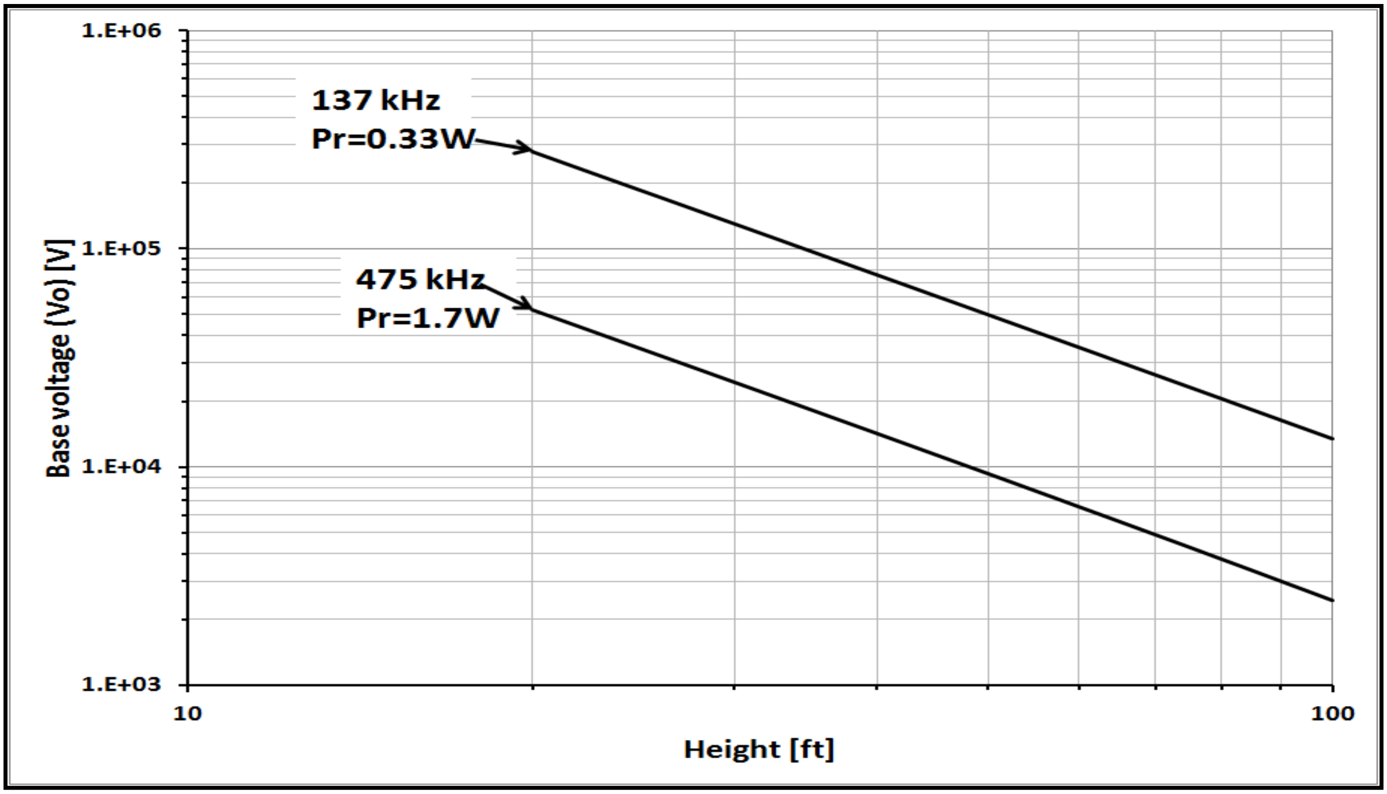

Figure 2.17 - Base voltage when radiating allowed Pr.

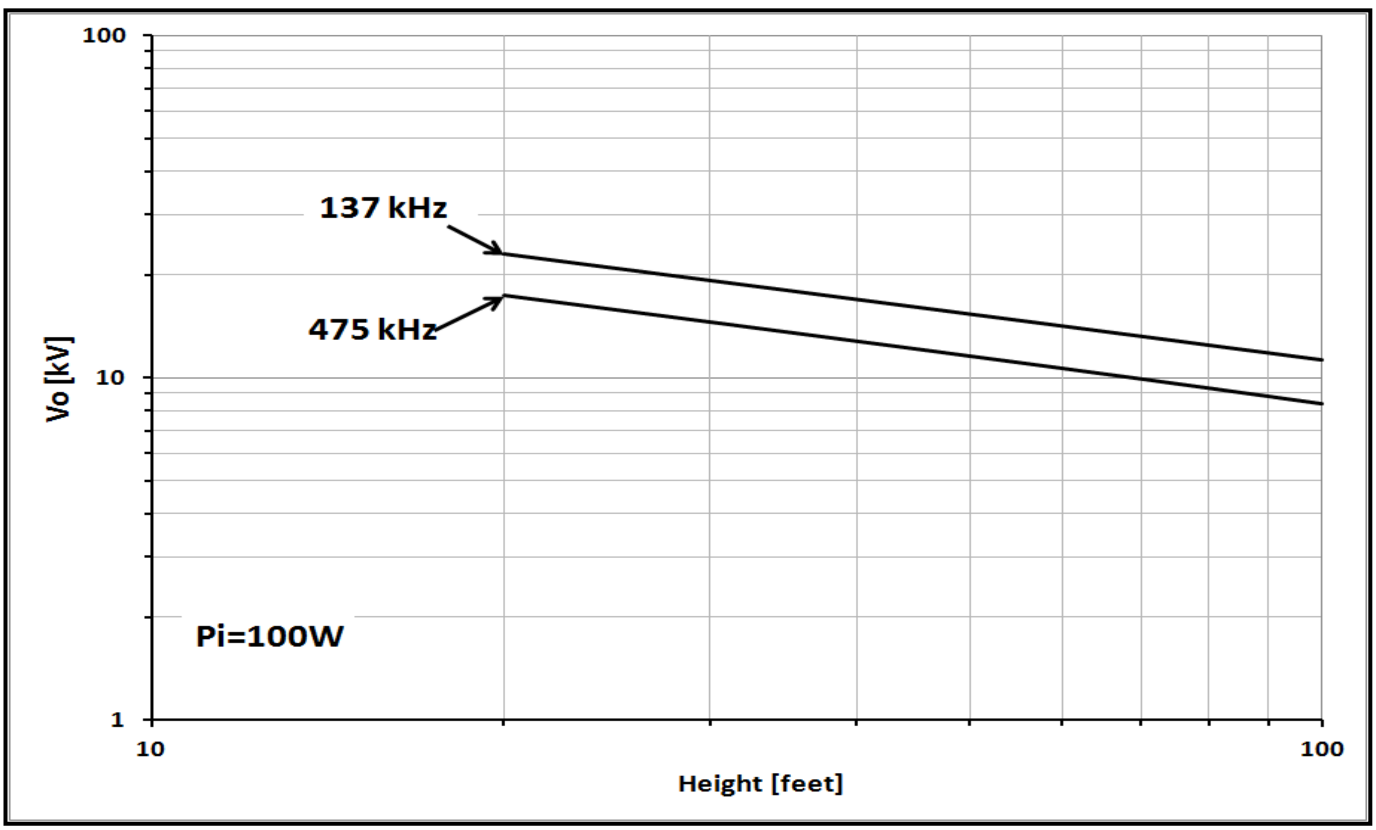

Figure 2.18 - Vo with Pi=100W.

For most amateurs Pi≤100W is more realisitc but even at this greatly reduced power Vo can still be many kV as shown in figure 2.18.

Why such a small reduction in Vo with a large reduction in Pi? Io and Vo vary as the square root of the power ratio:

$$\frac{V_1}{V_2} = \frac{I_1}{I_2} = \sqrt{\frac{P_1}{P_2}} \qquad (2.14)$$

Cutting the power in half only reduces Vo or Io by a factor of 0.707! This further reinforces the advice to minimize Xc. We must be very respectful of the voltages present on these antennas even at seemingly low power levels. BE CAREFUL!

Summary

This chapter makes the following points:

…make the height as tall as practical….

…short verticals require large lossy tuning inductors…

… inductor loss may totally dominate the efficiency…

…the base voltages across the tuning inductors will be very high even at low power levels…

References

[1] Terman, Frederick E., Radio Engineers Handbook, McGraw-Hill Book Company, 1943. This is a very useful book!

[2] Laport, Edmund, Radio Antenna Engineering, McGraw-Hill, 1952. You can find this one free on-line by Googling Edmund Laport.

[3] Schelkunoff and Friis, Antennas, Theory and Practice, page 426

Chapter 3

Capacitive Top-Loading

3.0 Efficiency

The discussion in chapter 2 made it clear that the equivalent series resistance (i.e. RL) of the tuning inductor can be a major loss contributor. A critical part of achieving higher efficiency is the reduction of capacitive reactance at the feedpoint (Xi) because decreasing Xi reduces the size of the loading inductor and its associated RL. In addition, often steps to reduce inductance can also increase Rr simultaneously.

Efficiency (**η**) in terms of Rr and RL:

$$\eta = \frac{R_r}{R_r + R_L} = \frac{1}{1 + \frac{R_L}{R_r}} \qquad (3.1)$$

Clearly we want RL small and Rr large! The reason for this rather obvious comment is that some top-loading schemes which reduce Xi also reduce Rr rather than increase it and at some point the efficiency may actually start to fall even though we continue to reduce Xi. This happens in top-loading schemes with sloping wires with currents opposing the current in the main vertical conductor. This is discussed in section 3.9. On most of the following graphs efficiency is stated either in percent (%) or as a decimal, although in some cases efficiency is given in -dB to illustrate the effect of losses or improvement on the signal for some change.

For amateurs new to LFMF the first antenna needs to be as simple as possible to get on the air. In the first part of this chapter we're going to keep it simple and assume we have only one or two supports, a roll of wire and some insulators. We'll begin with a example showing Rr, Xi and efficiency and then go on to explain why top-loading is so effective in reducing Xi and increasing Rr. Subsequent sections give examples of more complex top-loading arrangements. The discussion relies heavily on NEC modeling but, as in chapter 2, simple expressions and graphs for pencil and paper calculations are given.

At this point it will be assumed that the height (H) has been made as tall as practical and we are now turning to capacitive top-loading to improve efficiency. Many different variables affect the capacitance introduced by the top-loading structure:

- The number and/or length of umbrella wires
- Whether or not there is a skirt tying the ends of the wires together
- The location of the tuning inductor along the vertical conductor
- conductor sizing

As in chapter 2, tuning inductor QL=400 at 475 kHz and 200 at 137 kHz is assumed. Keep in mind that efficiency determined from only RL and Rr is an upper limit, i.e. the best we can do. Adding more losses only reduces efficiency. Once we've reduced RL as much as possible we can deal with other losses. Reducing Xi has the further benefit of reducing the voltages and currents at the base. For the most part, if you make some change in your antenna which reduces the inductance required to resonate the antenna that change is likely to improve your efficiency. This is a useful guide when "fiddling"! One important point, most examples have symmetric wire arrangements because it's easier to model but symmetry is not required! Irregular wire lengths work fine. In general we need to be opportunistic, connect top-loading wires to whatever support is available, even when the supports are at very different heights. This goes right back to Woodrow Smiths advice[1]:

".... the general idea is to get as much wire as possible as high in the air as possible....."

3.1 Efficiency with top-loading

We can use the "T" antenna shown in figure 3.1 to illustrate how effective capacitive top-loading is. Efficiency in percent (%) as a function of the length of the top wire (L) at 475 and 137 kHz is shown in figures 3.1 and 3.2. Note that there is a small sketch of the antenna under discussion on many of the graphs. This serves as a reminder of which antenna we're talking about. The notation "600m T1" is the title for that model in the NEC model files. Just some bookkeeping.

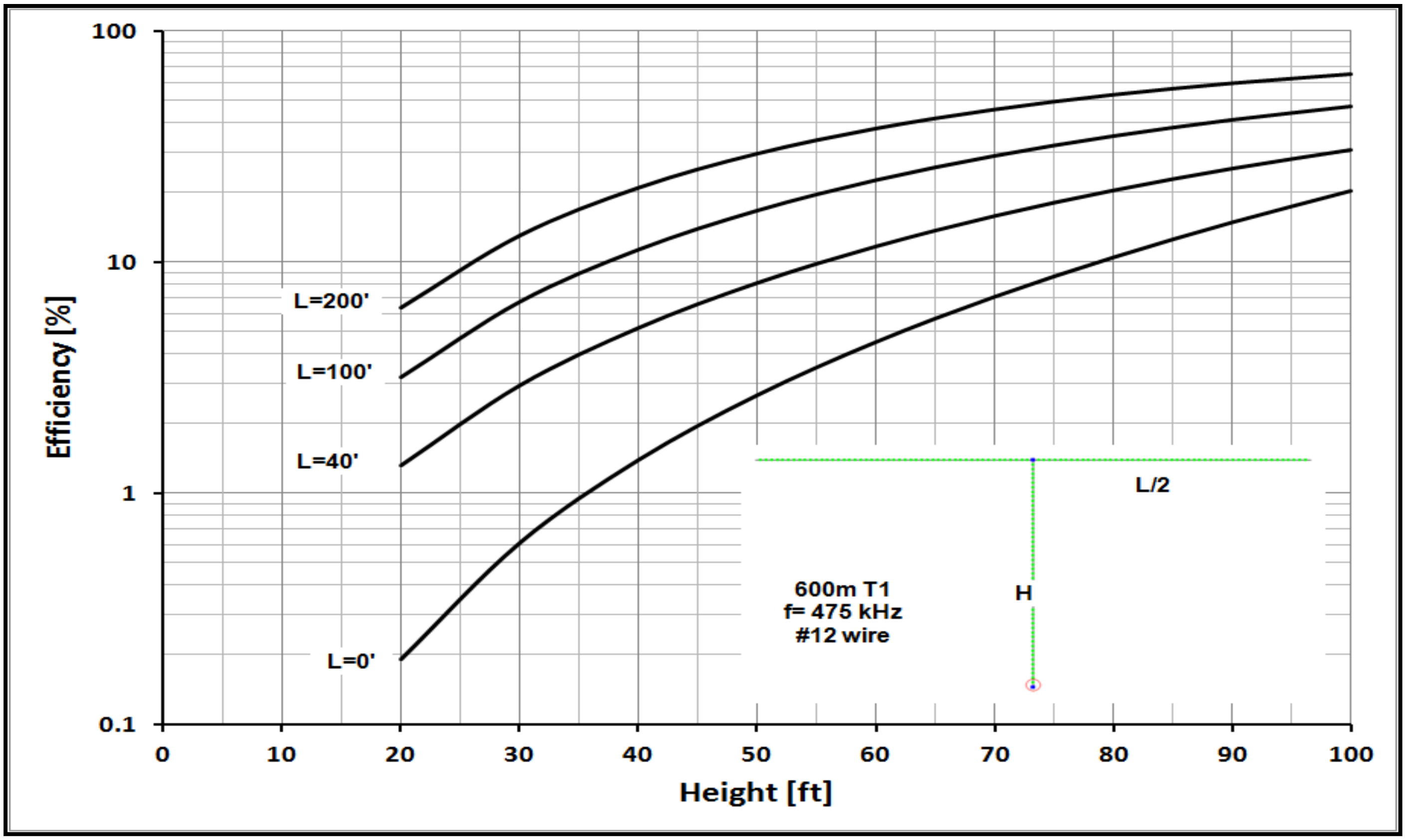

Figure 3.1 - Efficiency at 475 kHz with base tuning.

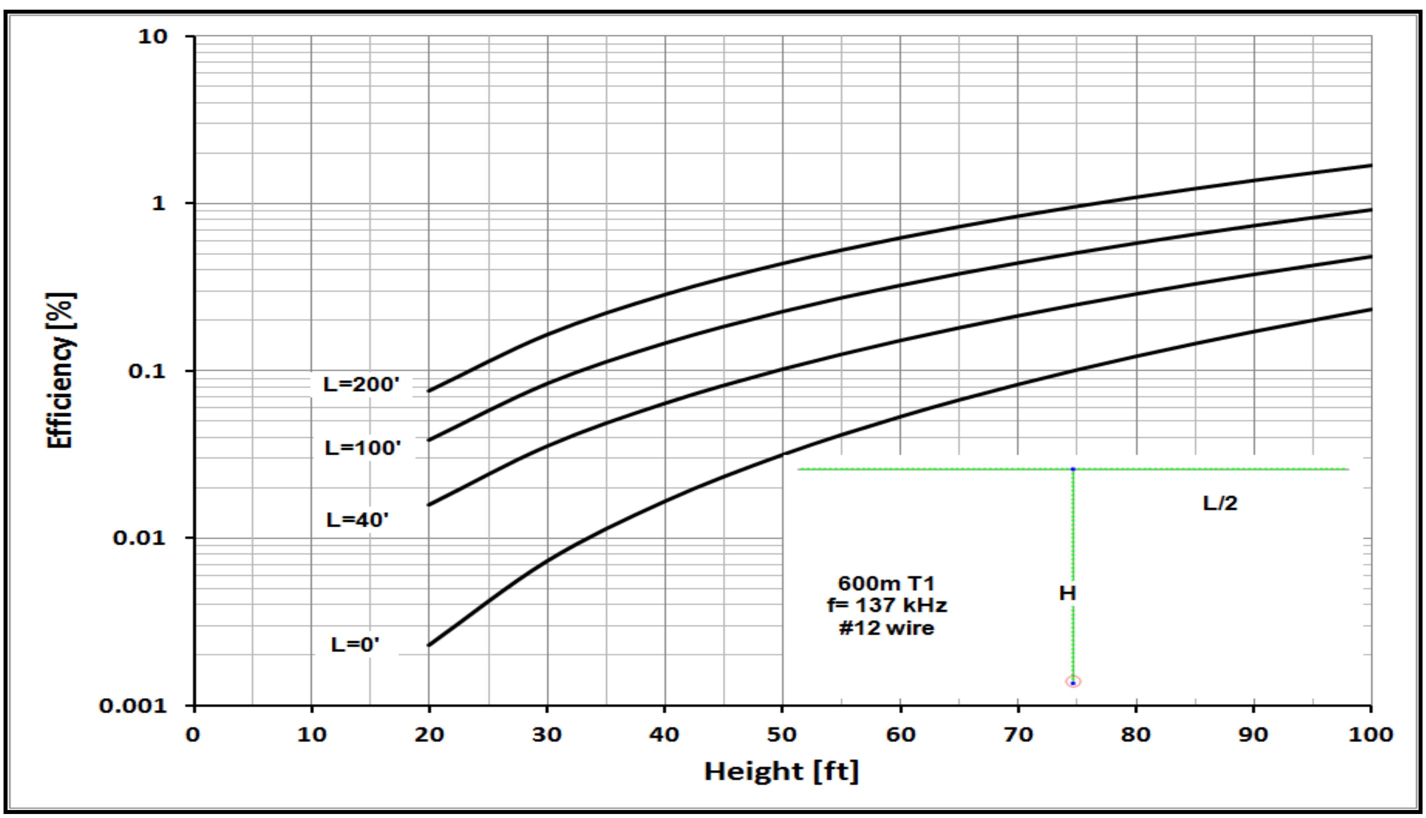

Figure 3.2 - Efficiency at 137 kHz with base tuning.

The L=0 line represents the case with no top-loading, just the bare vertical with a loading coil at the base. In these models the value of the tuning inductor was adjusted to maintain resonance as L and H were changed. #12 wire conductors are assumed. Even a small amount of top-loading increases efficiency. As an example, for L=0 and H=20', η=0.19% at 475 kHz. Keeping H=20' but adding a 40' top-wire, η=1.3%, a factor of 6.8X! Taking L to 100' increases the efficiency by 18X.!

Height and capacitive top-loading are keys to improving efficiency!

3.2 Xi with top-loading

To understand why the efficiency improves we need to look closer. For the T antenna the efficiency improvement is driven by increasing Rr and decreasing Xi simultaneously. This section explains what's happening with Xi and the next section looks at Rr.

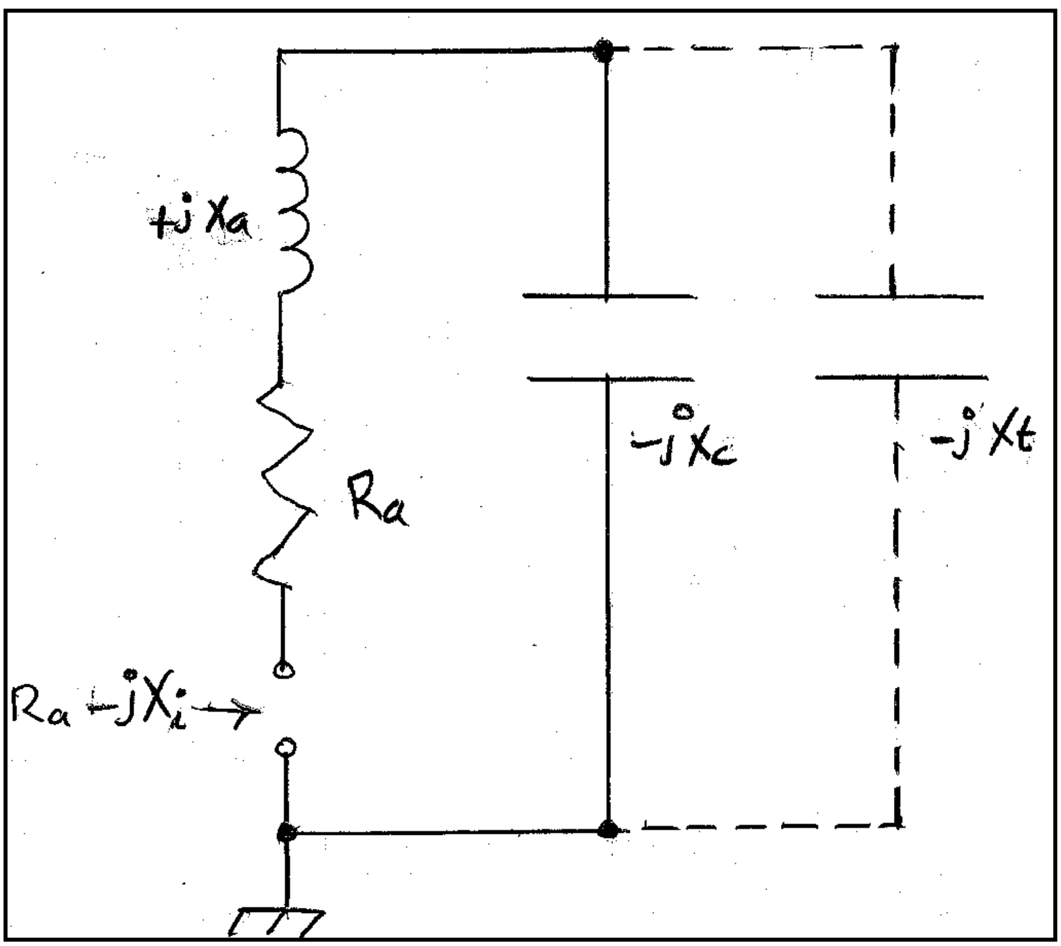

Figure 3.3 - Equivalent circuit for a vertical with capacitive top-loading.

Figure 3.3 shows an equivalent circuit for a vertical with capacitive top-loading. Ra, Xa and Xc represent the vertical, Xt represents the shunt capacitance introduced by the top-hat. Xi and Xt are:

$$\mathbf{Xi} \approx \frac{\mathbf{XcXt}}{\mathbf{Xc+Xt}} \qquad (3.2) \qquad\qquad \mathbf{Ct} = \frac{\mathbf{1}}{\mathbf{(2\pi f)Xt}} \qquad (3.3)$$

Where Ct is the capacitance corresponding to Xt. Xt can be determined from modeling or in simple cases, calculated. Xt is in parallel with Xc reducing Xi. Xa is usually small in the short verticals and can be ignored with only a small effect on the approximations. Figures 3.4 and 3.5 show values for Xi associated with figures 3.1 and 3.2.

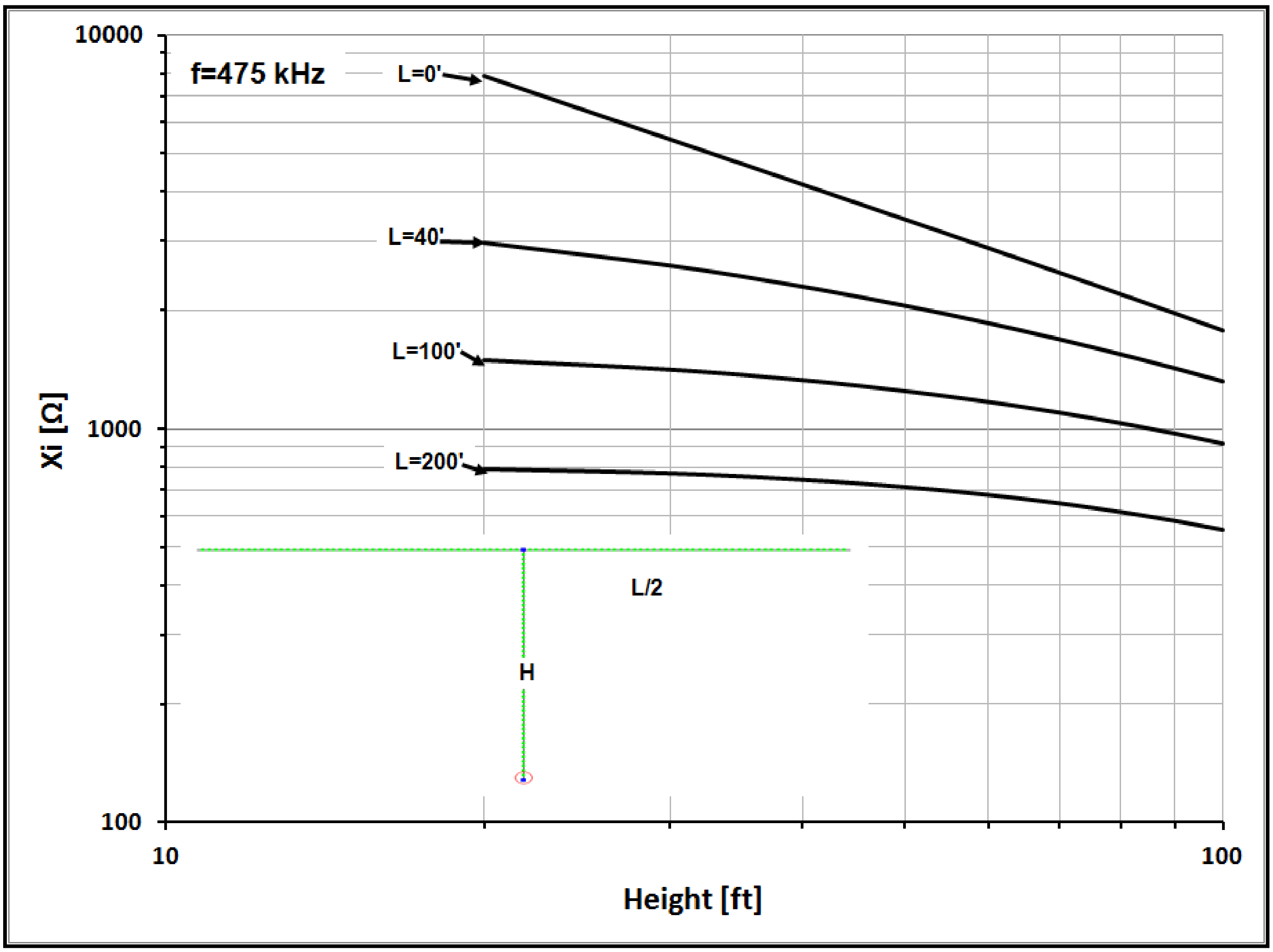

Figure 3.4 -Xi at 475 kHz.

The effect of even a small amount of top-loading on Xi is significant. For example in figure 3.4, at H=20', with no loading Xi≈7900Ω. However, when we add 40' of top-wire

Xi≈3000Ω which is a reduction of almost 3X. Increasing L to 100' reduces Xi by another factor of two, Xi≈1500Ω. RL will be reduced by the same factors which is part of the reason for the efficiency improvement.

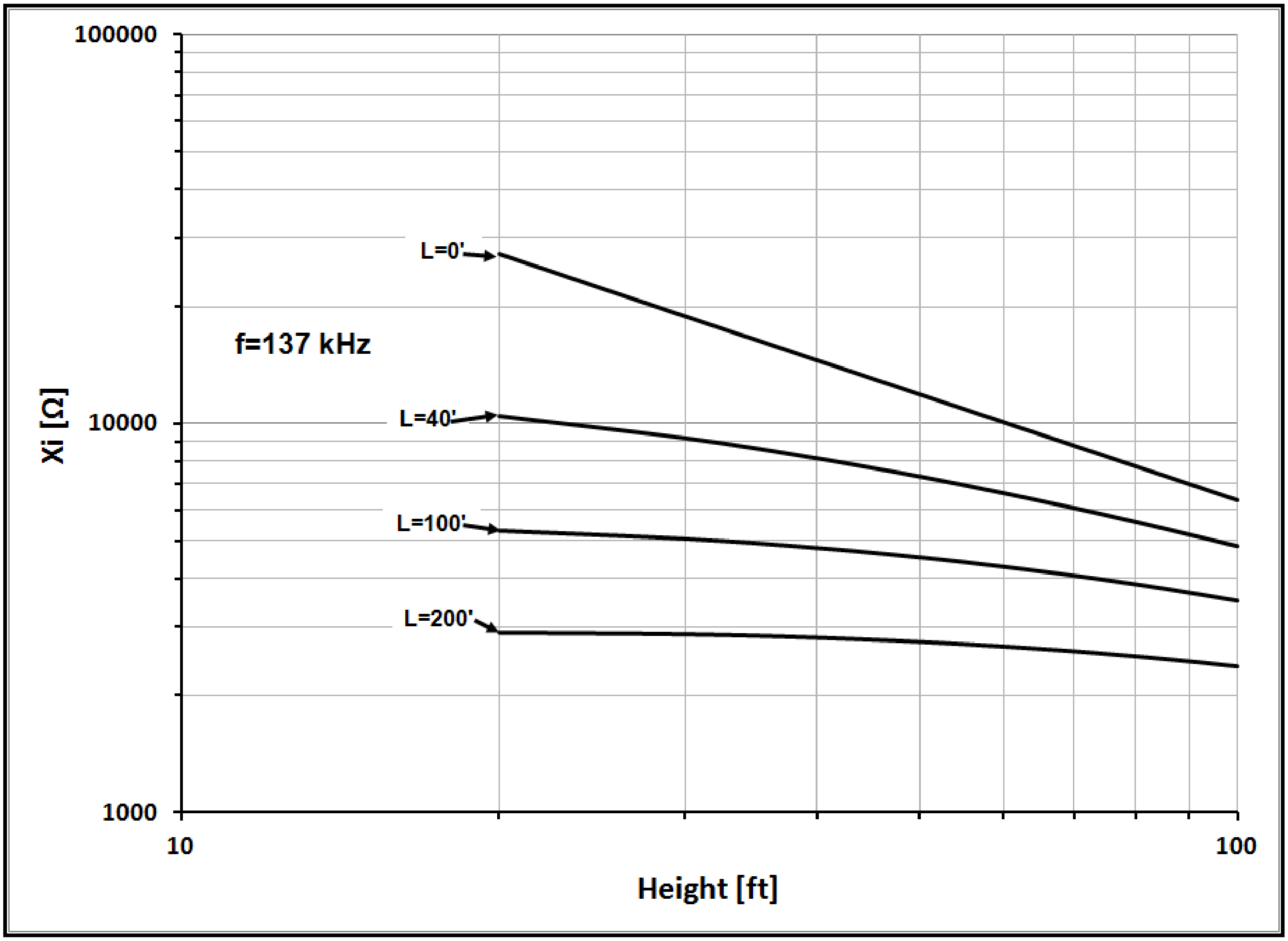

Figure 3.5 - Xi at 137 kHz.

The graphs in figures 3.4 and 3.5 were obtained from NEC modeling. We could also have derived this information with reasonable accuracy from calculations. We can consider each of the horizontal wires connected to the top of the vertical wire to be a single wire transmission line with an characteristic impedance of:

$$\mathbf{Zt = 138Log\left[\frac{4H}{d}\right] \quad [\Omega]} \quad (3.4)$$

Where H is the height above ground of the wire and d is the wire diameter. We are free to chose the units for H and d but both must use the same units.

As an example, a #12 wire (d=0.081" or 0.00675') at 50' will have Zt=617Ω. With a value for Zt in hand we can calculate Xt for each wire from the open circuit transmission line equation (chapter 2, equation (2.9)):

$$\mathbf{Xt} = \left(\frac{\mathbf{1}}{\mathbf{N}}\right)\left(\frac{\mathbf{Zt}}{\mathbf{tan(L')}}\right) \qquad (3.5)$$

When L is the physical length of the top-wire from the top of the vertical to the end, L' is the length in either degrees or radians at the operating frequency and N is the number of wires.

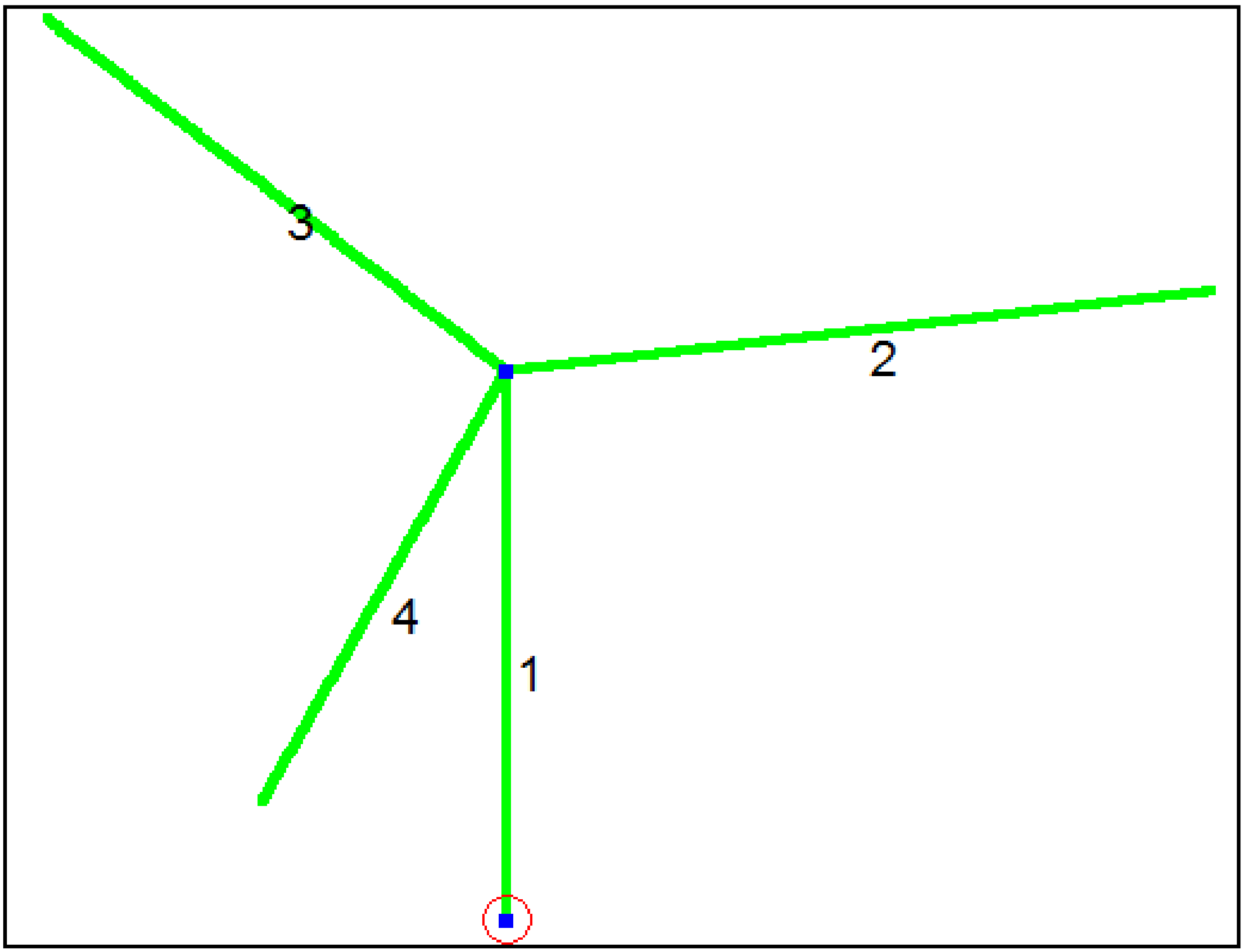

Figure 3.6 - Example for calculation.

We can use this approach to calculate Xt for N top-loading wires as shown in figure 3.6. In this figure N=3 but we will vary N from 0 to 8. If H=50' and L=50'. L' at 475 kHz, where λ≈2072', will be:

$$\mathbf{L}' = \frac{\mathbf{L \cdot 360^\circ}}{\boldsymbol{\lambda}} = \mathbf{8.69^\circ/0.152\ radians} \qquad (3.6)$$

From equation (3.4) Zo=617Ω and from equation (3.5), for each wire:

$$\mathbf{Xt' = 4037\Omega} \qquad (3.7)$$

Using NEC we can determine Xi and Xt for each number of top-wires (N) and compare that to Xt'/N:

Table 1 - Xt comparison.

N	Xi	Xt NEC	Xt calc	error
0	3360	0	0	0.00
1	1788	3822	4037	5.63
2	1239	1963	2019	2.84
3	961	1346	1346	0.02
4	795	1041	1009	3.09
5	685	860	807	6.16
6	607	741	673	9.18
7	549	656	577	12.12
8	504	593	505	14.89

For N<6 the error is <6% which gives a reasonable estimate for the tuning inductor value (XL=Xi). As N is increased the top-wires are closer together and the estimate of Xt is too low (i.e. the calculated value of Ct is larger than it should be). More complex top-hats should be modeled with NEC.

3.3 Rr with top-loading

Top-loading can also improve Rr as shown in figures 3.7 and 3.8. In these figures H is varied from 20' to 100' and L is between 0 and 200'. Rr increases substantially (up to 4X!) as we add more top wire.

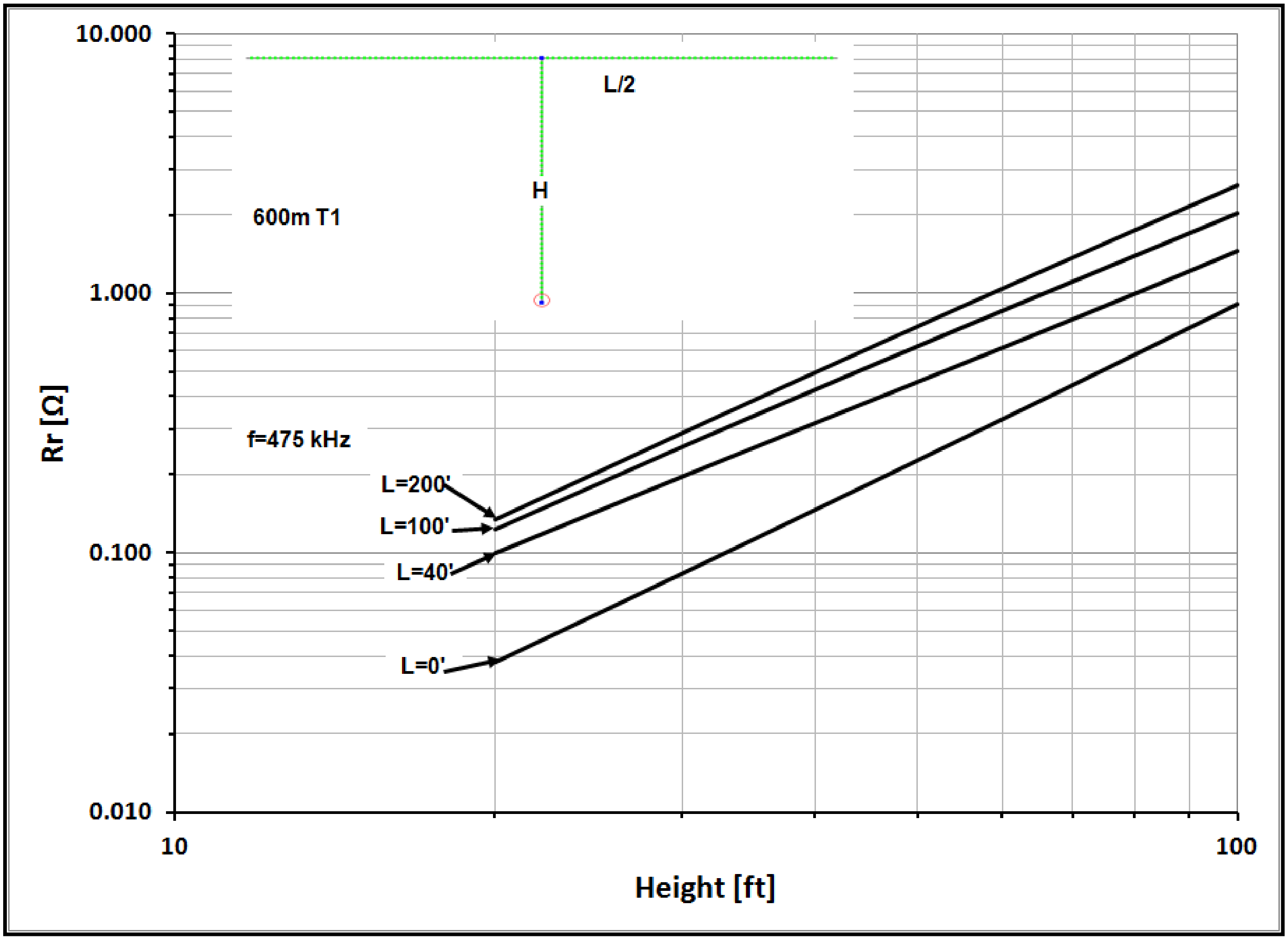

Figure 3.7 - Rr for a T antenna with a single top-wire at 475 kHz.

These Rr graphs are derived from NEC modeling. Rr can be calculated but it's a bit trickier than the unloaded vertical case discussed in chapter 2. The current along the vertical is not a simple triangle, it's trapezoidal and the dimensions of the trapezoid vary with top-loading. Figure 3.9 gives an example of the current distribution along the vertical for several values of L. Io is kept constant at 1A. With no top-loading the current at the top of the vertical (It) is close to zero but as the loading is increased It increases.

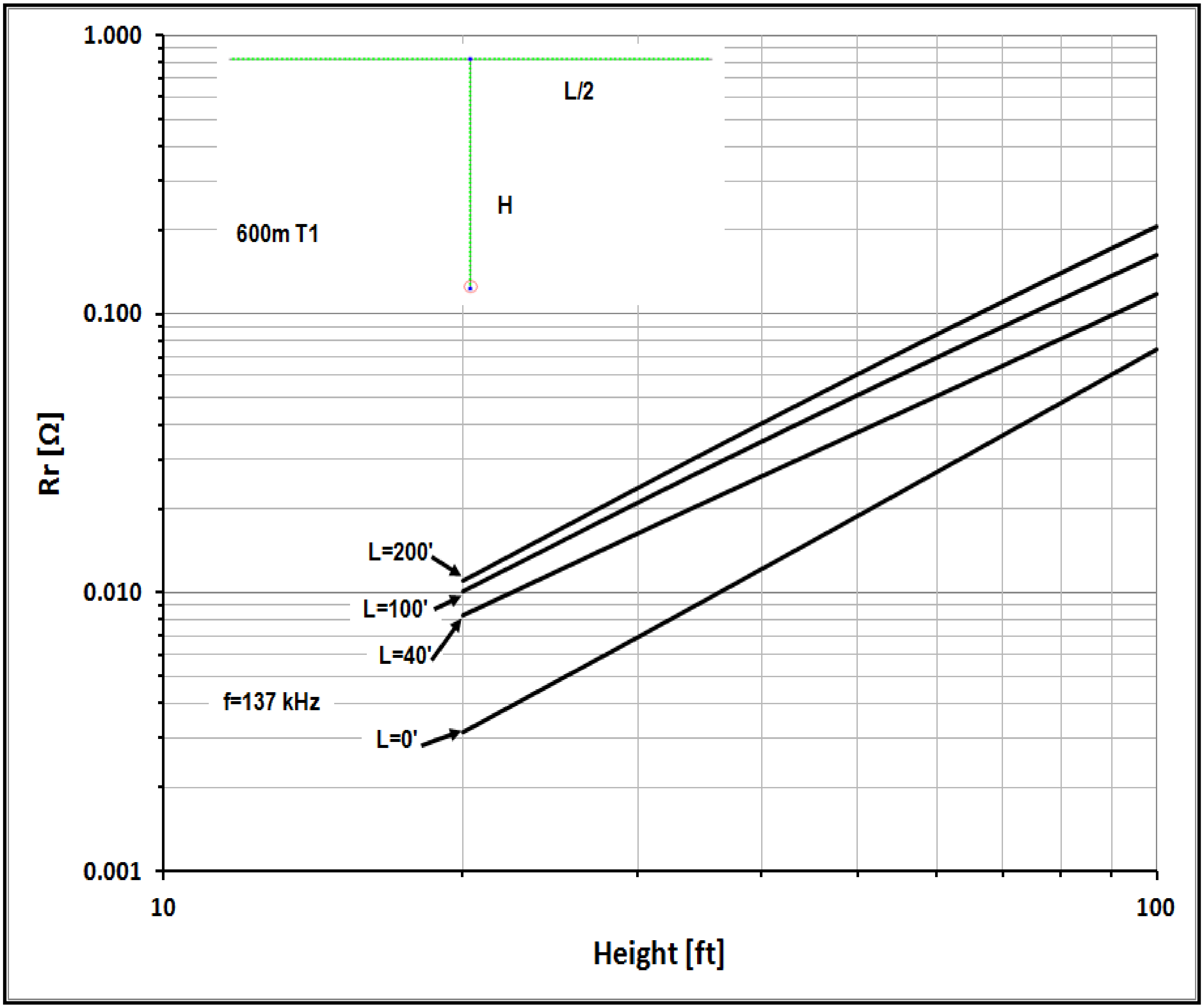

Figure 3.8 - Rr for a T antenna with a single top-wire at 137 kHz.

Equation (2.6) from chapter 2 is still valid:

$$Rr=0.01215A'^2\ [\Omega] \qquad (3.8)$$

But for trapezoidal current distributions equation (2.7) must be modified:

$$A' = \frac{G_v}{2}\left(\frac{It}{Io}+1\right) \quad \text{[Ampere-degrees] (3.9)}$$

Gv=electrical height in degrees.

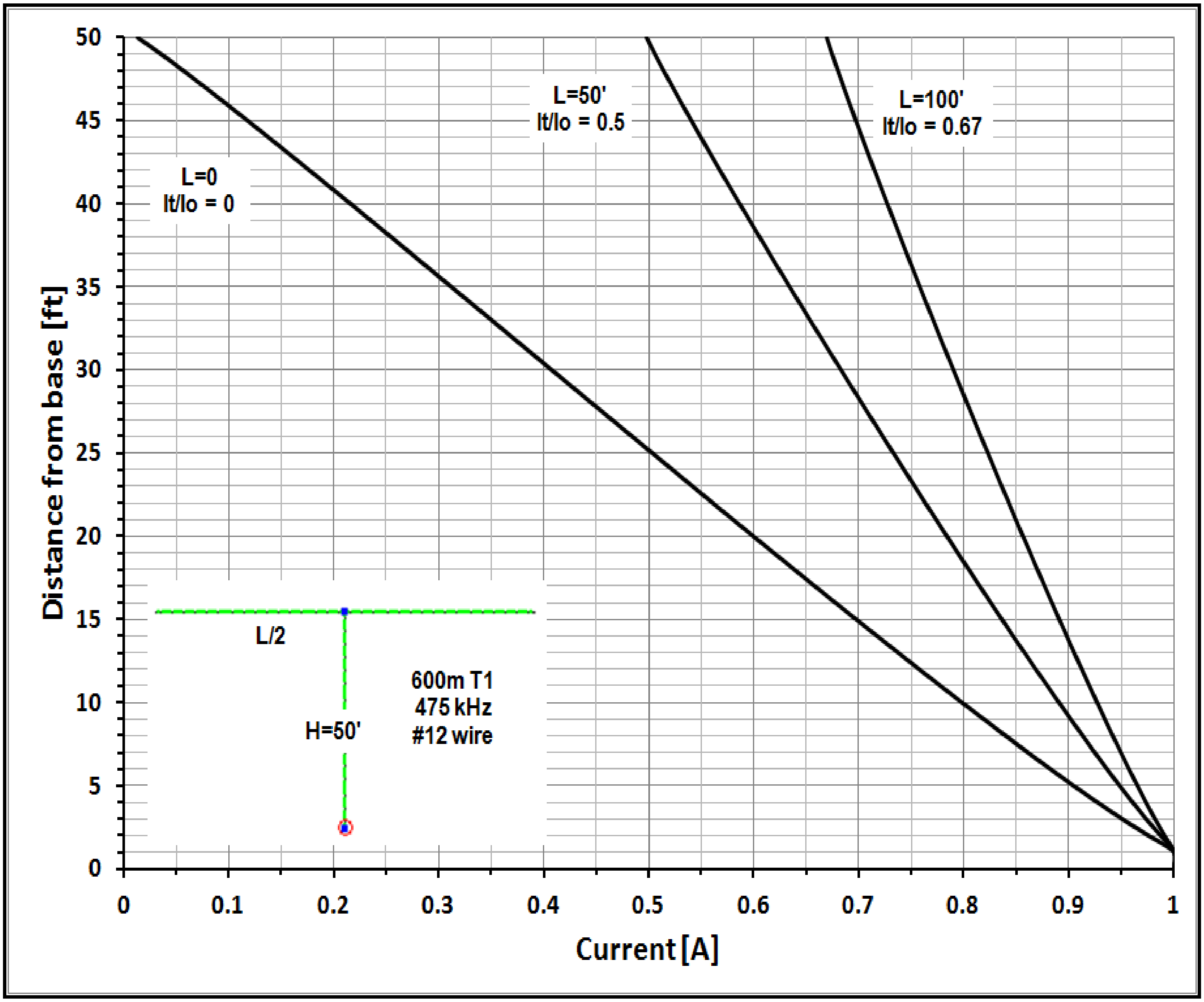

Figure 3.9 - Current distribution on the vertical part of a "T" antenna.

As shown in figure 3.10, inserting equation (3.9) into (3.8) we can create a universal graph for Rr as a function of antenna height in degrees (Gv) with the current ratio It/Io as a parameter. The It/Io ratio is varied from 0 (no top-loading) to 1, which corresponds to heavy top-loading and constant current on the vertical radiator (i.e. It/Io=1). As It/Io goes from zero to 1, Rr increases by a factor of 4X. Equations (3.8) and (3.9) are valid for Gv<45 degrees.

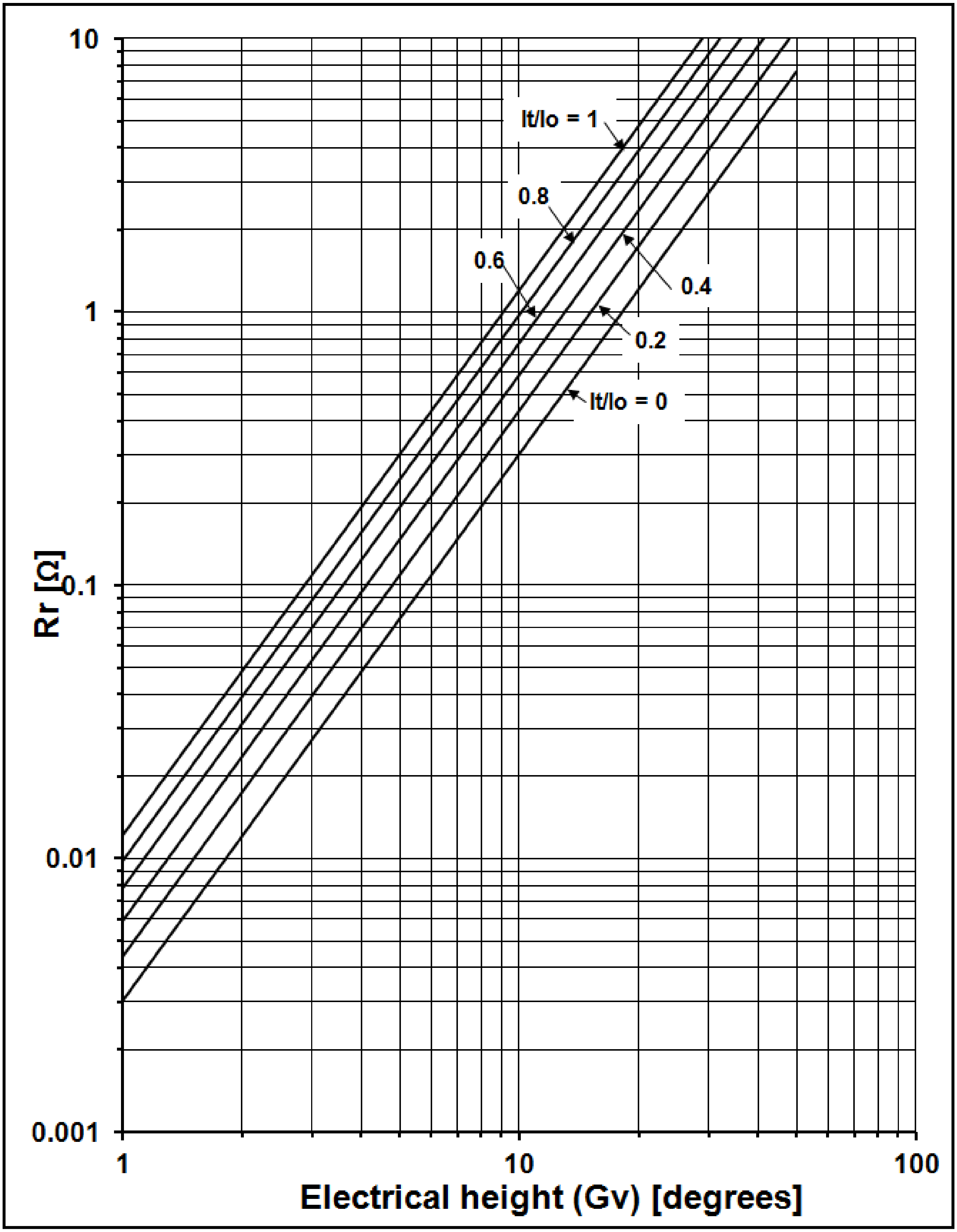

Figure 3.10 - Rr as a function of height in degrees.

For convenience we can convert the graphs in figure 3.10 to height in feet at 137 and 475 kHz as shown in figure 3.11.

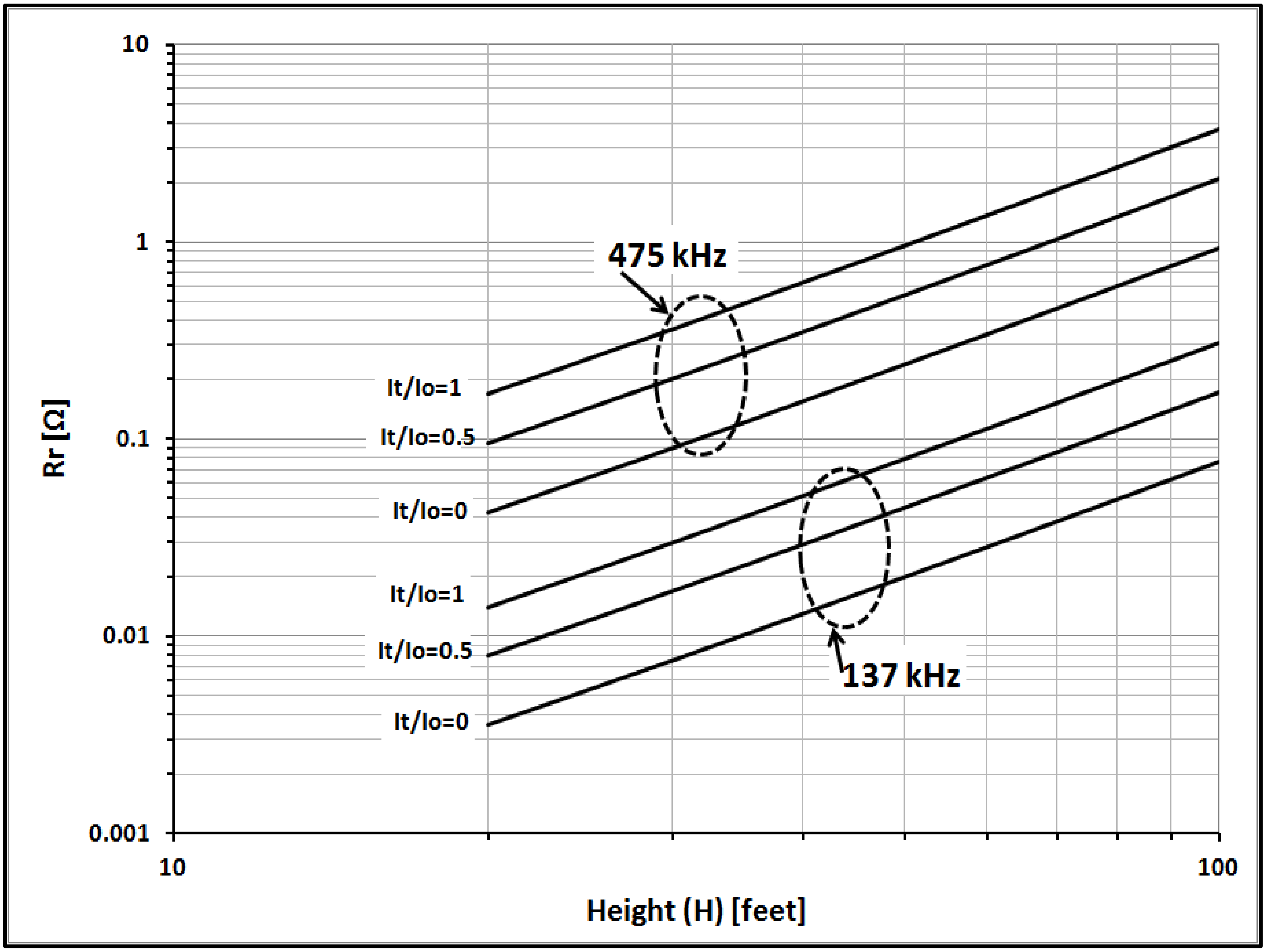

Figure 3.11 - Rr as a function of height in feet.

If we know It/Io then we can simply read Rr off the graph. Finding It/Io with modeling is easy but manually it's a bit tricky. I have never seen a discussion on calculating It/Io so on a hunch I did an experiment with NEC using the antenna shown in figure 3.6 with N varying from 0 to 8, H varying from 20' to 100' and L varying from 0 to 80'. This was done a 137 and 475 kHz. At each data point I recorded Xi. With no top-loading (N=0) Xi=Xc. Using that value for Xc and the value for Xi at each data point, I calculated Xt from:

$$\mathbf{Xt} = \frac{\mathbf{XiXc}}{\mathbf{Xc - Xi}} \quad (3.10)$$

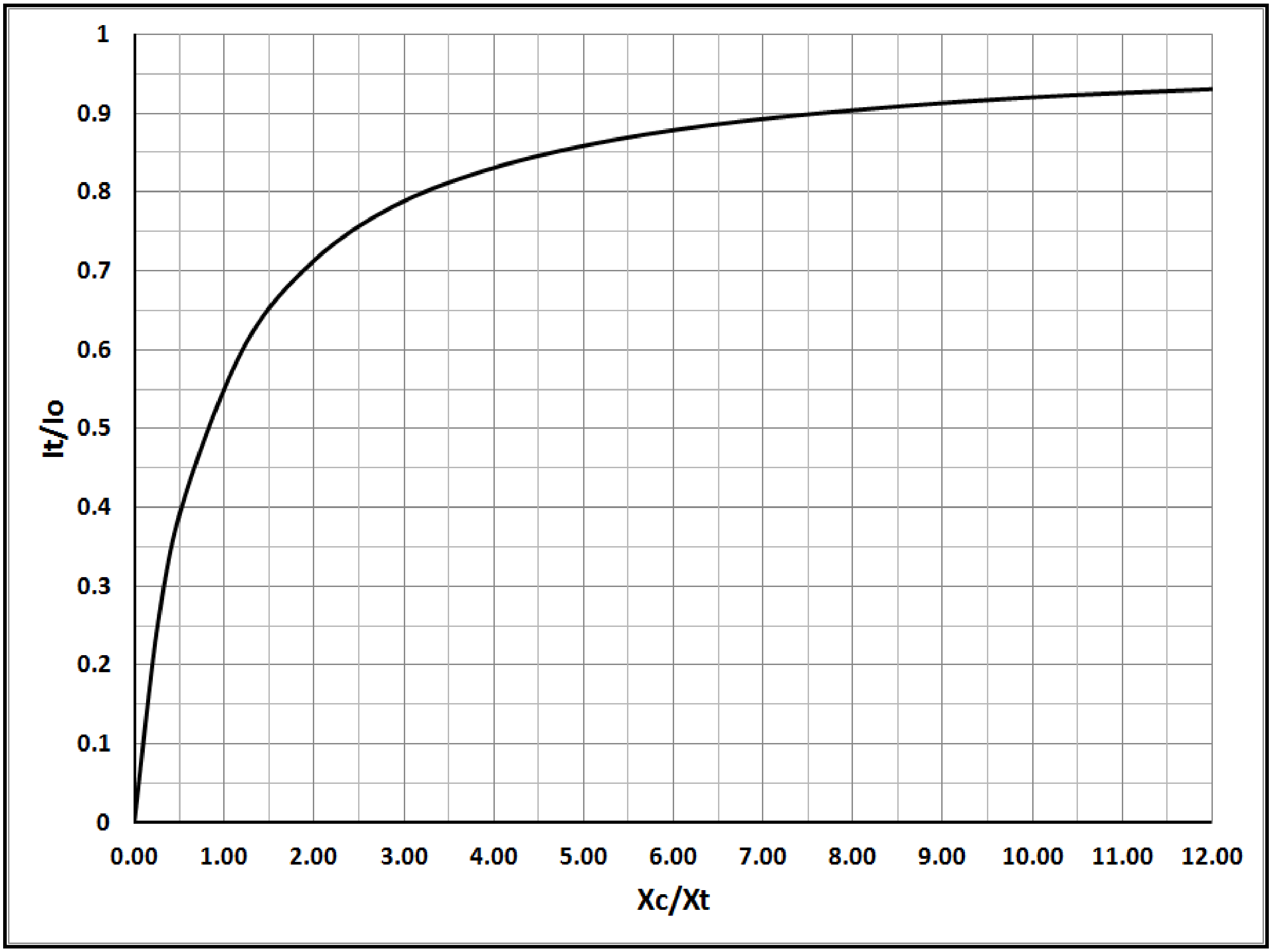

Figure 3.12 - Xc/Xt ratio.

I also recorded the value for It/Io at each point and then graphed **It/Io** versus **Xc/Xt** as shown in figure 3.12. The single line on the graph represents all values of H, L and frequency! This was a complete but very pleasant surprise! The curve is for the antenna with 8 radial top-wires but The individual values for N<8 lie along lower sections of the same curve (i.e. smaller Xc/Xt). The increase in Rr associated with a given value for Xc/Xt is shown in figure 3.13.

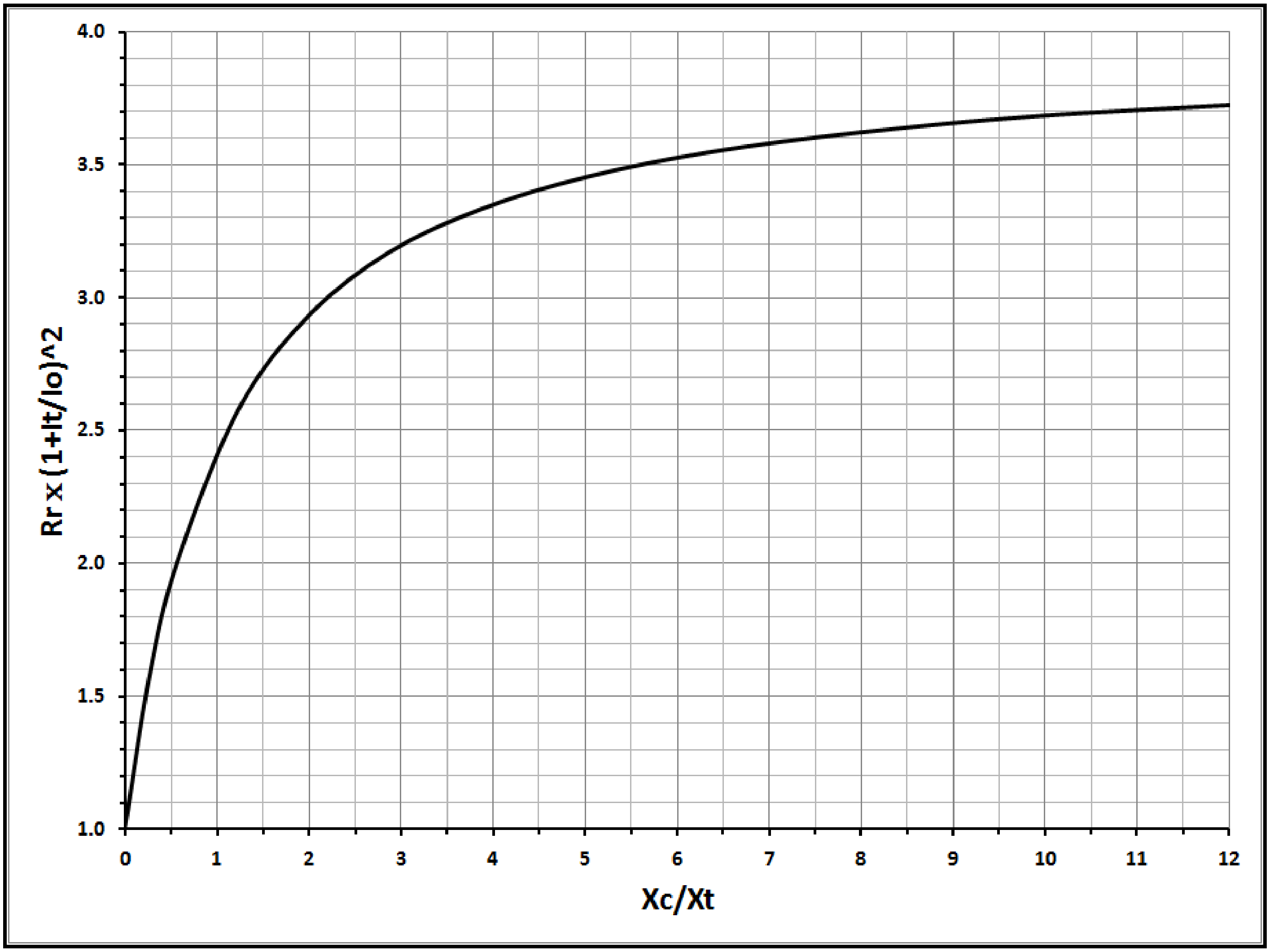

Figure 3.13 - Increase in Rr with top-loading (Xt).

We can calculate values for Xc and Xt using Equations (2.8), (2.9), (3.4) and (3.5), take the ratio and determine It/Io from figure 3.12. With a value for It/Io we can then go to figure 3.10 or 3.11 or use equations (3.8) and (3.9) to get a good approximation for Rr. One thing to notice in figures 3.12 and 3.13 is the flattening of the curves for Xc/Xt>2. Adding more top-loading will initially increase Rr substantially but there is a point of vanishing returns. However, as we increase Ct further (i.e. make Xt smaller) we still get an almost linear reduction in Xi which reduces XL and the associated inductor loss.

The small electrical size of our antennas allows us to simplify the expression for **Xc/Xt**. At 475 kHz 100' is ≈0.05λ, which corresponds to 18º or 0.314 radians. The Tan(0.314)=0.325, with is a difference of only 3%. This difference becomes even smaller for shorter lengths or lower frequencies. We can use this to replace Tan(H') with H in radians (H').

$$\mathbf{Xc} \approx \frac{\mathbf{Za}}{\mathbf{H'}} \quad \text{and} \quad \mathbf{Xt} \approx \left(\frac{\mathbf{1}}{\mathbf{N}}\right)\left(\frac{\mathbf{Zt}}{\mathbf{L'}}\right)$$

$$\frac{\mathbf{Xc}}{\mathbf{Xt}} = \left(\frac{\mathbf{Za}}{\mathbf{Zt}}\right)\left(\frac{\mathbf{L'}}{\mathbf{H'}}\right) = \left(\frac{\mathbf{Za}}{\mathbf{Zt}}\right)\left(\frac{\mathbf{L}}{\mathbf{H}}\right) \quad (3.11)$$

Note! The right side of equation (3.11), the ratio L/H, L and H can be in any units as long as both use the same units which is the same rule for Za and Zt. It is not necessary to convert H and L into radians (H' and L')!

For convenience we can restate:

$$\mathbf{Za} = \mathbf{60}\left[\mathbf{ln}\left(\frac{\mathbf{4H}}{\mathbf{d}}\right) - \mathbf{1}\right] \quad [\Omega] \quad (2.8)$$

$$\mathbf{Zt} = \mathbf{138Log}\left[\frac{\mathbf{4H}}{\mathbf{d}}\right] \quad [\Omega] \quad (3.4)$$

$$\frac{\mathbf{Xc}}{\mathbf{Xt}} = \left(\frac{\mathbf{Za}}{\mathbf{Zt}}\right)\left(\frac{\mathbf{1}}{\mathbf{N}}\right)\left(\frac{\mathbf{L}}{\mathbf{H}}\right) \quad (3.11)$$

Just remember to use the same units throughout!

Equations (2.8), (3.4) and (3.11) all include approximations so we need to check the effect of these on the computed value for Xc/Xt. We can do this by comparing the values derived from NEC modeling to those derived from the equations. figures 3.14 and 3.15 make that comparison using the figure 3.6 antenna with H=20' and 80' and the number of top wires (N) set to 2 and 8. In table 1 we saw that the error in Xt was small for N<7 and we see the same behavior in figures 3.14 and 3.15. For small values of N the agreement is quite good. For larger values of N Equation (3.11) over estimates Xc/Xt but for values of Xc/Xt>3 the It/Io curve flattens so the error in Rr prediction is not all that great. This illustrates the point that manual calculations work fine for simple top-loading arrangements but modeling is really best way for more complicated arrangements.

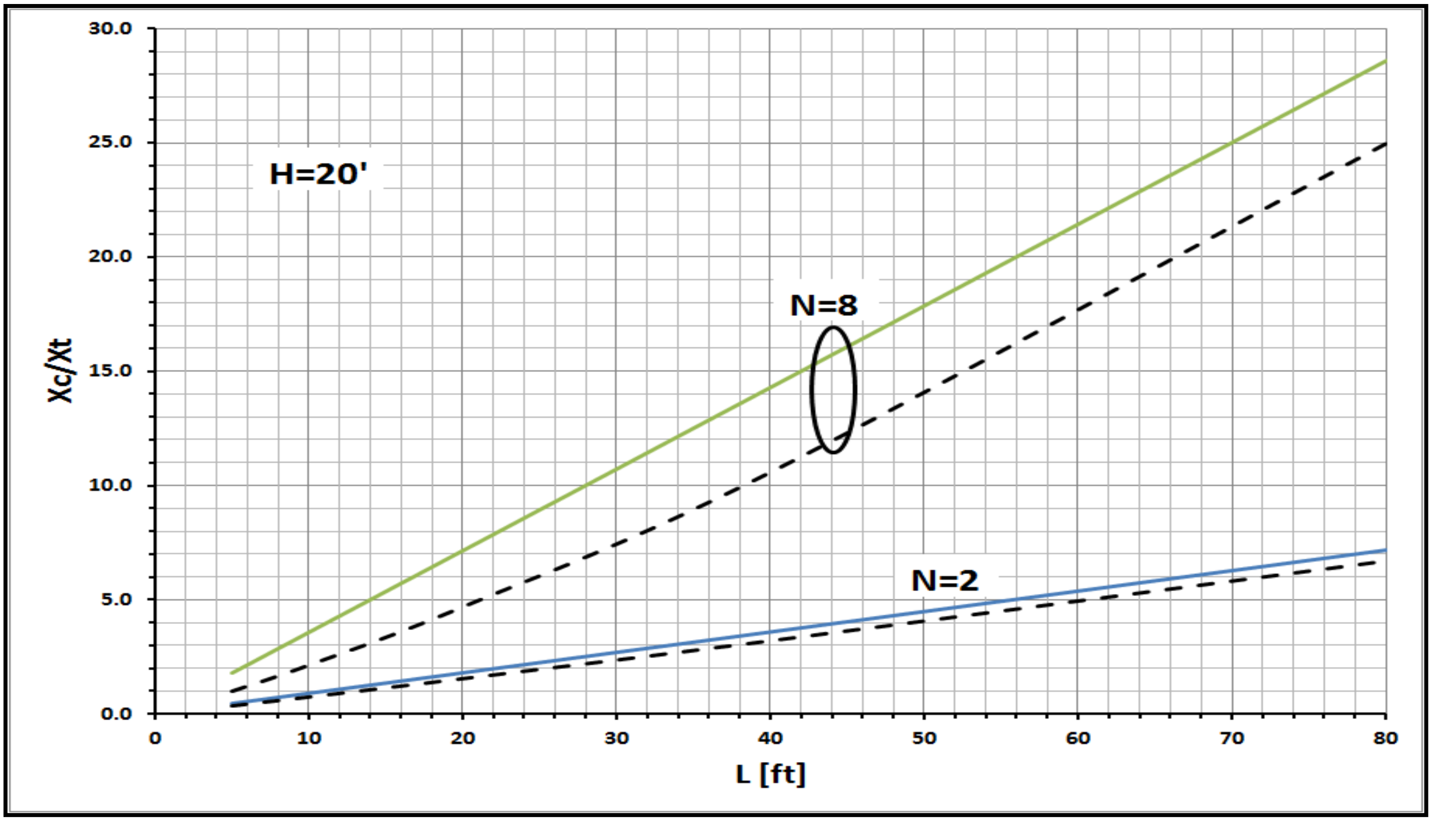

Figure 3.14 - Comparison of Xc/Xt derived from NEC and computation for H=20'.

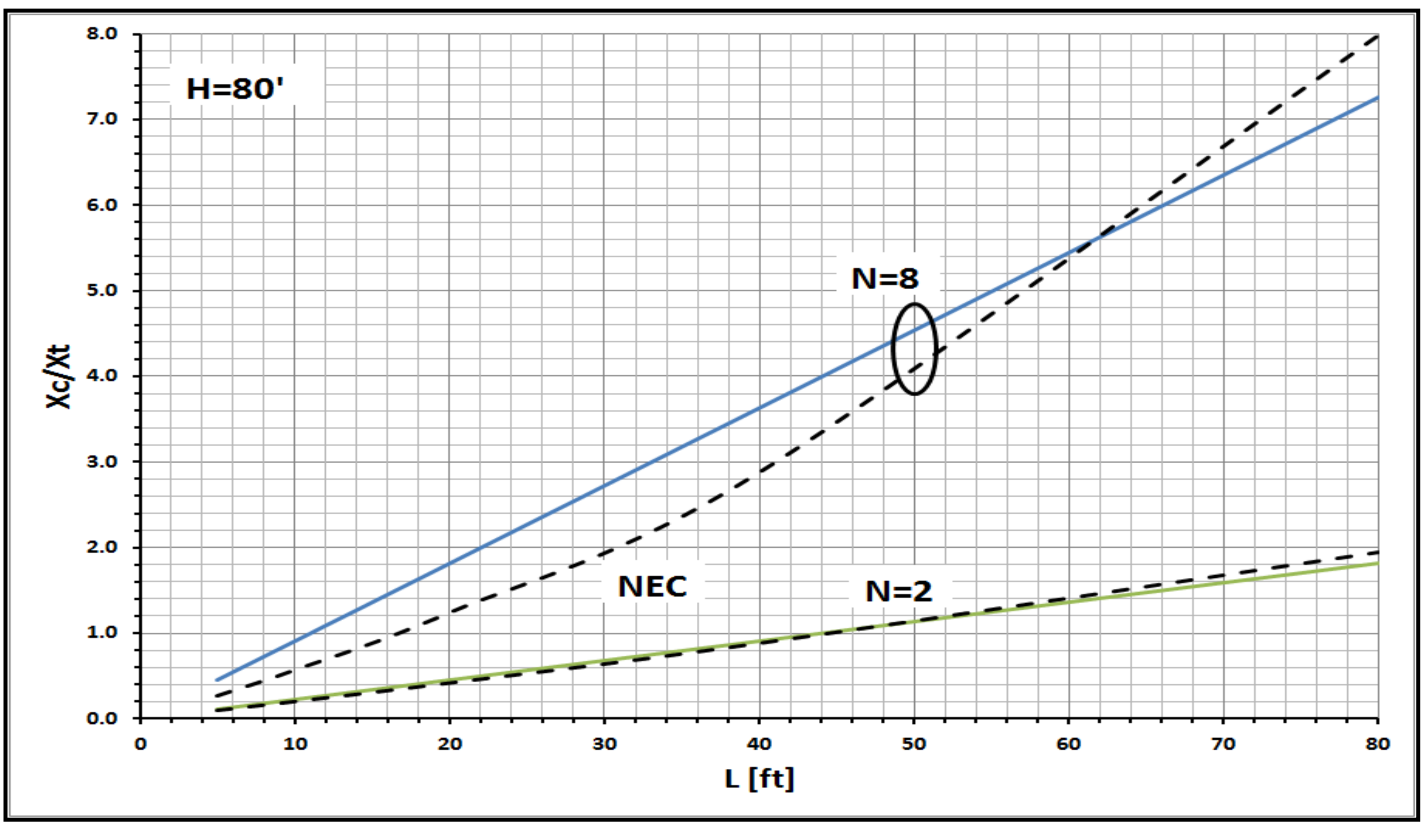

figure 3.15 - Comparison of Xc/Xt derived from NEC and computation for H=80'.

3.4 More realistic antennas

In the real world we won't have a perfectly symmetric T. The realities of a given QTH force us to fit within the available space and supports. This section explores the effect on efficiency of deviations from ideal.

Whenever you stretch a wire between two supports you must have at least some sag to limit wire tension in regions subject to ice loading! Another problem arises when a wire is suspended between trees. Trees move in the wind and two trees 50'-100' apart can be oscillating in opposite directions at the same moment. Both of these considerations can require significant sag in a top-wire. Figures 3.16 and 3.17 illustrate the effect of sag on efficiency. In this example the spacing of the support is 100' with the downlead attached at the center. The ends of the top-wire are fixed at 50' while the height of the center is varied from 25' to 50'. Certainly 25' of sag is excessive but 5' (H=45') is not. With 5' of sag in the 475 kHz antenna the efficiency drops by ≈1.5%. At 137 kHz the efficiency drops from 0.23% to 0.20%. Clearly we want to use as little sag as possible while still meeting the mechanical requirements.

In some installations it may be more convenient to attach the downlead at a point other than the center of the top-wire. As shown in figures 3.18 and 3.19, we can attach the downlead anywhere along the wire and we are also free to place the ground end of the downlead pretty much where we want with little effect on efficiency. Note that in these two figures the efficiency scale is expanded which tends to magnify the differences which are actually rather small.

In some cases the top-wire may not be directly over the point on the ground where we would like connect the downlead. As figures 3.20 and 3.21 show, we can move the downlead ground point many feet off the side with almost no effect on efficiency.

Using supports already on hand (trees, poles, structures), the two ends of the top-wire are likely to be at different heights. Figures 3.22 and 3.23 illustrate the effect top-wire ends at different heights. The horizontal axis on the graphs shows the end height offset from the center height, i.e. for example if one end is 10' higher than the center, the other end will be 10' lower than the center. This means the top-wire slopes downward from one end to the other. As can be seen in the graphs, for a given center height (50' in this example), when we raise one end the efficiency goes up even though we've lowered the other end.

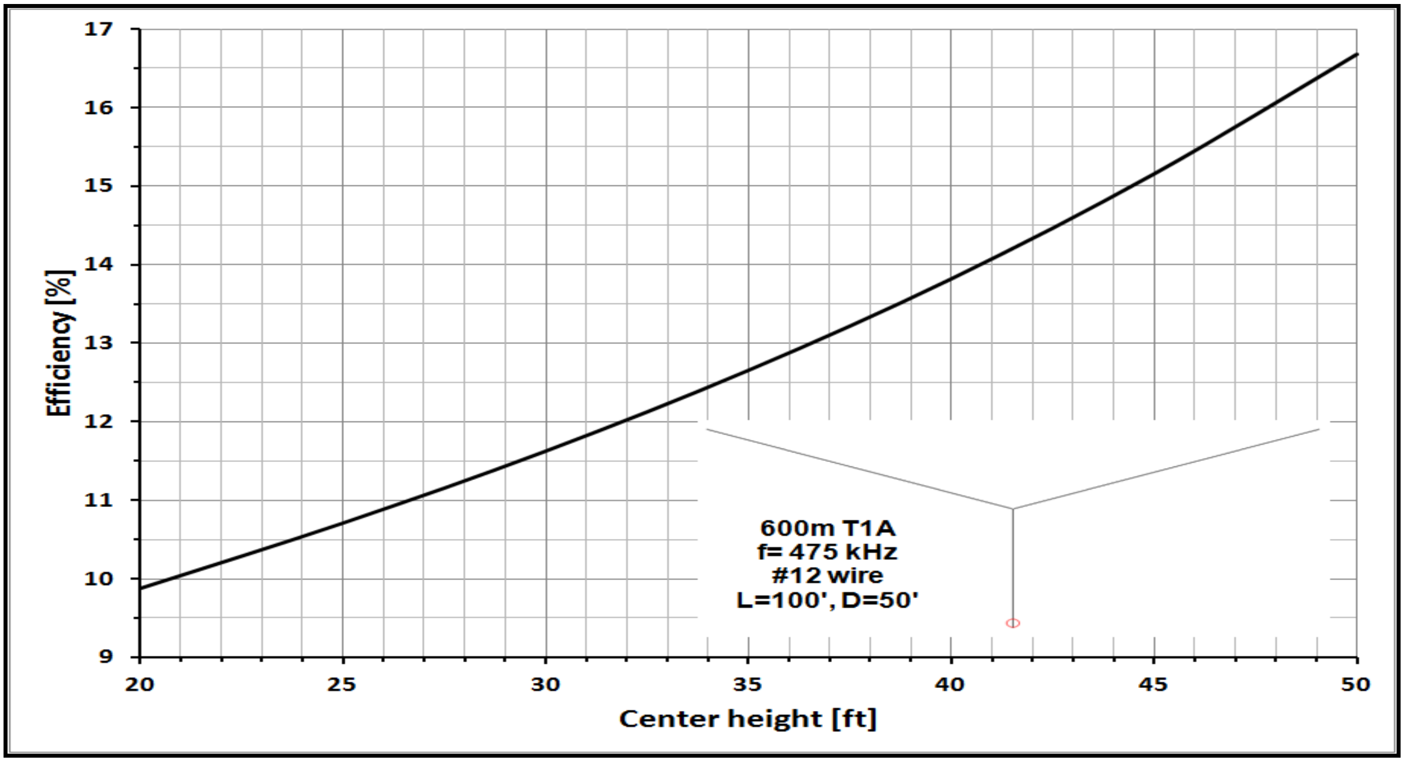

Figure 3.16 - Effect of sag on efficiency at 475 kHz.

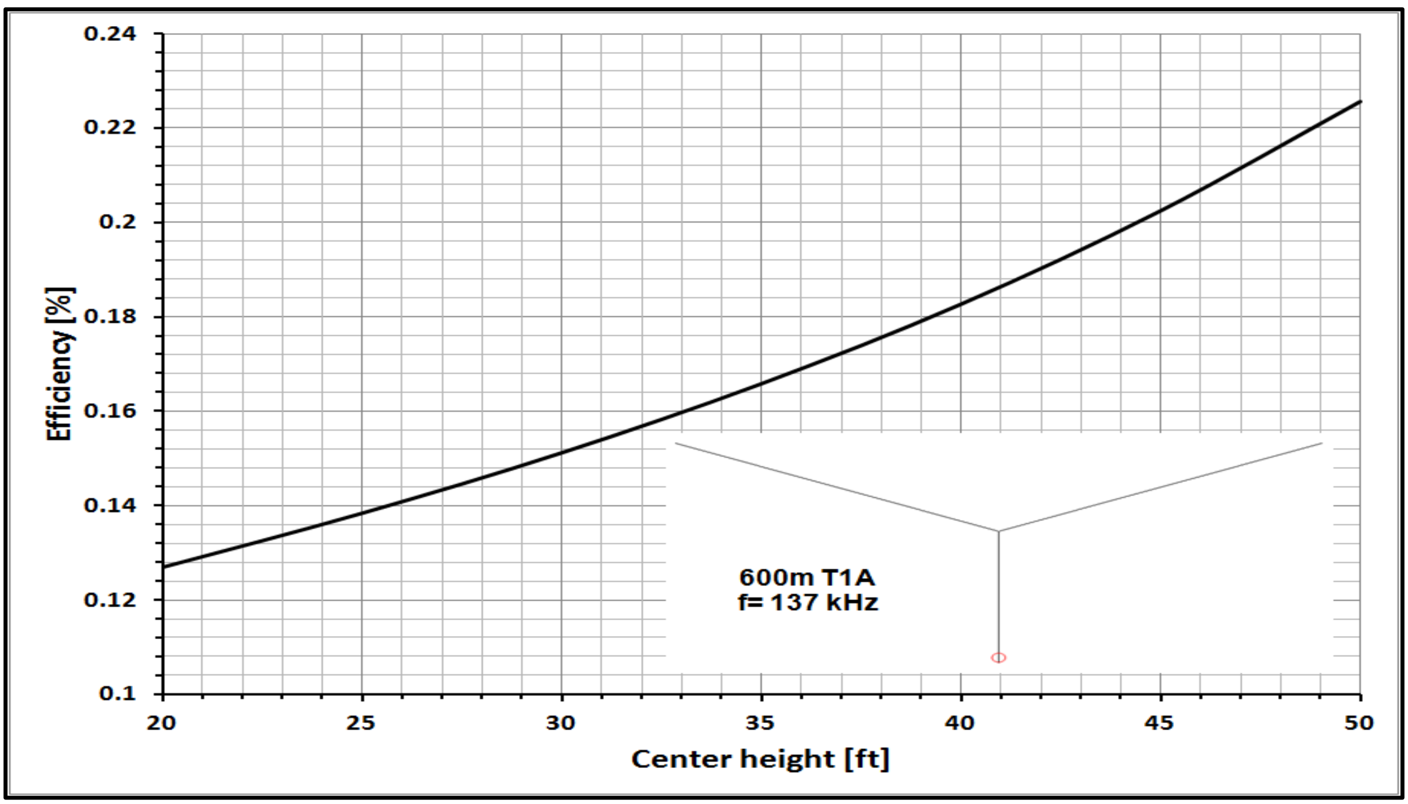

Figure 3.17 - Effect of sag on efficiency at 137 kHz.

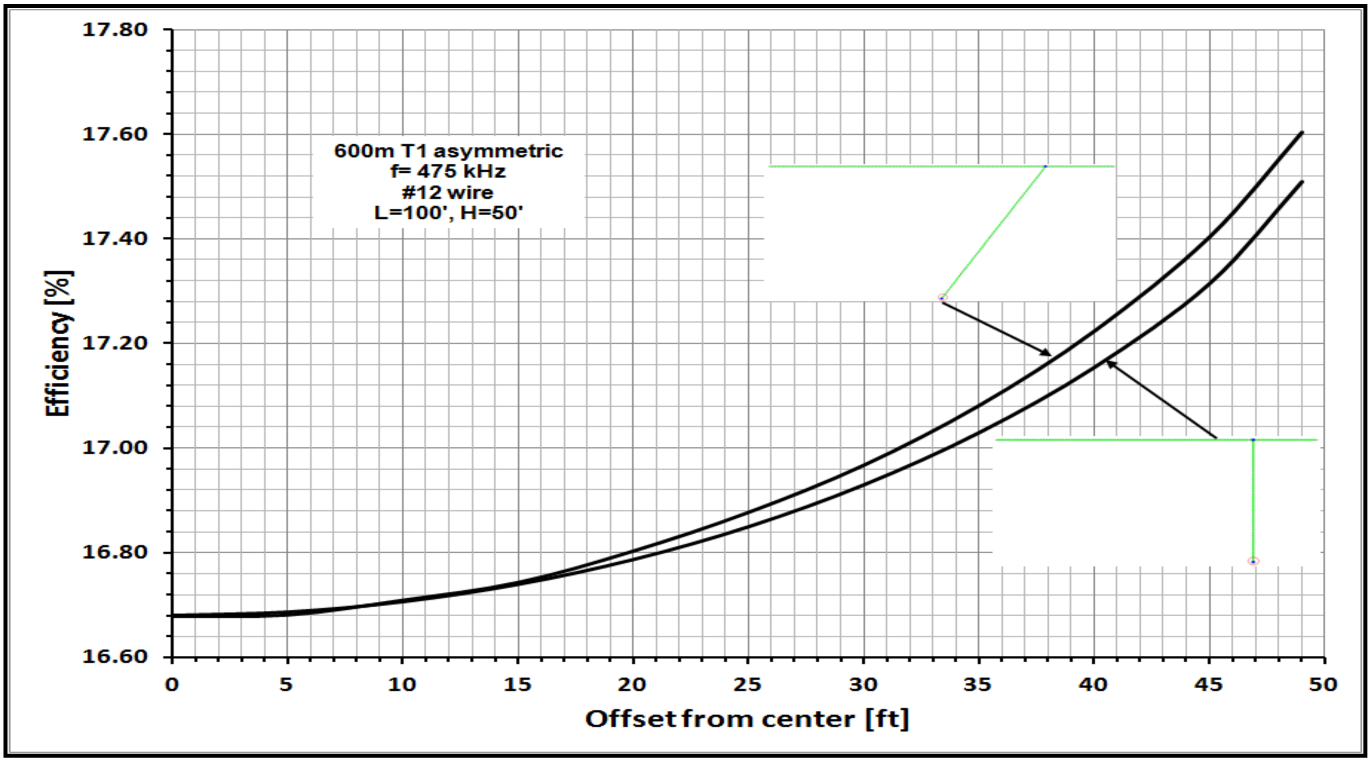

Figure 3.18- Effect of downlead attachment point, 475 kHz.

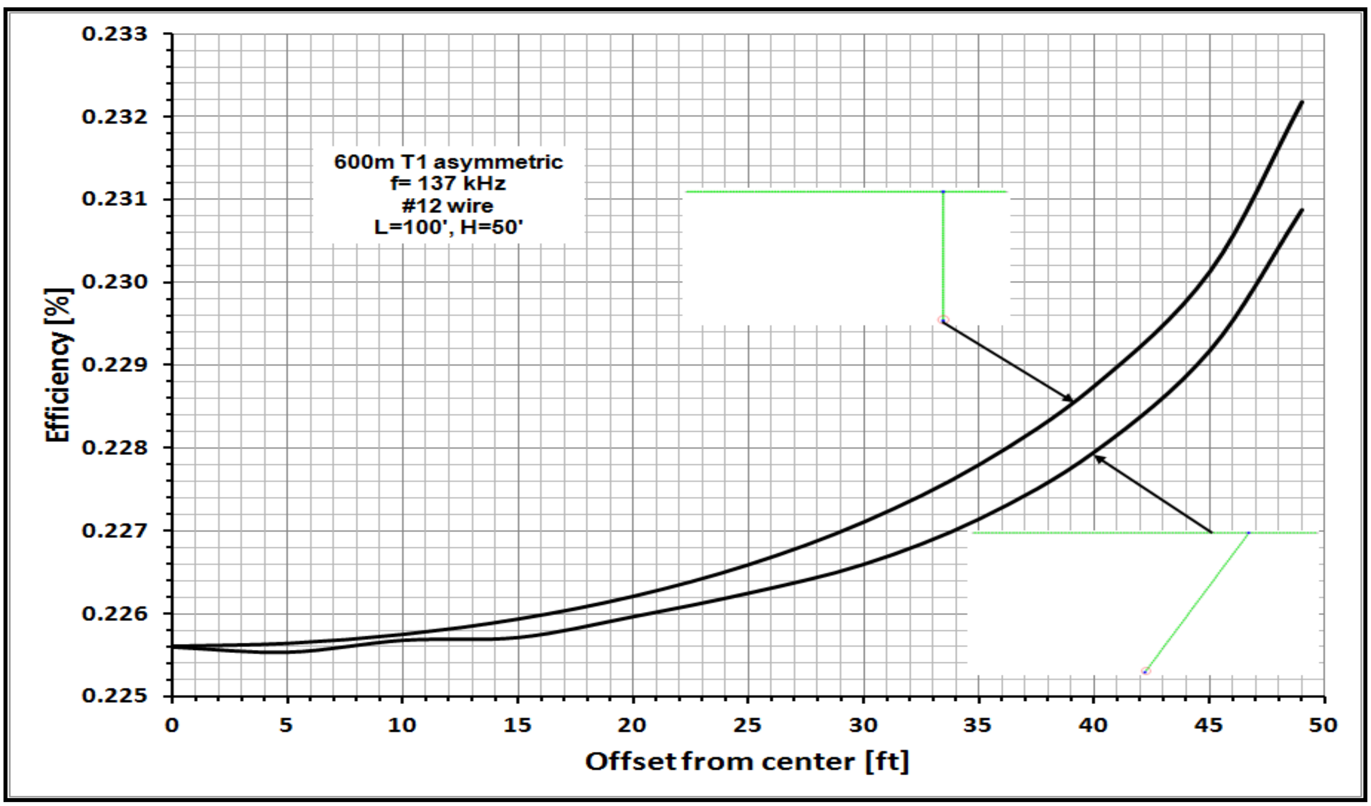

Figure 3.19 - Effect of downlead attachment point, 137 kHz.

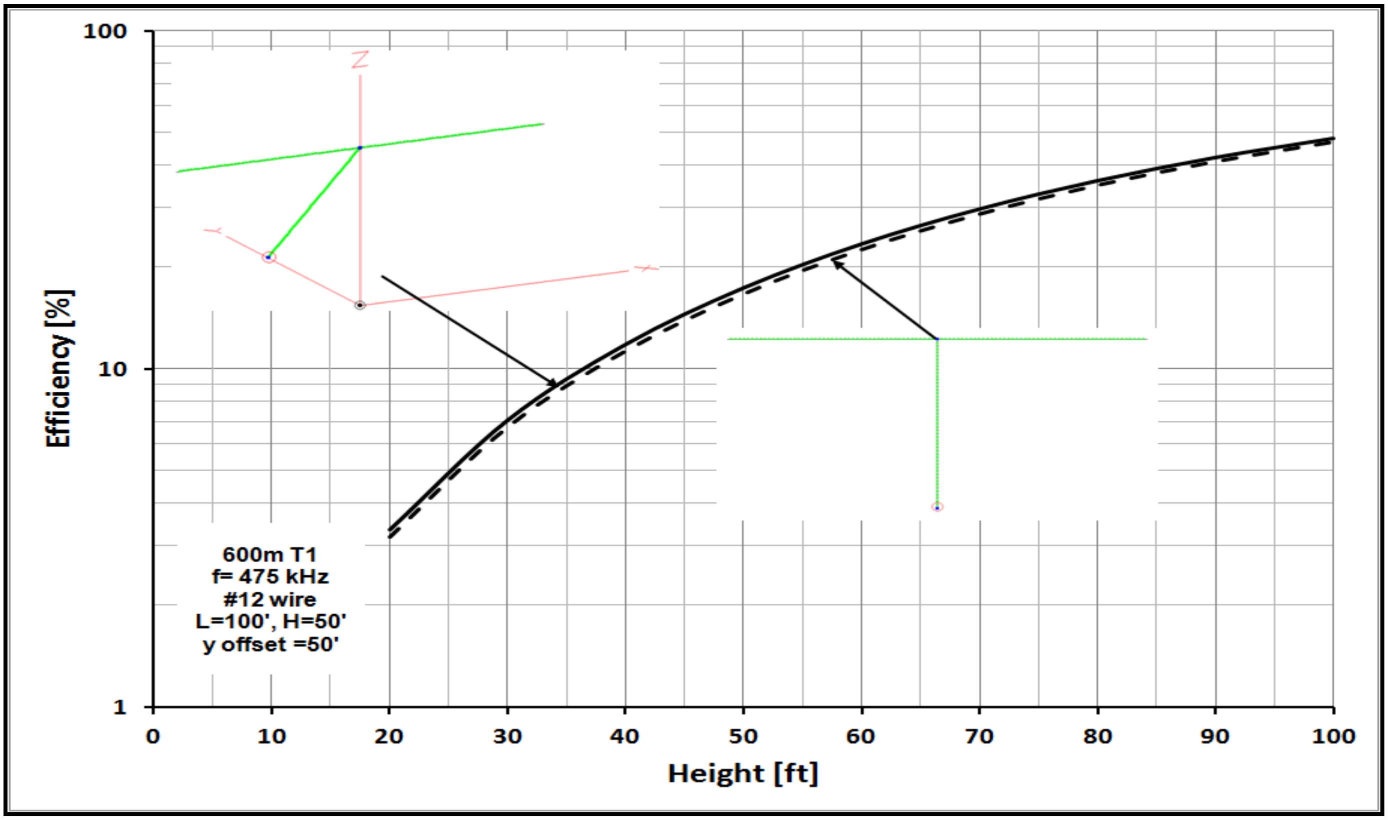

Figure 3.20 - Effect of downlead ground end offset, 475 kHz.

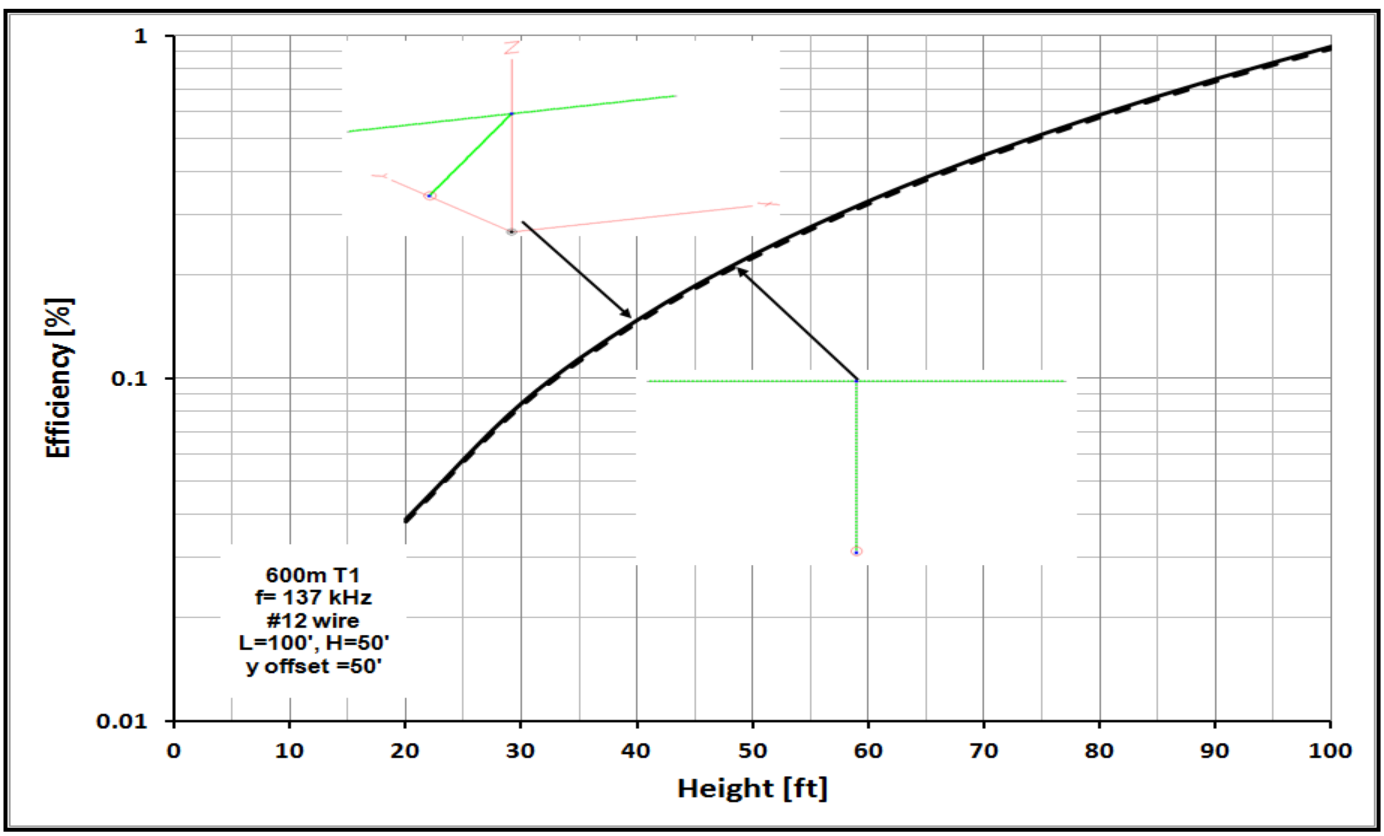

Figure 3.21 - Effect of downlead ground end offset, 137 kHz.

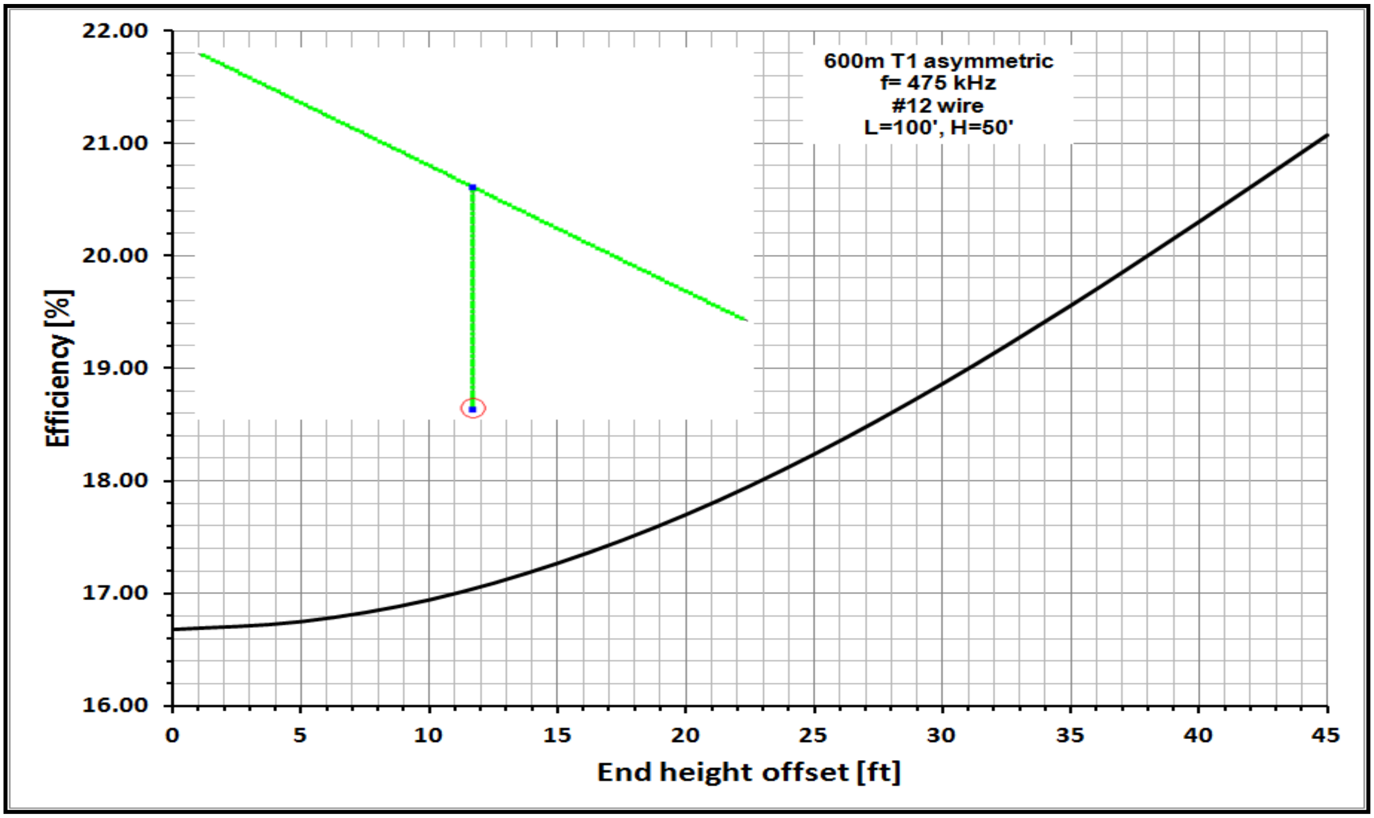

Figure 3.22 -Effect of slope in the top-wire, 475 kHz.

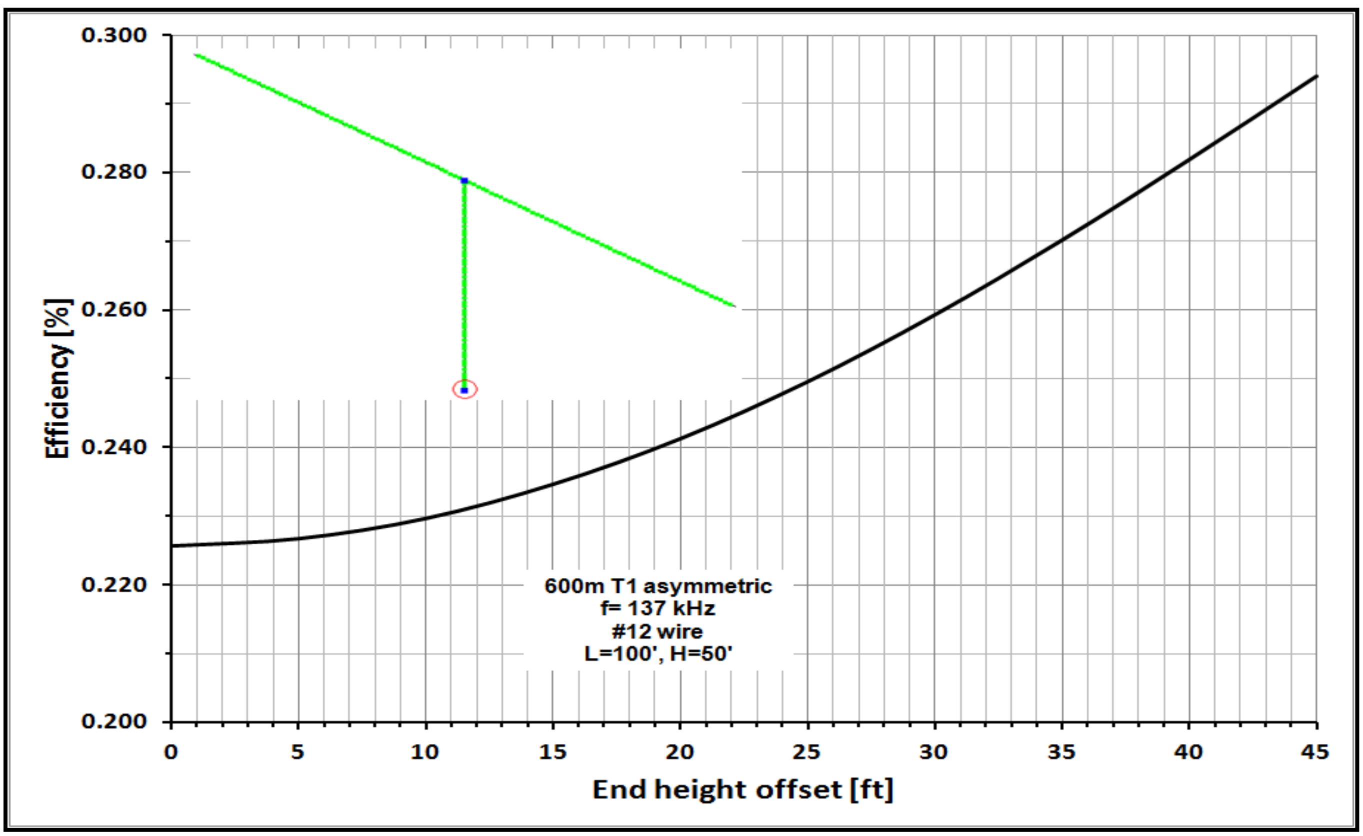

Figure 3.23 - Effect of slope in the top-wire, 137 kHz.

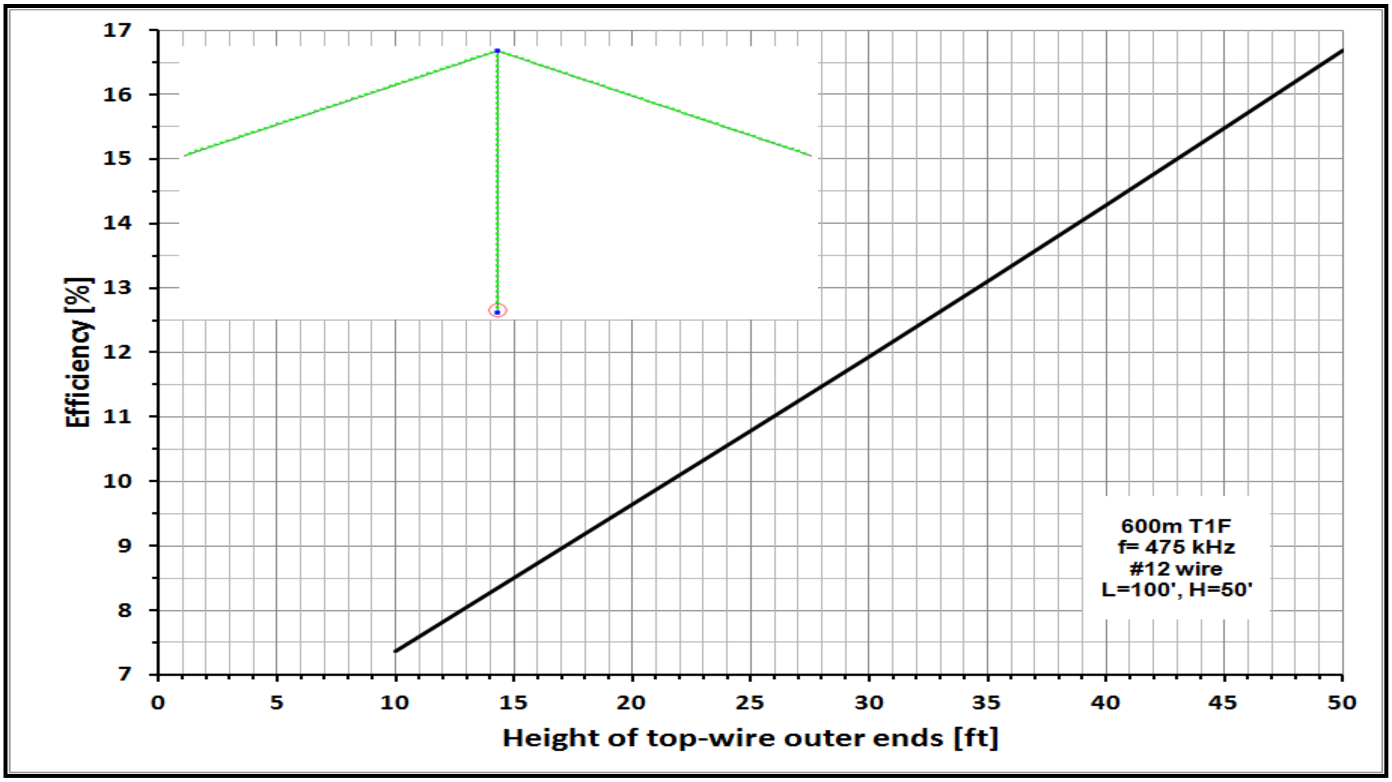

Figure 3.24 -Single support with sloping top-wires, 475 kHz.

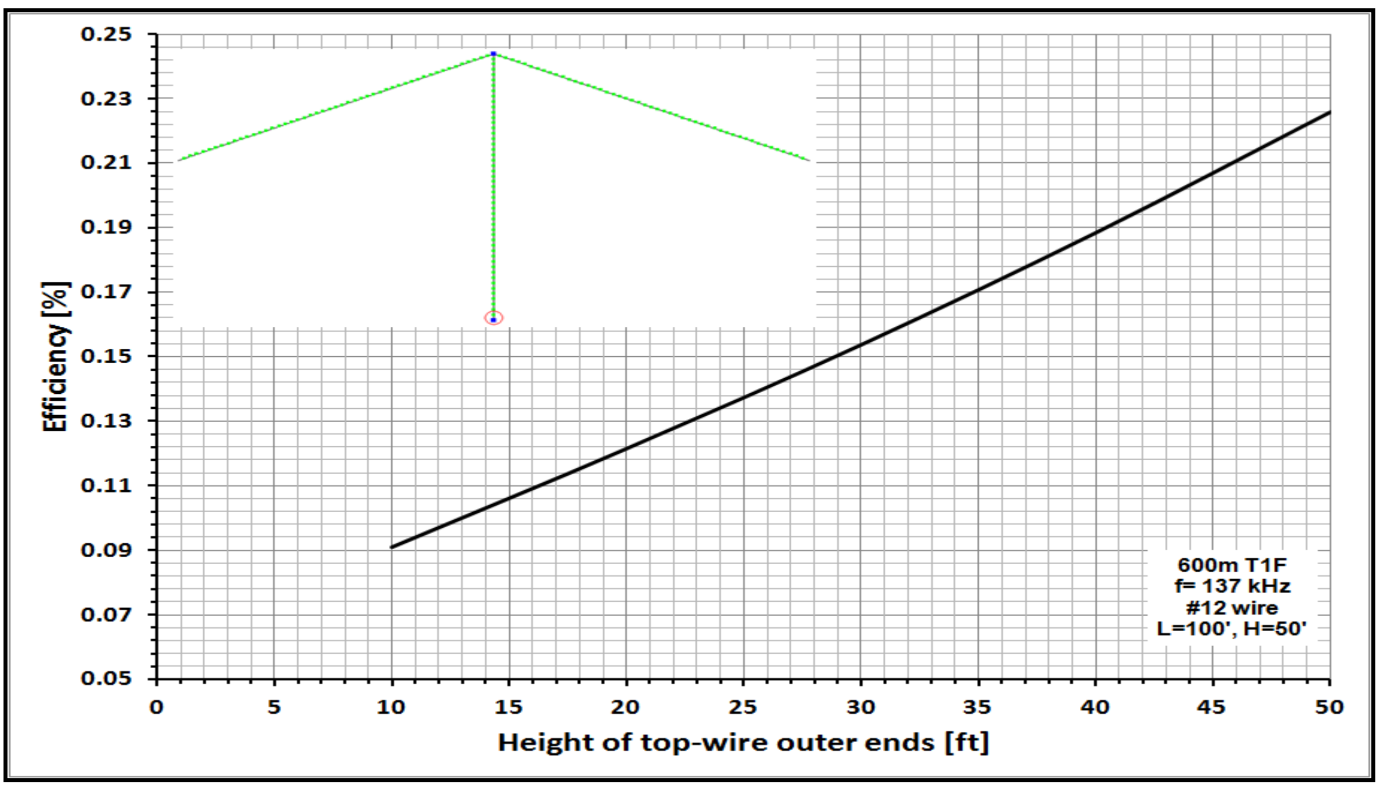

Figure 3.25 - Single support with sloping top-wires, 137 kHz.

Sometimes only one support will be available and the top-wires will have to slope downward from the center supports as indicated in figures 3.24 and 3.25. In this example the height of the vertical is assumed to be 50' and the two top-wires are 50' long. We see that the efficiency is a strong function of the height of the top-wire ends. The higher the better! This is a case where the current in the top-wires has a component that partially cancels the current in the vertical, reducing Rr.

In most cases the available spacing between the supports will be limited. For example, suppose L is limited to 100' and H=50', we can still improve things a bit by adding drop-wires to the ends of the top-wire as shown in figure 3.26 which also shows how the efficiency changes as we vary the drop-wire lengths. When there are no drop wires the efficiency is about 17% but when the drop-wires are 25-30' long (roughly H/2) the efficiency peaks about 3.6% higher. The current in the drop-wires is ≈180° out of phase with the current in the vertical so there is some cancelation which reduces Rr while reducing RL. Initially, as we make the drop wires longer Rr falls somewhat but Xc falls more rapidly until a peak in efficiency is reached. Note however, that the length of the end-loading wires is not very critical.

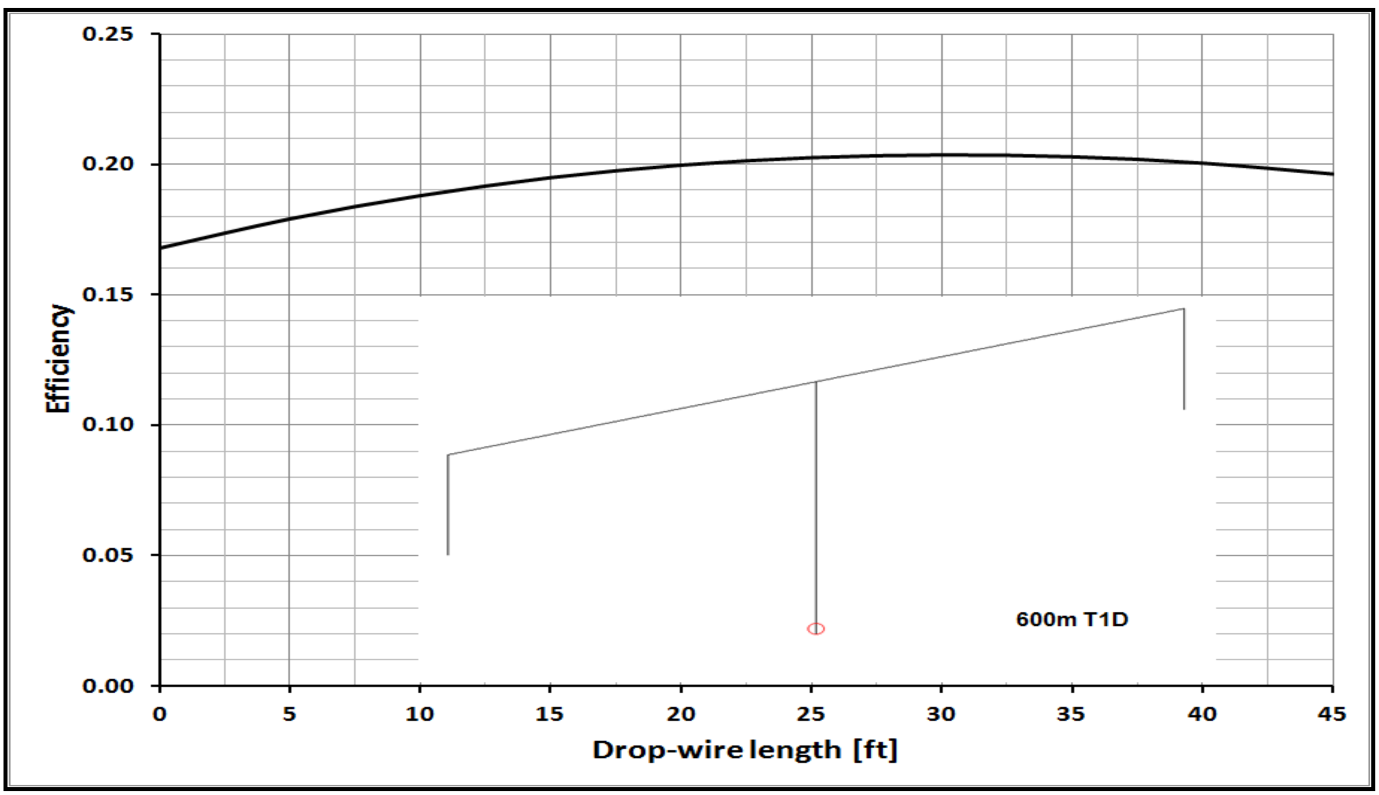

Figure 3.26 - T vertical with end-loading wires. L=100', H=50'.

We don't have to use a straight wire for top-loading, it may be bent as shown in figure 3.27. When the top-wire center angle is 180° the loading reactance for resonance is 1246Ω and the efficiency is about 16.7% (including only the inductor loss!). When the center angle is changed to 90°, the loading reactance increases only slightly to 1255Ω and the efficiency decreases to 16.6%, a very small change. It appears that the center angle for the top-wire is not at all critical and has to made quite small (<45°) before it matters very much.

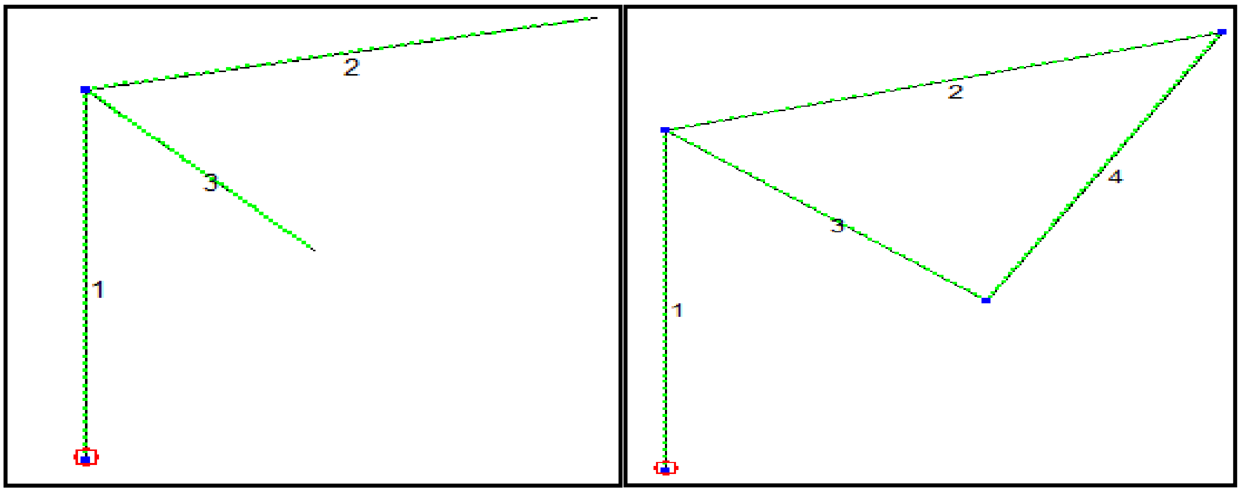

Figure 3.27 - 90° between top-loading wires. With and w/o skirt wire.

Note that when the top-wire is bent it may be possible to add a "skirt" wire as shown on the right (wire 4). For a 90° center angle adding the skirt wire reduces the loading reactance to 904Ω and the efficiency increases to 23.7%. Adding the extra wire is well worth doing! It should be pointed out that wire 1, the vertical, does not have to come straight down to ground. As shown earlier the wire can be slanted towards a more convenient point. In figure 3.27 wires 2, 3 and 4 constitute a loop "top-hat". The downlead could be connected at <u>any</u> point on the loop with only modest effect on efficiency! To generalize a bit further, if multiple support points, at different heights, are available a loop can be strung between these points to form a top-hat. An irregular top-hat works just fine. <u>Exploit the opportunities at your QTH!</u>

At HF we would expect the radiation pattern for the inverted-L to have significant asymmetry. However, the antennas in figures 3.15 through 3.27 are nowhere near self resonance without a loading inductor. The far-field patterns show very little difference between the antennas, i.e. circular to a fraction of a dB. The message is that these antennas are very tolerant of mechanical variations.

3.5 Using a Tower for support

A tower can be used as a support or as a radiator. Exciting the tower will be discussed in chapter 4. The immediate question is: what is the effect of coupling between a grounded tower and the vertical downlead? The simple answer is that it will reduce the efficiency somewhat but usually only a few percent because the tower, even with multiple Yagis for loading, is unlikely to be resonant anywhere near 475 kHz, not to mention 137 kHz. The coupling can be minimized by spacing the top anchor point as far out from the tower as possible, several feet would be helpful. Pulling the bottom and/or the top of the downlead away from the tower as shown in figure 3.28 can also help. The effect on a specific installation is best explored with modeling.

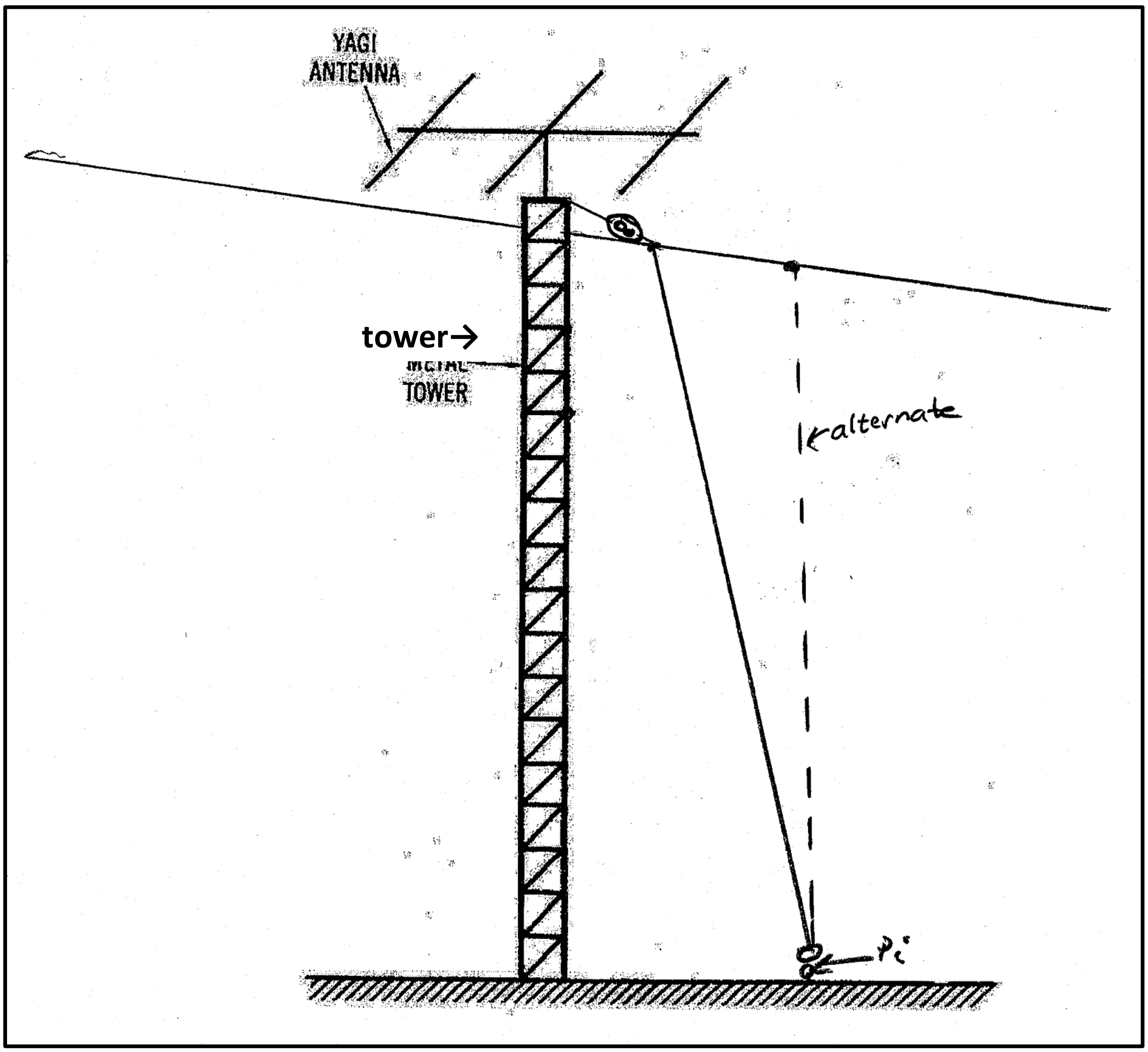

Figure 3.28 - Using a grounded tower as a center support.\

3.6 More complicated top-loading

While a vertical with a single top-wire is attractive, we're not limited to a single wire for top-loading. More complex top-loading can substantially improve efficiency. For example, we can add spreaders and use two or more wires as shown in figure 3.29. One note, the T and L models for this study show only a single conductor down-lead. When multiple top wires are used, the down-lead can also have multiple wires for at least part of its length which can make a small improvement in efficiency by reducing Rc. One of the problems with spreaders is that they tend to rotate and twist. Extra downleads can act as stabilizers. Light non-conducting lines can also be used to stabilize the spreaders.

Going from a single wire to two wires with 10' spreaders, makes a huge difference. For example, with 100' top-wires and H=50': for one wire the efficiency is ≈17% but with two wires and 10' spreaders that jumps to ≈28% and when we go to 30' spreaders the efficiency is ≈34%. However, there are practical limits to spreader length. 10' is easy, 20' takes some doing but spreaders longer than 20' are challenging. When the spreader length is increased to 20' the efficiency increases to 32%, significant but not nearly as great as the initial increase. For two wires 20' spacing is well into the region of vanishing returns and it's time to consider adding another wire or two. Figure 3.30 compares examples using 2, 3, 4 and 5 wires with a spreader length of 20' and an overall length of 100'. Clearly adding more wire to the hat reduces Xc and leads to higher efficiency but as can be seen in figure 3.30, for a given spreader length the <u>rate</u> of improvement falls pretty quickly and in this case using more than five wires gains very little. Adding more wire to the top-hat helps to reduce Xc but there's an important drawback to more wire up in the air: if you live in a area with ice storms, the antenna becomes much more vulnerable. From figures 3.29 and 3.30 it's clear that the effective capacitance (Ct) of a top-hat with parallel wires depends on the separation between the wires. Unfortunately the calculations for Xt shown in section 3.2 are not useful for closely spaced wires. An excellent and very complete paper on calculating the capacitances associated with LF-MF loaded verticals, written in 1926 by Fredrick Grover, is available on-line[2]. Excerpts from the Grover paper can be found in Terman[3]. In general it's much easier to use modeling for more complex top-loading structures.

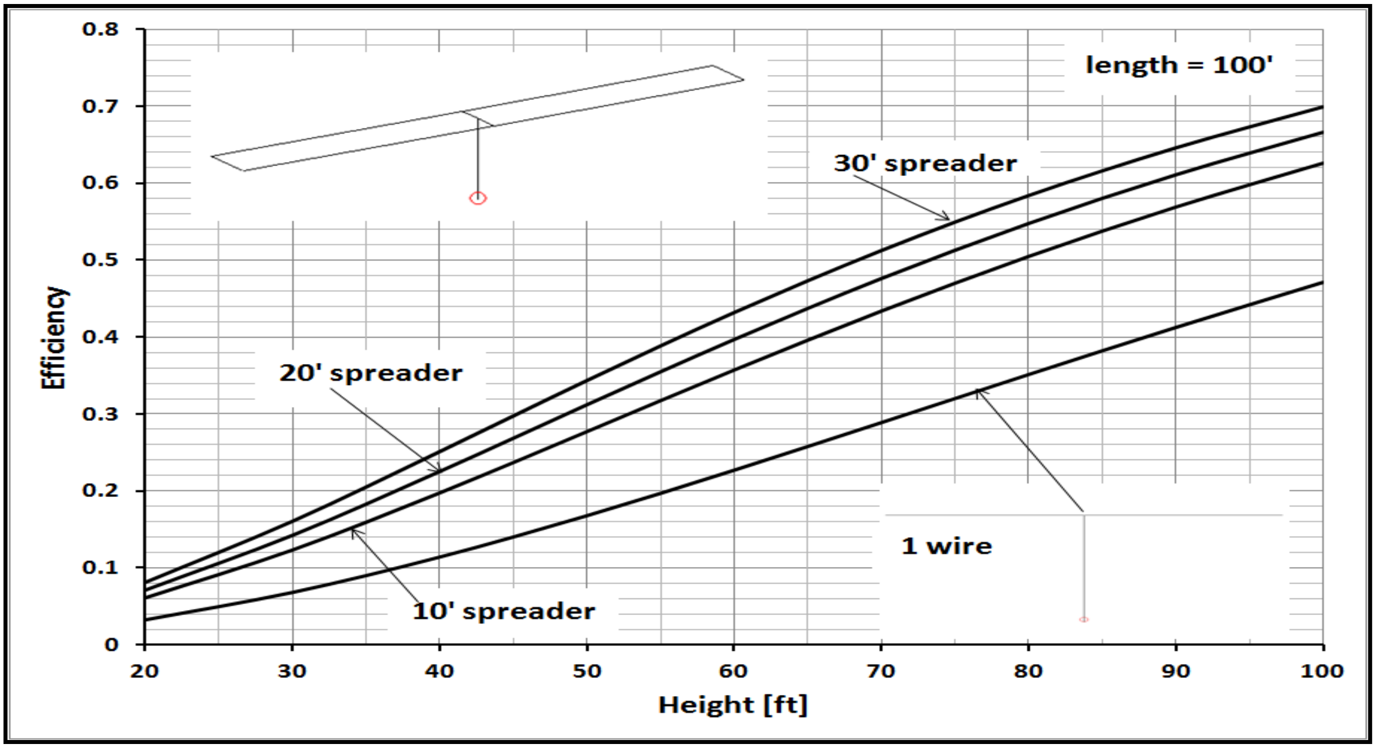

Figure 3.29 -Comparison of top-loaded verticals using a single wire or 2 horizontal wires with 10', 20' or 30' spreaders.

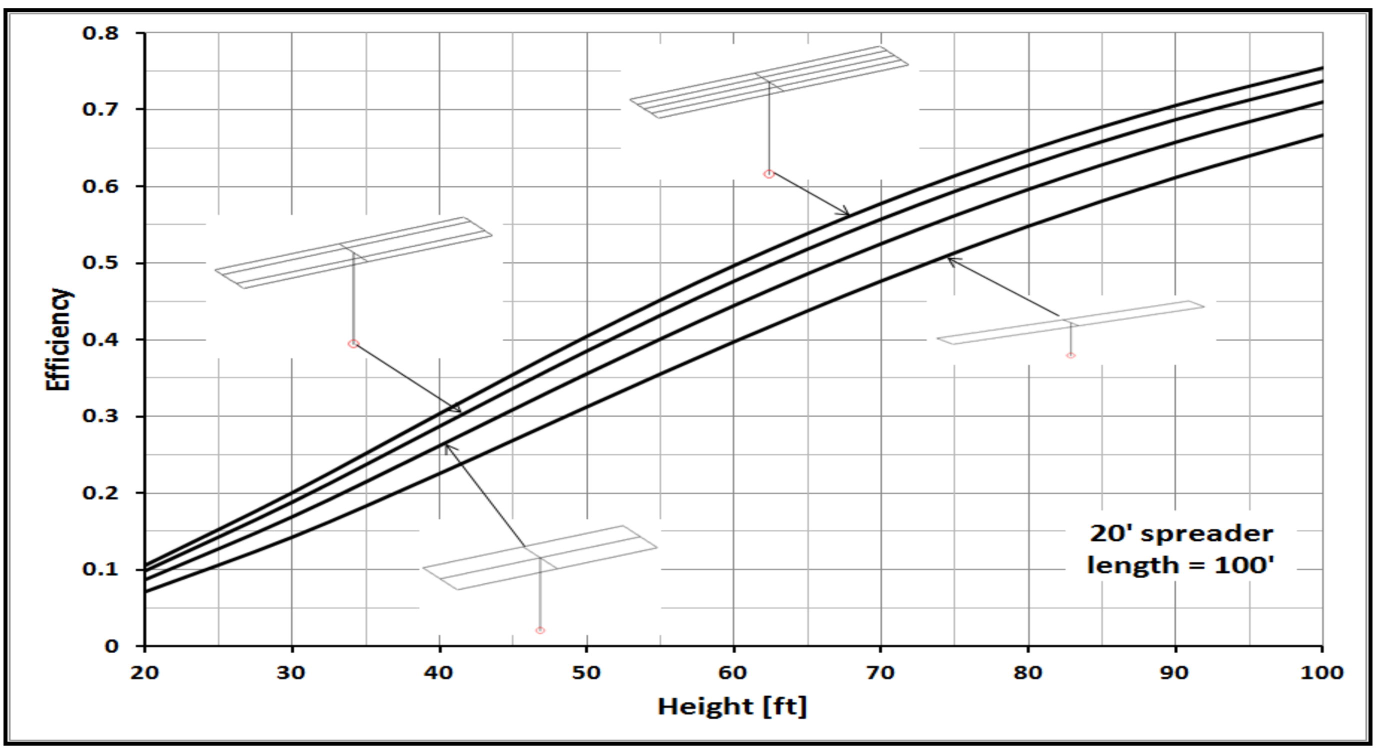

Figure 3.30 - Examples using more wires in the top-hat.

As shown in figure 3.31 in a multi-wire top-hat it may be possible to dispense with the center spreader, reducing the weight at the mid-point of the hat. The efficiency falls by ≈2% compared to the flat top but the reduction in sag due to the decrease in center-weight may compensate for that. One problem with the "bow-tie" arrangement, especially if the ends with the spreaders are supported only by a single line, is the tendency to twist in the wind. With a spreader at the midpoint and a fan for the downlead, twisting is much less of a problem. For the bow-tie configuration you will have to add some restraining lines at the ends of the spreaders.

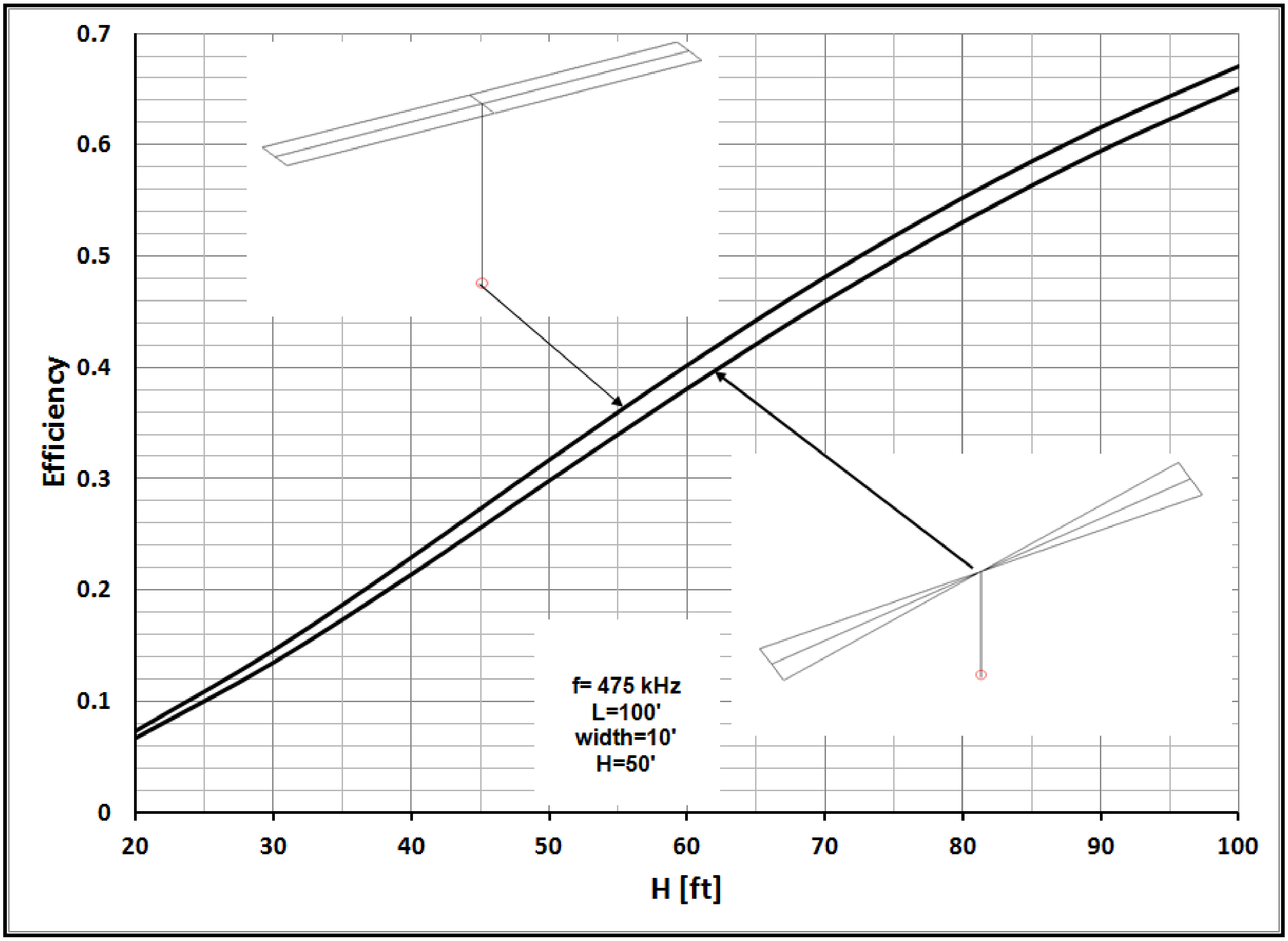

Figure 3.31 - 3-wire top-hat examples.

3.7 Umbrella top-loading

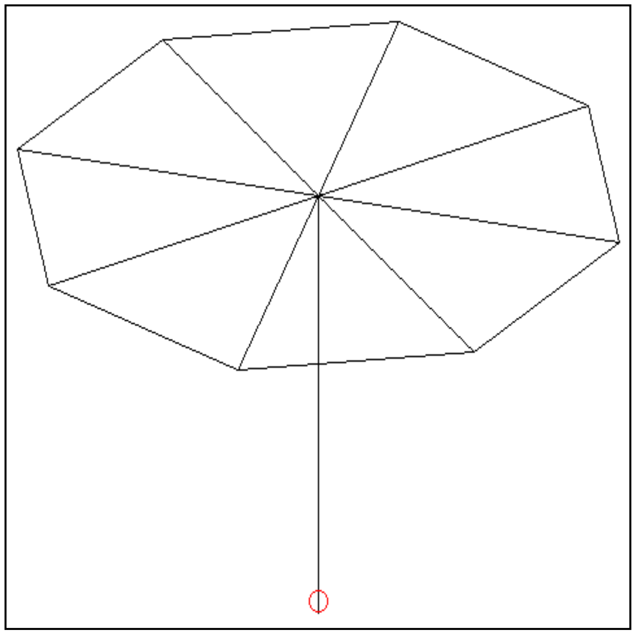

Figure 3.32 - Example of an "umbrella" for capacitive top-loading.

We've seen how helpful capacitive loading with various long top-wires can be. Now let's extend our investigation to symmetric top-hats which resemble umbrellas as shown in figure 3.32. Top-loading structures like this are often used when only one support (the vertical itself for example) is available.

There are many ways to construct an umbrella:

- Use several rigid radial supports like a wagon wheel. For example, the supports can be aluminum tubing or F/G poles supporting wires or some combination of the two. The practical limit for a self-supporting structure is probably a radius of 20' or so. Using the hub(s) and spreaders from an old HF quad can be a very simple way to fabricate an umbrella.
- Set up a circle of poles (3 to 8) at some distance from the base of the vertical to support the far ends of the umbrella wires. The radial dimension of the umbrella could be quite large but the length of the poles, which establishes the height of the outer rim of the umbrella, is a limiting factor.
- Attach a number of wires to the top of the vertical, sloping them downward at an angle towards anchor points located at some distance from the base of the vertical.
- You can add sloping wires to the outer rim of a horizontal umbrella to increase the loading.

- You do not have to connect the outer ends of the umbrella radial wires together with a "skirt" wire but a skirt-wire significantly increases the loading effect of an umbrella of a given radius.
- While most of this discussion shows symmetric umbrellas, symmetry is not required. Supports at different distances with different heights can also be used to good effect. Take advantage of what's on site!

For much of the modeling an 8-radial umbrella with a skirt wire is used. This represents a compromise. As few as three wires (with a skirt!) can make a useful umbrella but the performance improves substantially as you go from three to four and then eight wires. While the jump from four to eight wires gives a useful improvement the law of diminishing returns starts to set in and sixteen wires is about the useful limit. It's also possible to add more skirt wires inboard of the outer skirt wire. These are issues best explored with modeling.

Figure 3.33 is a generic sketch of the dimension designators used in the discussion.

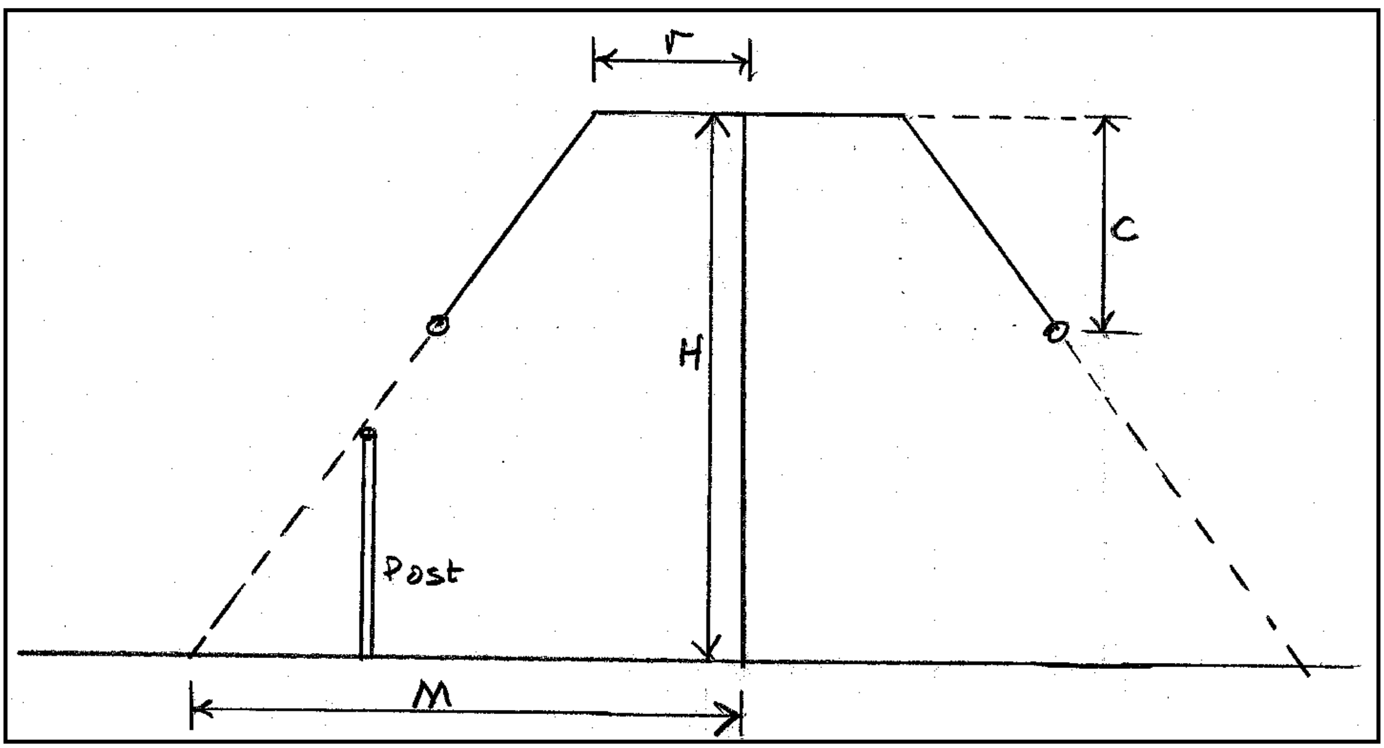

Figure 3.33- Definition of antenna dimensions.

Where: H=vertical height, r=radius of a flat umbrella, C=depth of a sloping umbrella, M=distance from the base to an anchor point,. When the available space restricts the radius to the anchors (M), as shown, a post can be used to increase the slope angle.

3.8 Horizontal or flat umbrellas

Figure 3.34 shows how Rr and Xc vary with H and r for a simple horizontal umbrella. r is varied from 0 (no umbrella) to 10' and then to 50'. We can see that even an umbrella radius as small as 10' makes a marked improvement in both Rr and Xc.

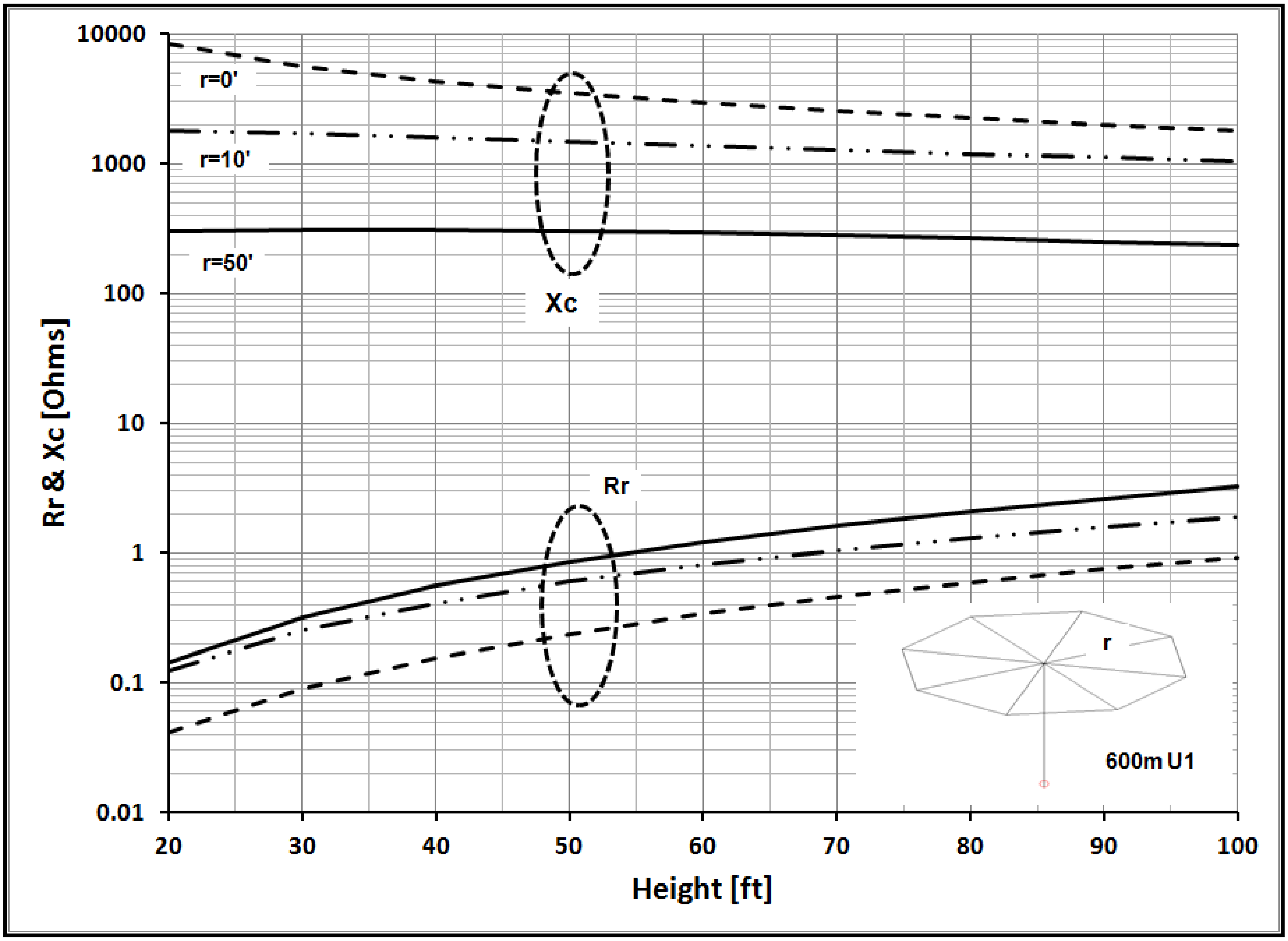

Figure 3.34 - Rr & Xc as a function of H and r.

Taking the data from figure 3.34 and assuming QL=400 we get the efficiencies shown in figure 3.35. The larger the umbrella, the better the efficiency! Figure 3.36 compares T top-loading to umbrella loading. A circular umbrella with r=20' is almost the same as a 3-wire T with 10' spreaders and 100' wires.

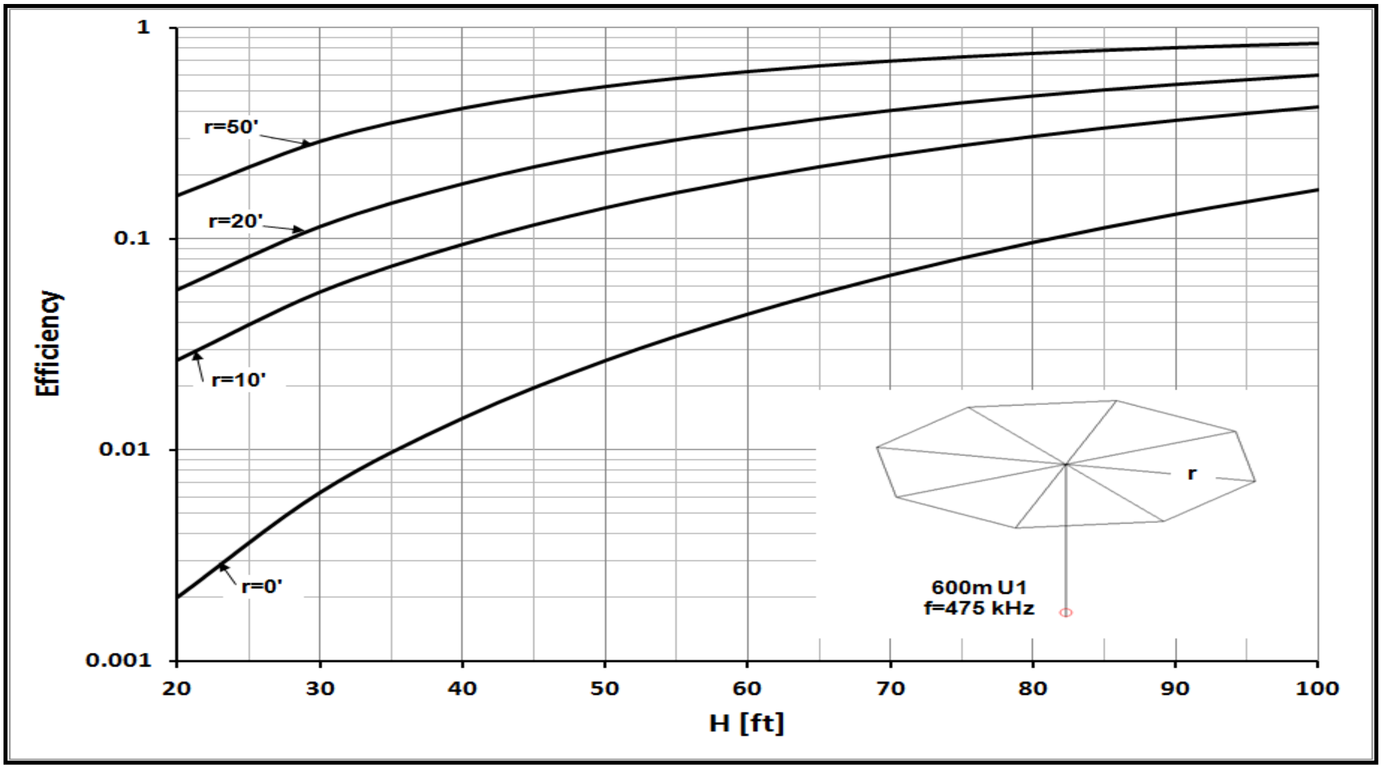

Figure 3.35 - Efficiency as a function of H and r.

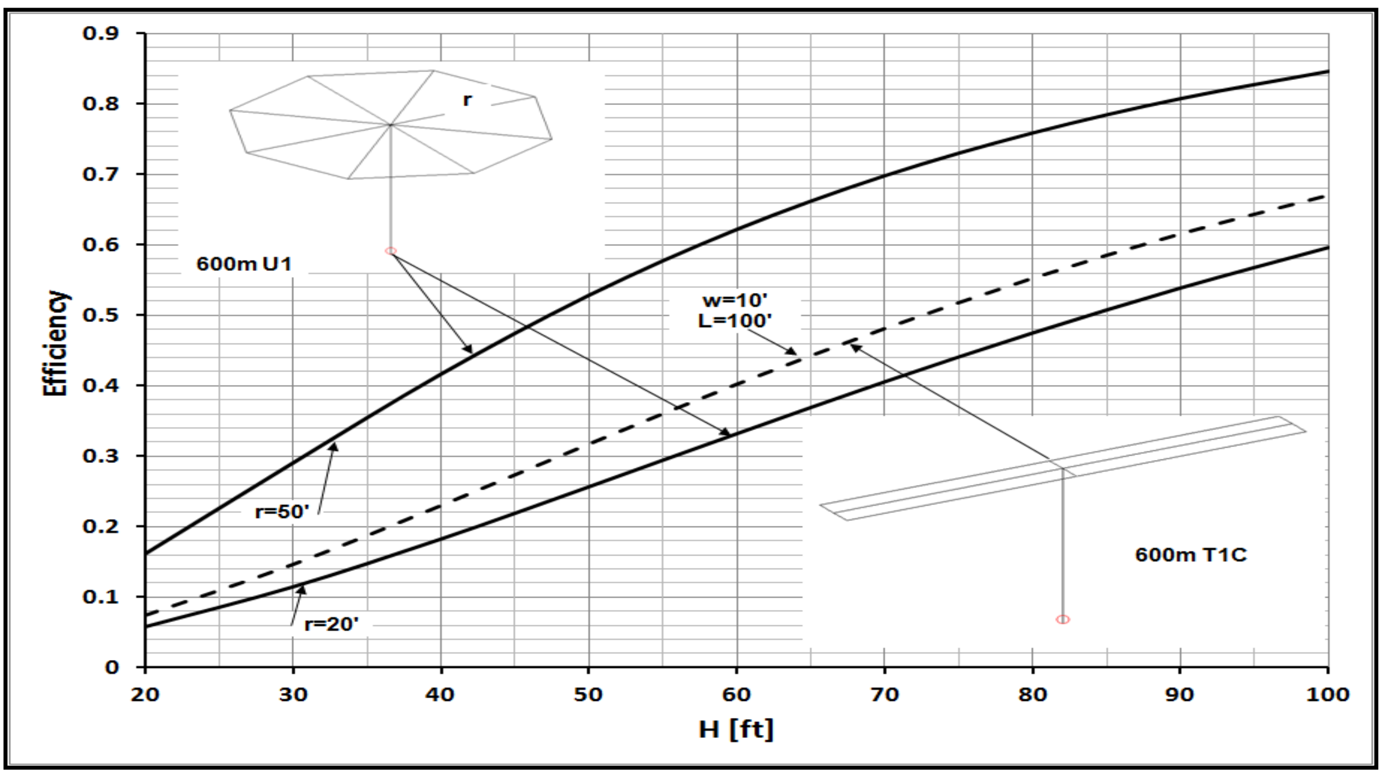

Figure 3.36 - Comparison between T and symmetric top-loading umbrellas.

3.9 Sloping wire umbrellas

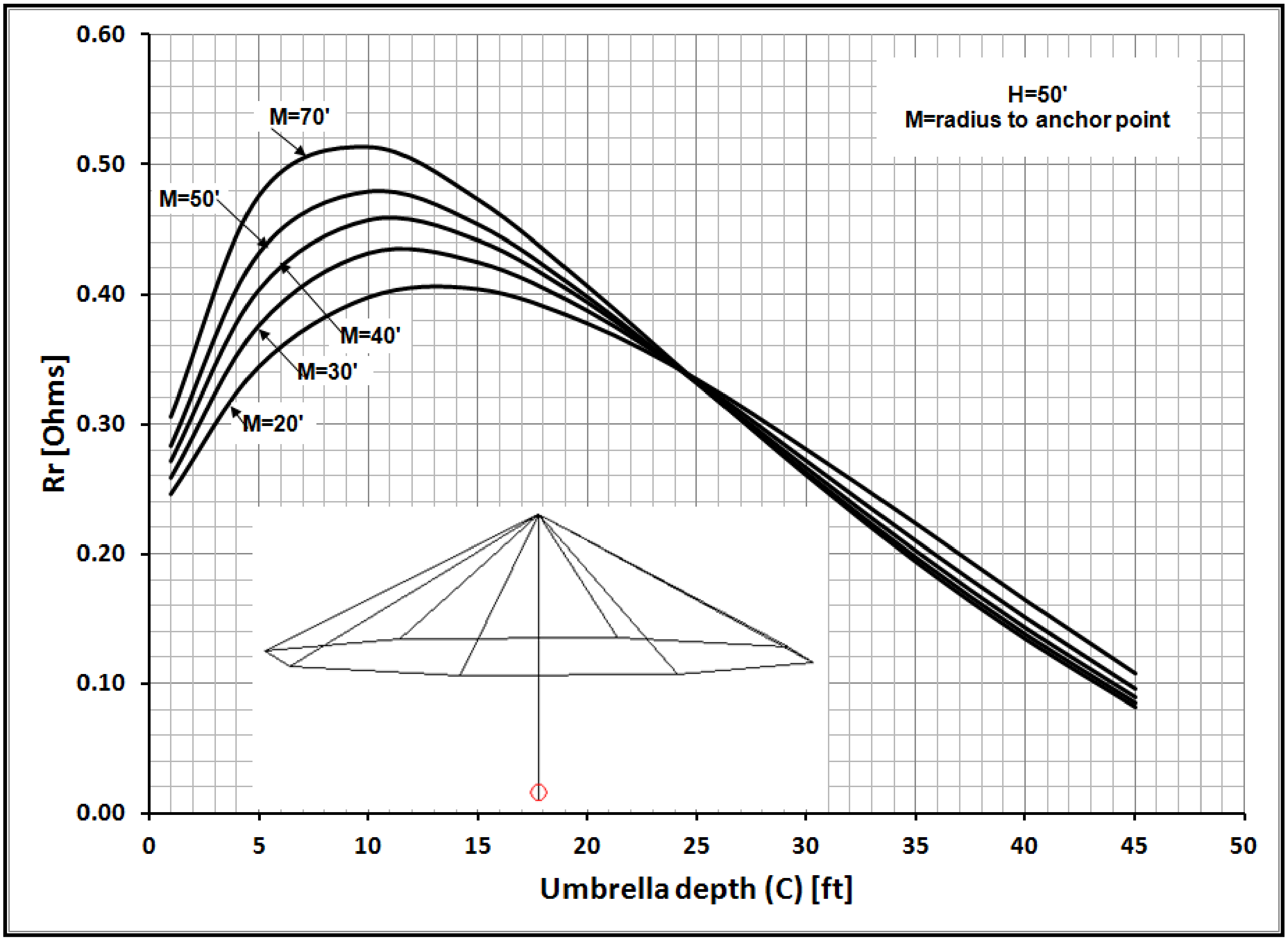

Figure 3.37 - An example of a sloping wire umbrella.

A large horizontal umbrella may not be practical. A alternative is to connect the umbrella wires to the top of the vertical and slope them downwards towards ground as shown in figure 3.37 which includes plots of Rr for H=50' over a range of anchor point distances (M) and umbrella depths (C). Other heights can of course be used and the results would be similar. The angle between the umbrella wires and the top of the vertical is: θ=ATAN(M/H). The larger we make M, the larger θ becomes. For a horizontal umbrella θ=90˚. As shown in figure 3.34, with a flat umbrella Rr increases steadily as the radius is increased but in the case of a sloping umbrella as we increase either M or C, Xi goes down but Rr doesn't increase continuously. For a given M, as we increase C, Rr rises initially, reaches a maximum and then decreases. This decrease in Rr is due to the vertical component of umbrella current opposing the current in the vertical.

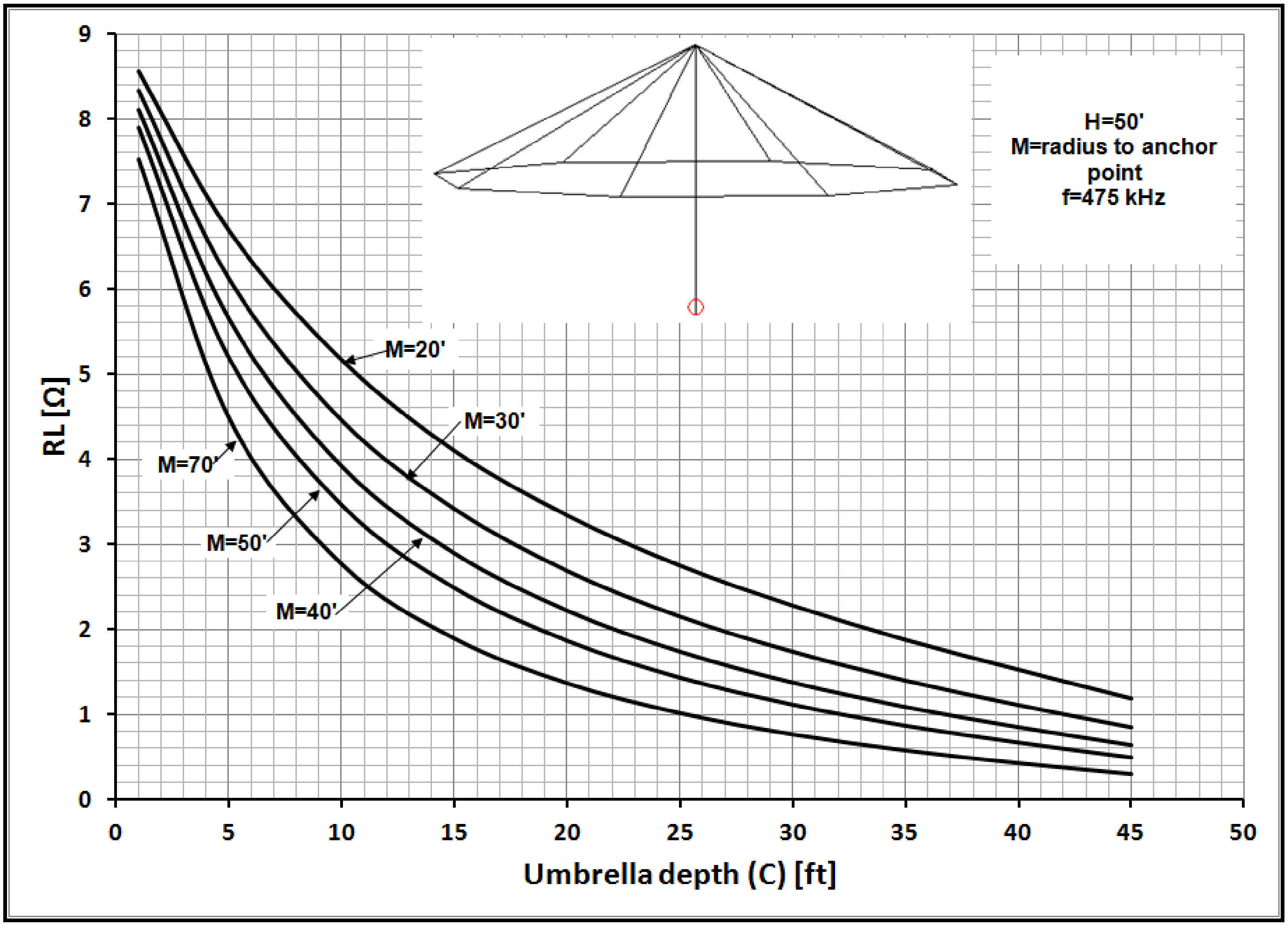

Figure 3.38- Variation of RL with A and C.

RL as a function of M and C is graphed in figure 3.38. We can take the information in figures 3.36 and 3.37 to create a graph of RL/Rr as shown in figure 3.39. We can see the RL/Rr ratio continues to decrease well beyond the peak for Rr as we increase C but a point is reached (C≈0.4-0.5 H) where the ratio flattens out. Beyond C≈0.4 there's no point on continuing to expand the umbrella.

We can take the numbers in figure 3.38 and use equation (3.1) to determine the efficiency as shown in figure 3.39 which indicates there is a optimum value for C with a given value of M. However, the optimum is very broad so the value for C is not critical. Figure 3.40 illustrates how increasing the distance to the anchor points increases efficiency.

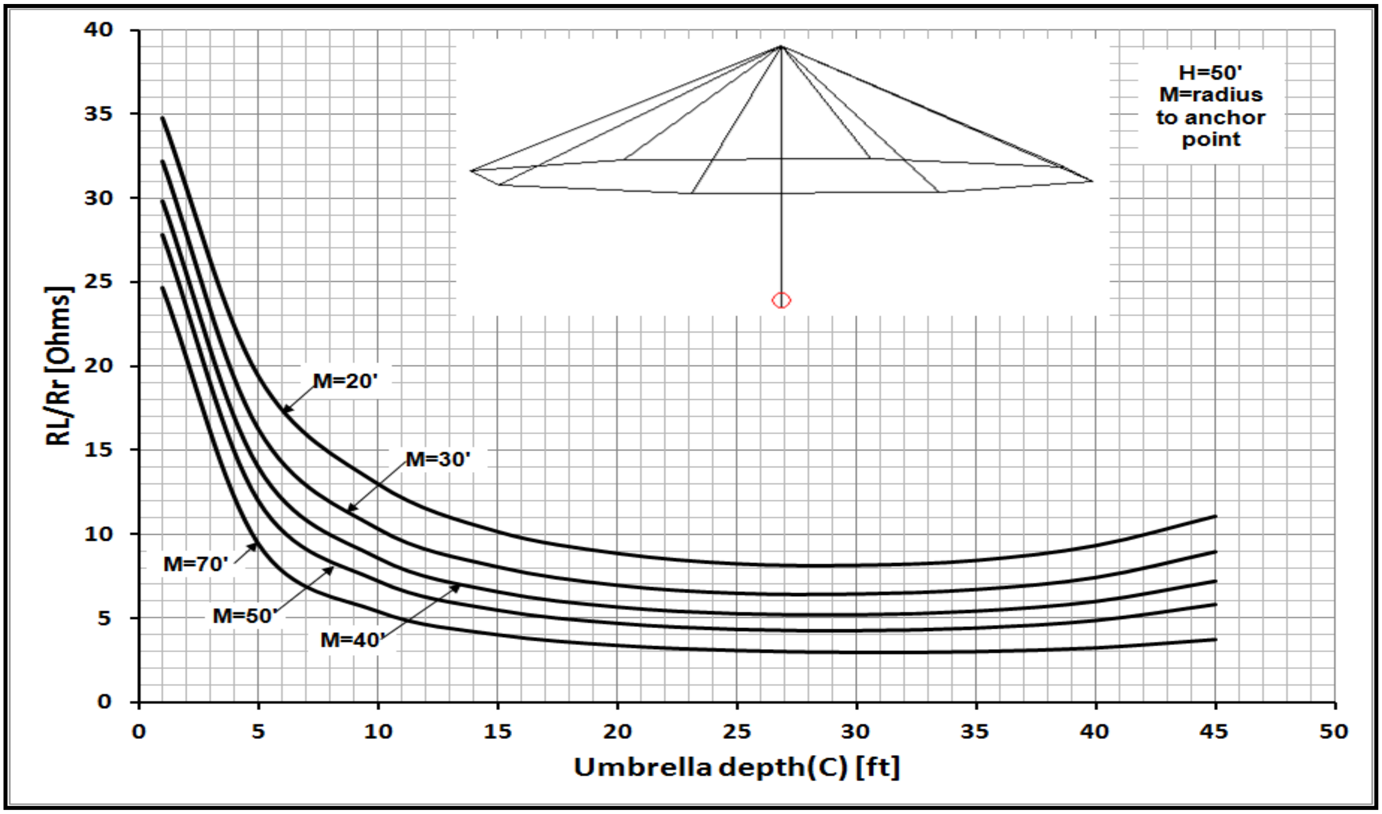

Figure 3.39 - Variation of RL/Rr with M and C.

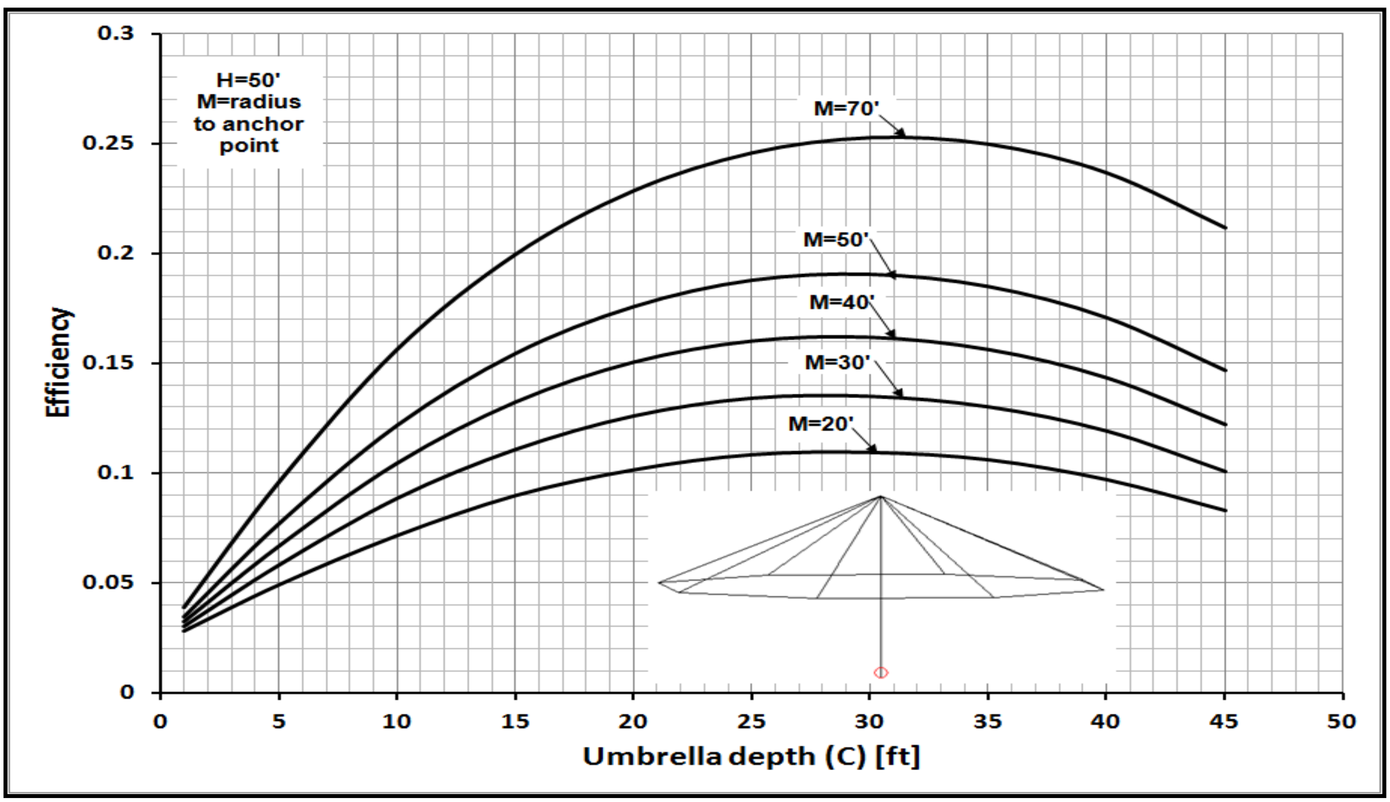

Figure 3.40- Efficiency as a function of M and C.

3.10 Combining sloping and flat umbrellas

There are practical size limits for a horizontal umbrella but we can improve the performance by adding sloping wires to the outer perimeter to form a composite umbrella. An example where the "sloping" wires actually hang straight down from the horizontal part of the umbrella is shown in figure 3.41. Due to the opposing currents in the drop wires Rr decreases as the drop wire length (C) is made longer. However, RL is also decreasing due to a falling Xc. The effect of C on efficiency is shown in figure 3.41.

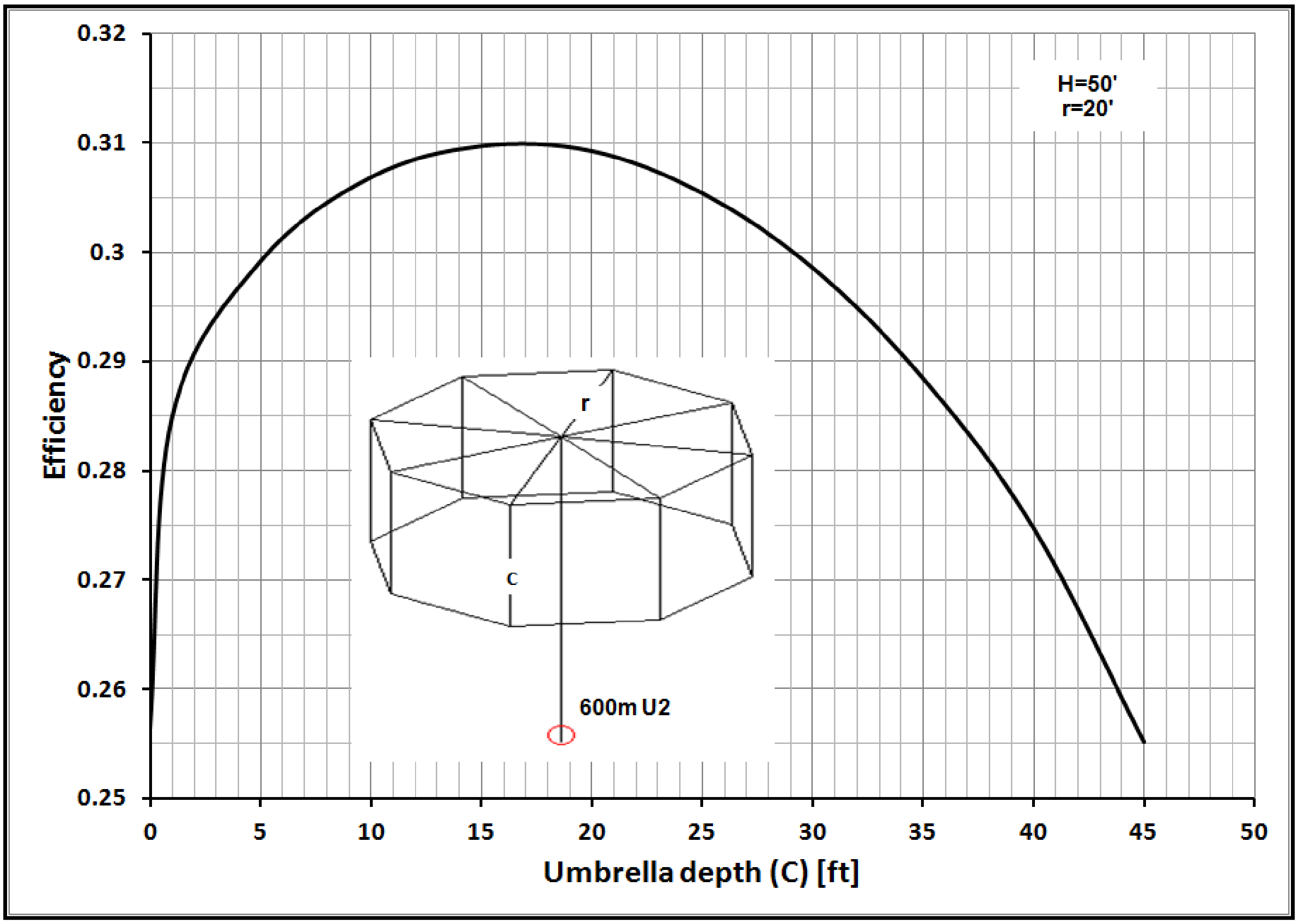

Figure 3.41 - Efficiency of a horizontal umbrella with vertical drop wires on the perimeter of the umbrella.

Adding drop wires allows an increase in efficiency of nearly 6% at C≈0.35H in this example. The peak is also quite broad so C is not critical. This type of umbrella might work very well in an urban backyard.

If there is space to move the anchor points for the drop wires further away then the drop wires can be sloping as shown in figure 3.42, where H=50' and M is kept constant at 50' as r is varied.

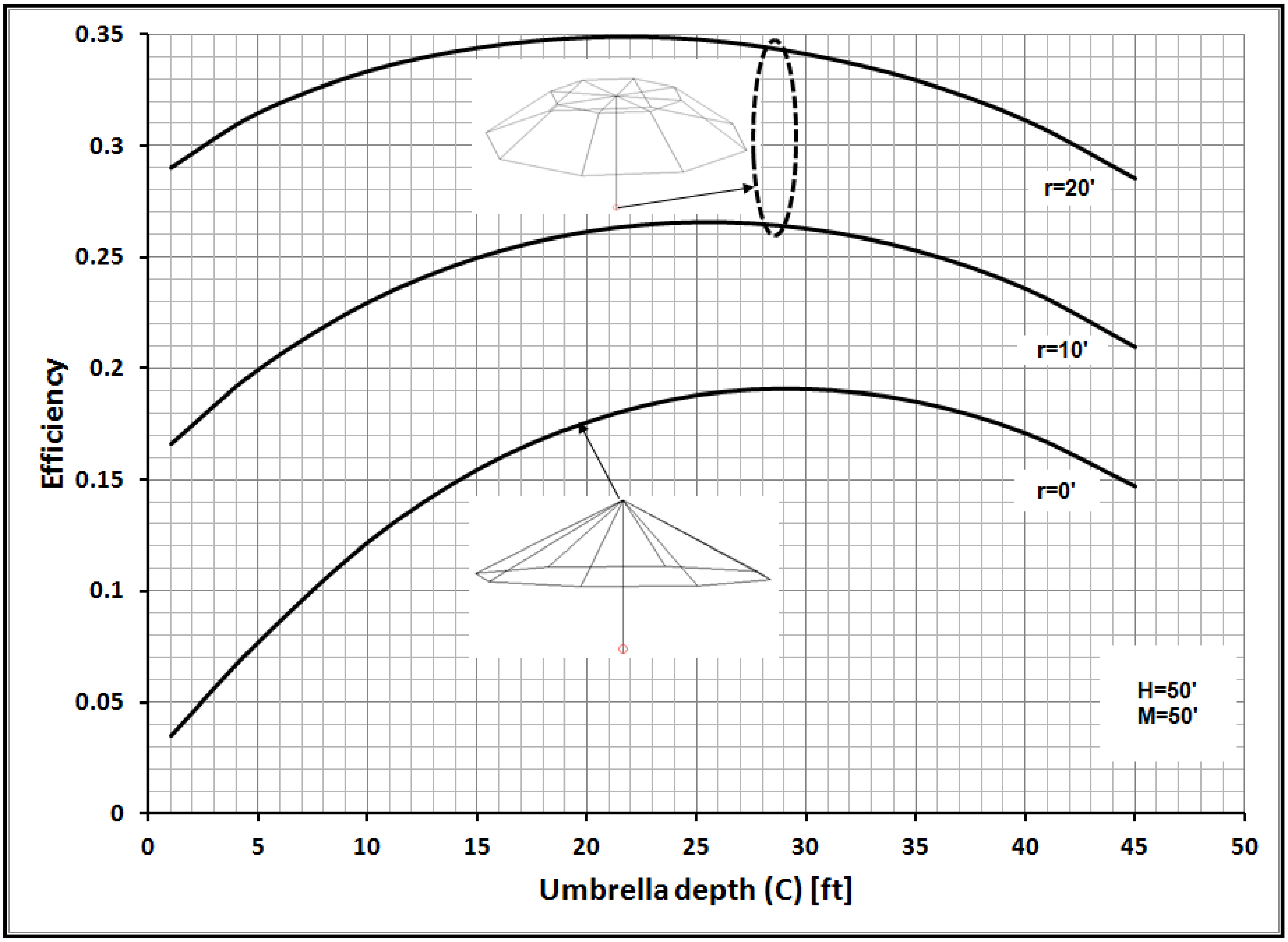

Figure 3.42 - Efficiency for a combined sloping and flat umbrella.

The r=0' trace represents the case of only sloping umbrella wires. The r=10' and 20' traces show the improvement gained by adding some horizontal component to the umbrella. For r=0', η≈19% but when r is increased to 10', η≈27% and for r=20', η≈35%. These are very worthwhile improvements which indicates that adding even a small horizontal section to the umbrella is worthwhile. Compared to figure 3.41 where the drop wires were vertical, sloping the drop wires away from the vertical shows an improvement in efficiency of ≈4%.

3.11 Rc and Conductors

Copper or aluminum wire and aluminum tubing are typical conductors. The wire may be bare or insulated. The choice of conductor is usually a matter of what's on hand and/or what's economical. Insulated wire intended for home wiring is of often the most economic choice for copper wire. Aluminum electric fence wire, available in #14 or #17 gauges, is a less expensive choice but aluminum has greater resistance than copper and because soldering aluminum is often not satisfactory, joints require special attention. We can work around the resistance issue by using multiple, well spaced, wires in parallel.

The resistance of a wire at DC (Rdc) is:

$$\boldsymbol{Rdc = \frac{l}{\sigma Aw}} \quad [\Omega] \quad (3.12)$$

Where l is the length of the wire, Aw is the cross section area and **σ** is the conductivity in Siemens/meter [S/m]. For copper $\sigma=5.8X10^7$ [S/m] and for aluminum $\sigma=3.81X10^7$ [S/m]. Rdc for copper wire can be found in wire tables.

Unfortunately the AC resistance of the wire (Rac) will be substantially different from Rdc due to two effects: skin effect and the effect of non-uniform current distribution.

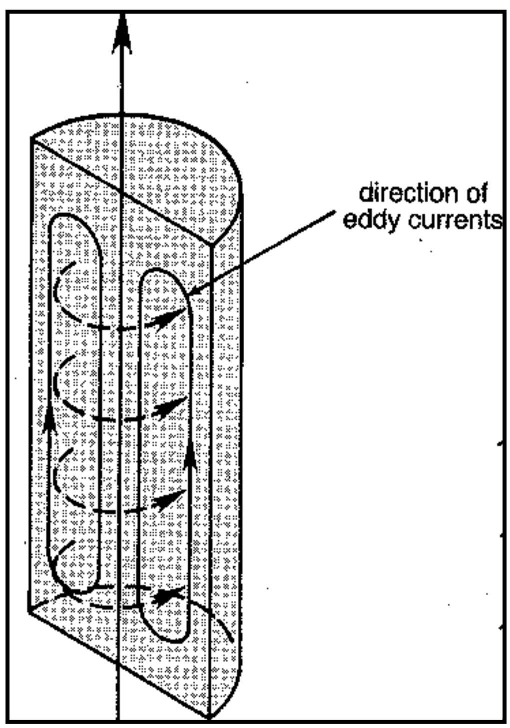

Figure 3.43 - Skin effect.

The cause of skin effect is the magnetic fields associated with the current flowing in the conductor as shown in figure 3.43. The dashed lines represent magnetic field lines resulting from the desired current flowing in the conductor. The solid lines represent currents induced in the conductor. At RF frequencies the current is concentrated in a very thin layer at the surface of the wire, hence the term "skin effect". Because the current is concentrated in a thin layer the resistance of a round conductor will proportional to the diameter (d), i.e. the wire circumference, rather than the cross sectional area. Changing from a solid conductor to a tubular one greatly reduces the amount of metal required. Typically solid wire up to #8 (d=0.13") is used. To lower losses further either two or more wires in parallel or thin wall aluminum tubing are used. Aluminum irrigation tubing, in sizes from d=2" to 6", is widely available, at least in rural areas. The larger sizes can be self supporting or need only limited guying. Making the vertical from tubing is very helpful when no other supports are available.

The "penetration" or "skin" depth δ is expressed by:

$$\boldsymbol{\delta = \frac{1}{\sqrt{\pi\sigma\mu f}}} \text{ [m]} \quad (3.13)$$

Where:

σ is the conductivity in [S/m].

μ is the permeability of the material which for Cu and Al $=4\pi x10^{-7}$ [H/m].

f is the frequency [Hz].

The skin depth in <u>mils</u> for copper is:

$$\boldsymbol{\delta = \frac{2.602}{\sqrt{f_{MHz}}}} \text{ [mils]} \quad (3.14)$$

At 137 kHz δ=7.03 mils (0.00703") and at 475 kHz δ=3.78 mils.

The resistance of the wire, Rc, can be represented by the product of the DC resistance (**Rdc**) and a factor attributed to skin effect (**Ks**).

$$\mathbf{Rac = Rdc \cdot Ks} \quad (3.15)$$

For the wire sizes typically used in tuning inductors, d/δ> 5, and Ks becomes a linear function of d/δ which can be closely approximated with a simple expression:

$$Ks = \left(\frac{1}{4}\right)\left[\frac{d}{\delta} + 1\right] \quad (3.16)$$

which is graphed in figure 3.44. Note that Ks is only a function of the wire diameter d and the skin depth δ at the frequency of interest (δ∝1/√f).

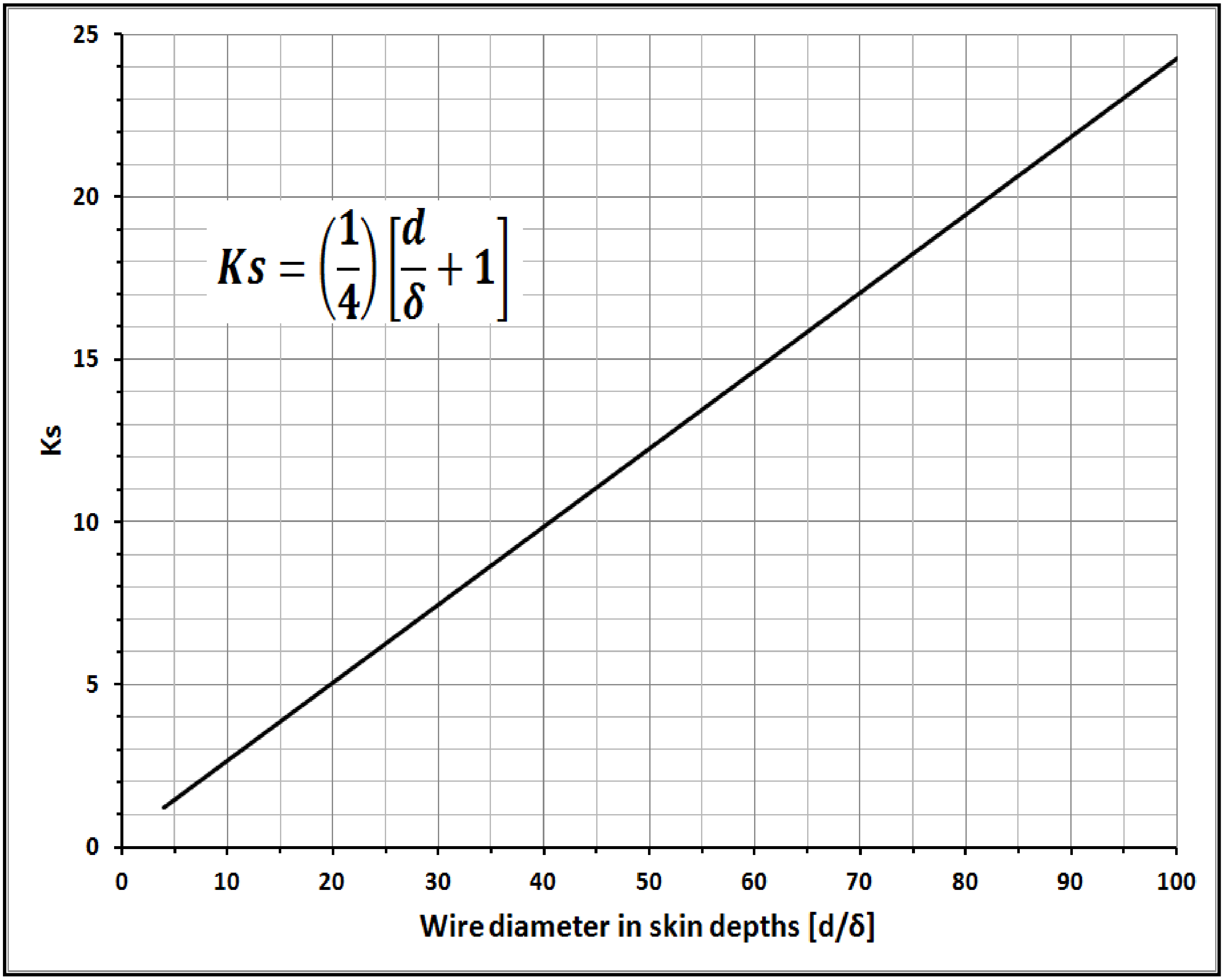

Figure 3.44 - Skin effect factor Ks versus wire diameter in skin depths (**d/δ).**

Table 3.2 gives Ks for typical wire sizes at 137 and 475 kHz.

Table 3.2 - Ks for typical wire sizes.

	137 kHz	475 kHz
Wire #	Ks	Ks
8	4.82	8.76
10	3.87	7.00
12	3.10	5.55
14	2.53	4.49
16	2.06	3.61
18	1.15	1.93

The current on antenna conductors will usually be non-uniform, i.e. the current amplitude will vary from one point to another. When the antenna is large enough to approach self-resonance the current distribution can be close to sinusoidal as shown in figure 3.45A. For small LF-MF antennas however, the current will be approximately linear as indicated in figure 3.45B.

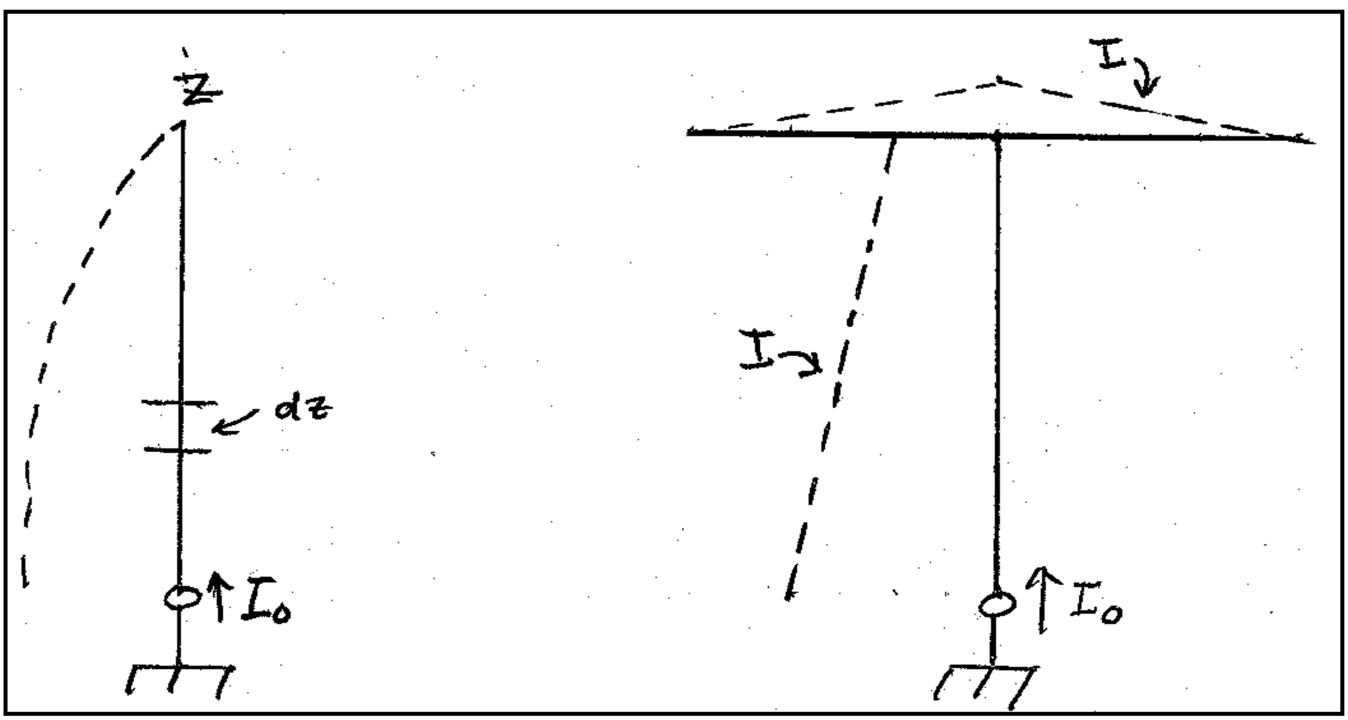

Figure 3.45 - Antenna current examples.

Io= rms current at the high current end of the wire. Re= effective resistance of the wire which results in the same power dissipation for a given Io as the actual power dissipation on the wire. Re/Rac is the resistance ratio due to non-uniform current distribution.

A graph of Re/Rac for a linear current distribution is given in figure 3.46.

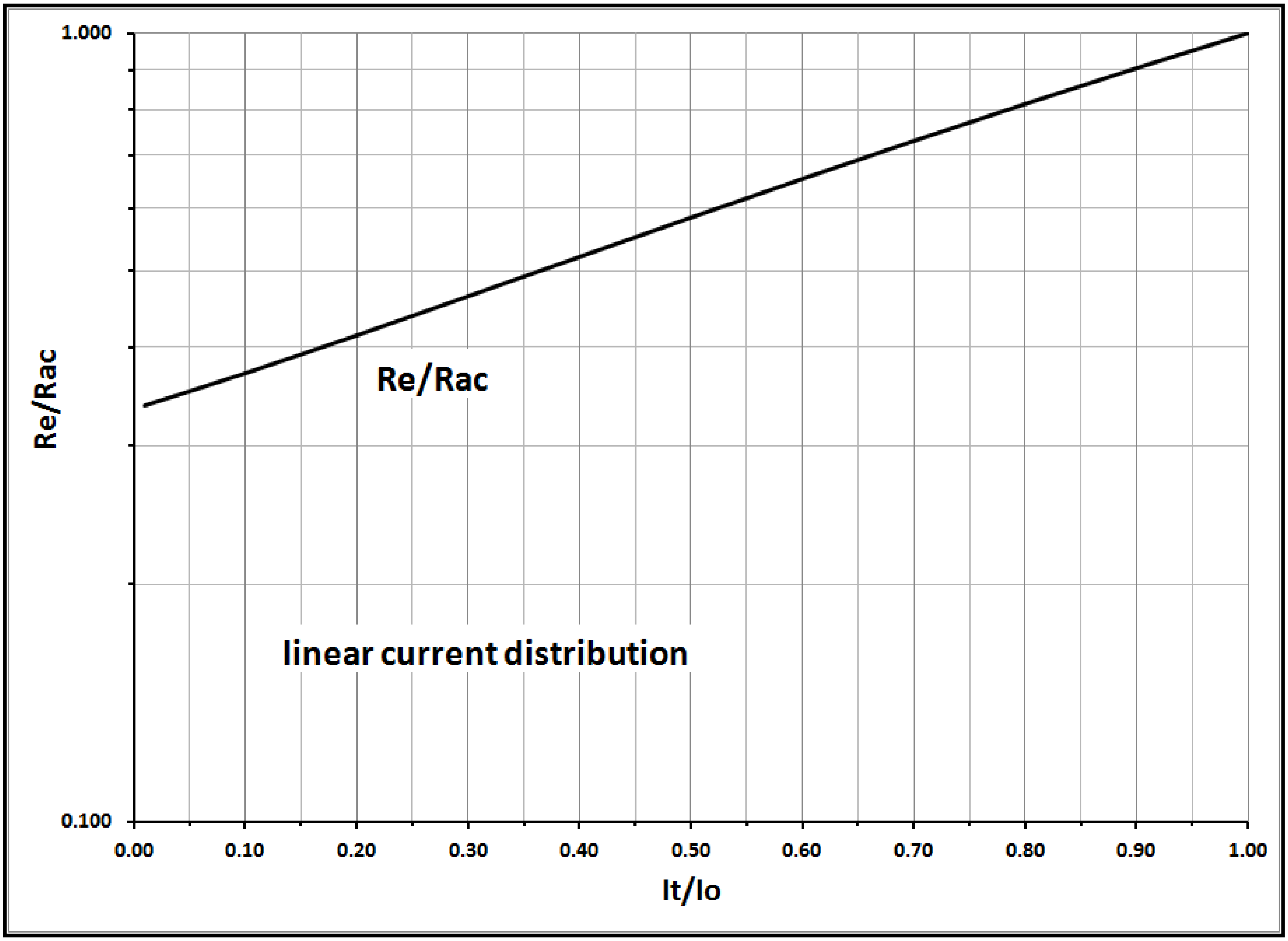

Figure 3.46 - Re/Rac versus It/Io.

For a vertical with no top-loading, It=0 and *Re/Rac=1/3*. More detailed information on Ks can be found in chapter 6.

The cost of the conductors may be a concern. For copper $\sigma=5.8x10^{7}$ [S/m]. Aluminum has somewhat lower conductivity, $\sigma=3.81x10^{7}$ [S/m], but is much less expensive and a better strength to weight ratio.

Replacing copper with aluminum:

$$\frac{\mathbf{Rc_{Al}}}{\mathbf{Rc_{Cu}}} = \sqrt{\frac{\boldsymbol{\sigma}_{\mathbf{Al}}}{\boldsymbol{\sigma}_{\mathbf{Cu}}}} = \sqrt{\frac{\mathbf{3.81}}{\mathbf{5.8}}} = \mathbf{0.81} \quad (3.17)$$

The conductor loss (Pc) penalty of switching from copper to aluminum is ≈19%.

Up to this point we've not considered ground loss (Rg) or conductor loss (Rc). Including these losses the efficiency becomes:

$$\boldsymbol{\eta} = \frac{\boldsymbol{R_r}}{\boldsymbol{R_r + R_L + Rg + Rc}} \quad (3.18)$$

Rg will be discussed in chapter 5 but to properly evaluate the effect of conductor loss on efficiency we need to include it at this point. Figure 3.47 shows a "T" antenna with multiple downleads. The height is 50' over average ground. The top-wire is 100' and there are thirty two 50' buried radials. We want to know the effect of wire size on efficiency and the effect of using multiple downleads. A key question is "how much effect does the conductor loss have on efficiency?" From equation (3.18) we can see that the value for Rc matters but also its value in proportion to the sum of Rr+RL+Rg. As the conductor size in increased Rc will go down but at some point Rc becomes small compared to sum of the other terms so further reductions in Rc have limited effect.

Figure 3.47 shows the effect of wire diameter (d) on efficiency for different numbers of downleads. At any given point all the wires have the same diameter. d=0.01" corresponds to a #30 wire. d=0.150" corresponds to a #6 wire. From a practical point of view wire sizes ranging from #18 to #8 are the most likely, with #12 a very common choice. Figure 3.47 also shows that using more downleads improves efficiency, no real surprise there, but the graph also shows how the initial increase in wire size rapidly improves efficiency but the curve soon flattens out as the point of vanishing returns approaches. Note that for larger wire sizes the efficiency only changes by less than 1%!

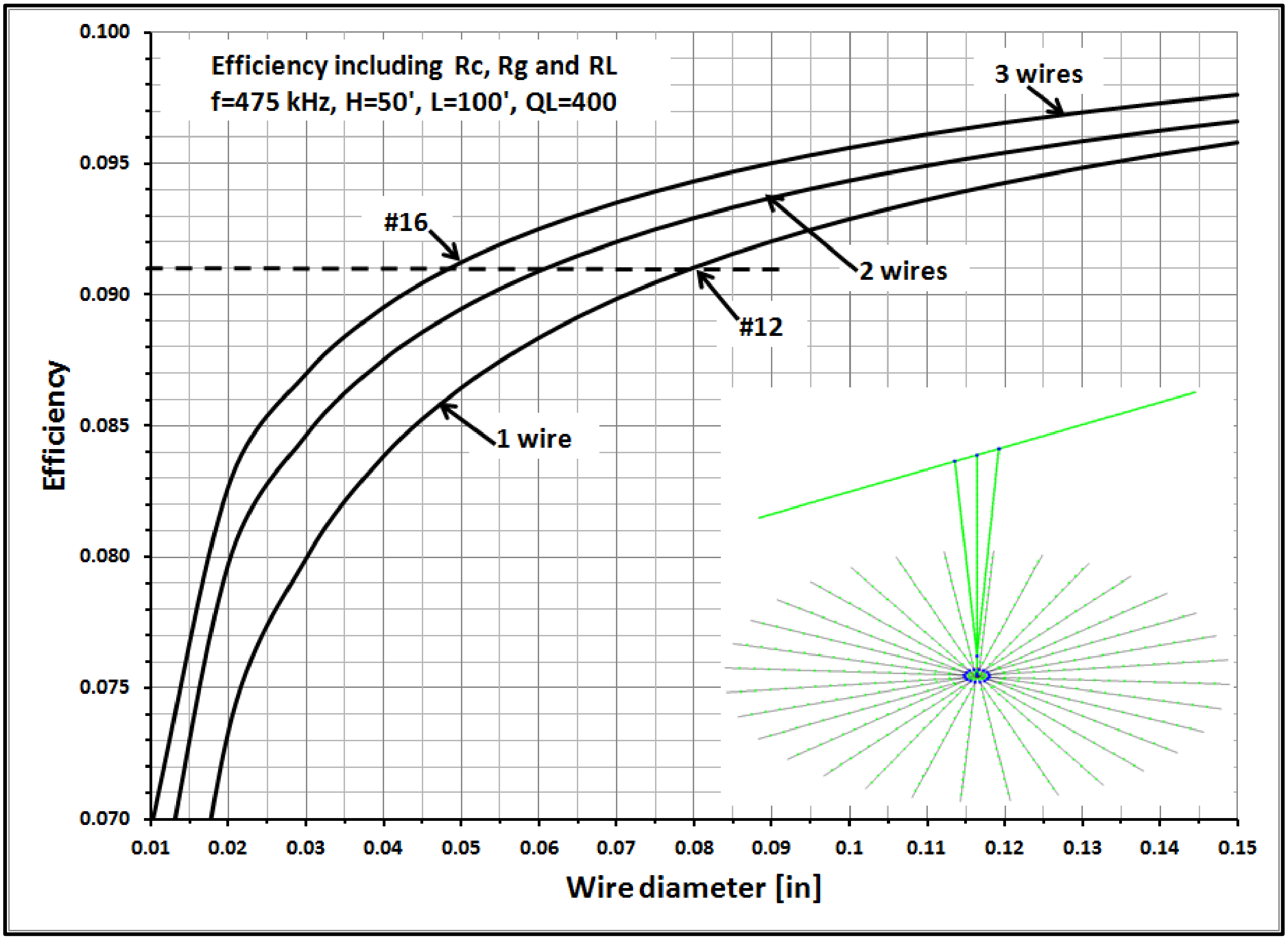

Figure 3.47 - Efficiency versus wire diameter for 1, 2 or 3 downlead wires.

Due to idiosyncrasies in wire pricing often there isn't an direct linear variation in wire cost with the amount of copper between different wire sizes but we can still graph the volume of wire as shown in figure 3.48 to give us an idea on the cost impact of different wire sizes and downlead numbers. What these graphs seem to be telling us is that the very common use of #12 wire is actually a very practical choice although we could save a bit by going down to #16 wire.

Insulated copper wire intended for home wiring is often used for antennas and ground systems, both elevated and buried. This wire is readily available at hardware and home improvement emporiums and often significantly less expensive than the equivalent wire without insulation (bare). Among amateurs there has been a recurring discussion whether it's necessary or even useful to strip the insulation. Stripping a few hundred feet isn't a serious chore but if you're laying out a radial field with thousands of feet of wire then stripping would be a chore.

For this example we will assume the antenna in figure 3.47 with only one downlead.

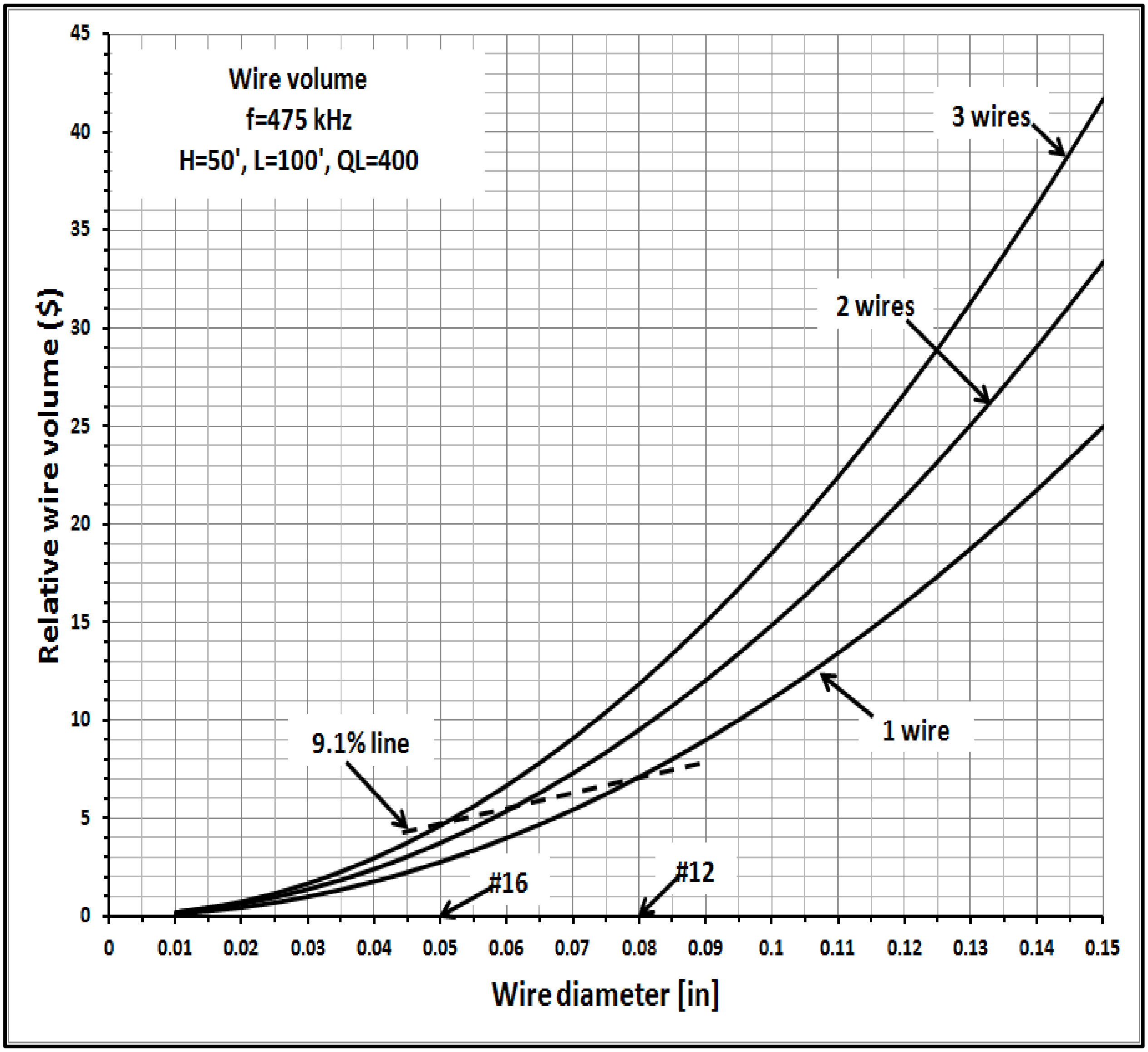

Figure 3.48 - Relative wire volume ($) versus wire size and downlead number.

1) The insulation introduces no additional loss even when the wire has been exposed to sun and weather for many years.

2) The insulation does introduce some dielectric loading which reduces Xi slightly, from 1245Ω to 1209Ω. This means that XL is slightly smaller reducing RL and increasing efficiency by ≈0.5% when QL=400.

3) The disadvantage of leaving the insulation on the wire is the increase in weight and diameter. The increase in diameter can lead to greater ice loading.

3.12 Voltage, current and power dissipation

At the end of chapter 2 we looked at the voltages (Vo) and currents (Io) at the feedpoint of a vertical without top-loading. The results were discouraging to say the least! Now it's time to see what happens to Vo and Io when top-loading is present.

Io is determined by the radiated power (Pr) and the radiation resistance (Rr):

$$\mathbf{Io} = \sqrt{\frac{\mathbf{Pr}}{\mathbf{Rr}}} \quad (3.19)$$

As explained in chapter 1, the maximum radiated power (Pr) is limited to 1.67W on 630m and 0.33W on 2200m. Combining the band specific values for Pr with Rr values derived earlier (figures 3.7 and 3.8) we can use equation (3.19) to create the graphs in figures 3.49 and 3.50. Note that L is the overall length of the top-wire in feet in all of the graphs in this section. L=0 represents the no top-loading condition.

Vo is the voltage at the feedpoint:

$$\mathbf{Vo} = \mathbf{XiIo} = \mathbf{Xi}\sqrt{\frac{\mathbf{Pr}}{\mathbf{Rr}}} \quad (3.20)$$

We can use typical Xi values from figures 3.4 and 3.5 to generate values for Vo as shown in figures 3.51 and 3.52. Note that the y-axis is logarithmic, emphasizing the rapid change in Vo with top-loading.

These graphs illustrate one important point:

Even a small amount of top-loading significantly reduces Io and greatly reduces Vo!

Despite the low radiated powers (Pr) Vo can easily approach 1kV on 630m and be even higher on 2200m. This must be kept in mind when selecting a base insulator. It's clear that substantial height (H) and top-loading are required on 2200m if Vo is to be kept below 10 kV. Given that we are trying to radiate only 330 mW that may come as a shock!

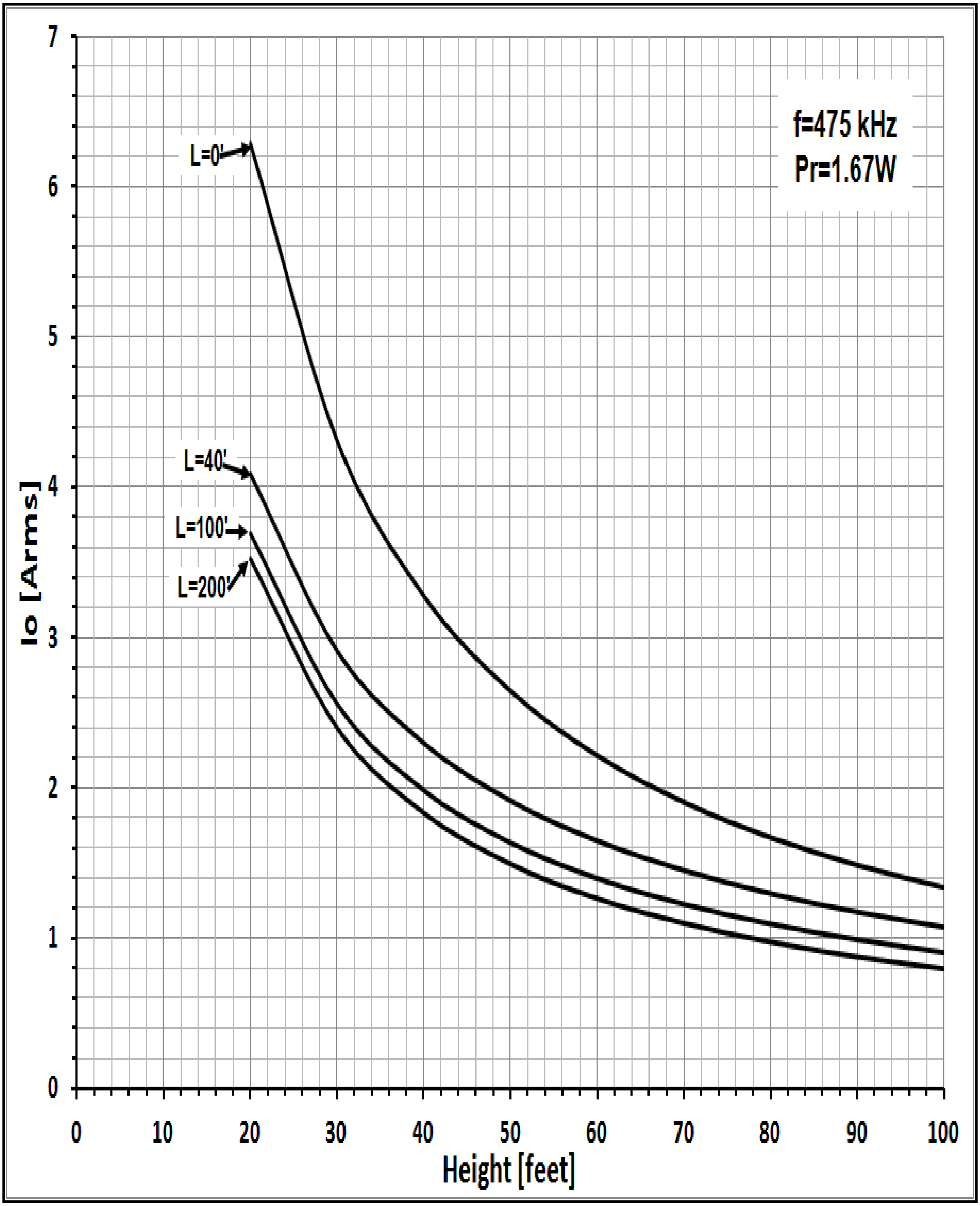

Figure 3.49 - Io for Pr=1.67W at 475 kHz.

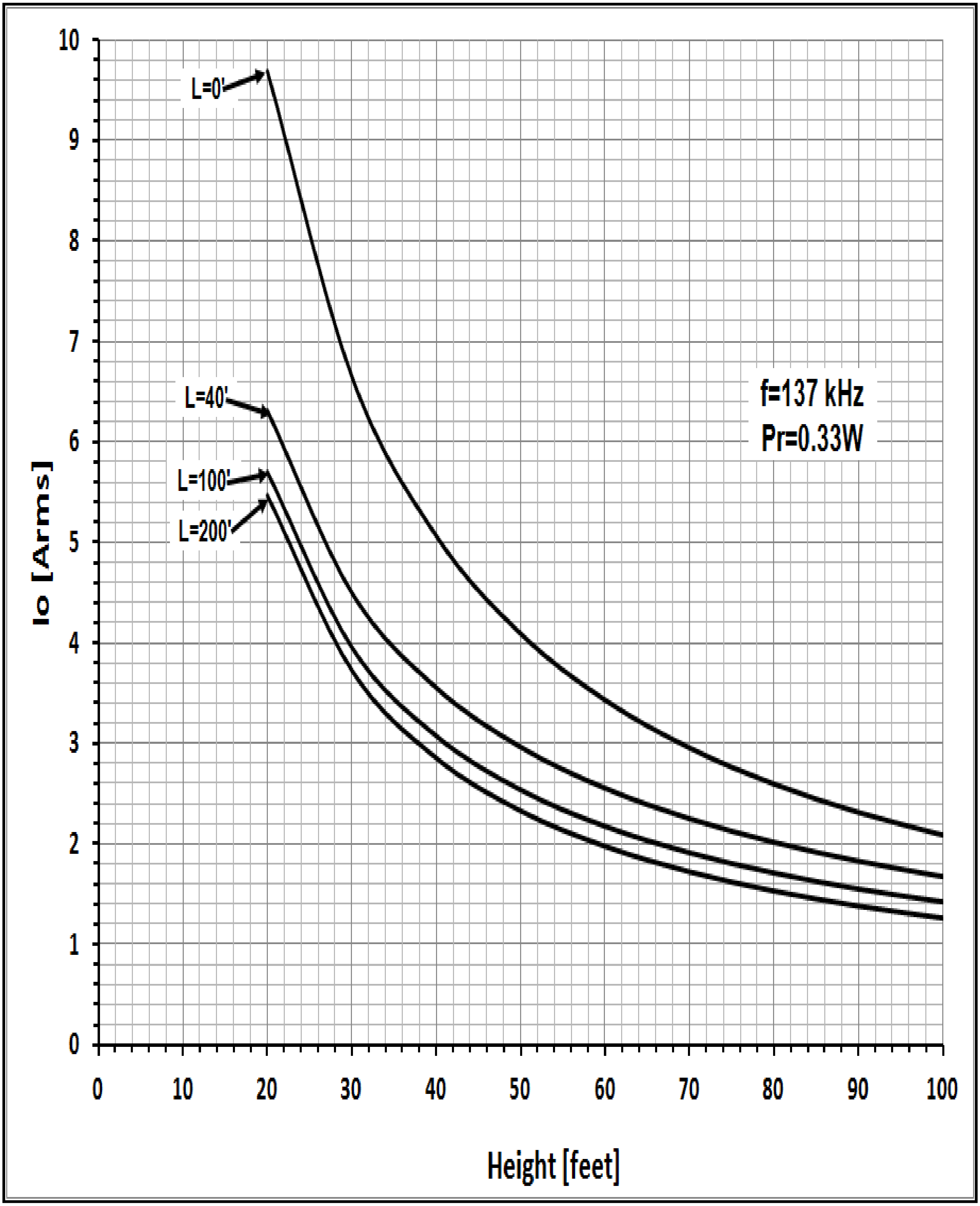

Figure 3.50 - Io for Pr=0.33W at 137 kHz.

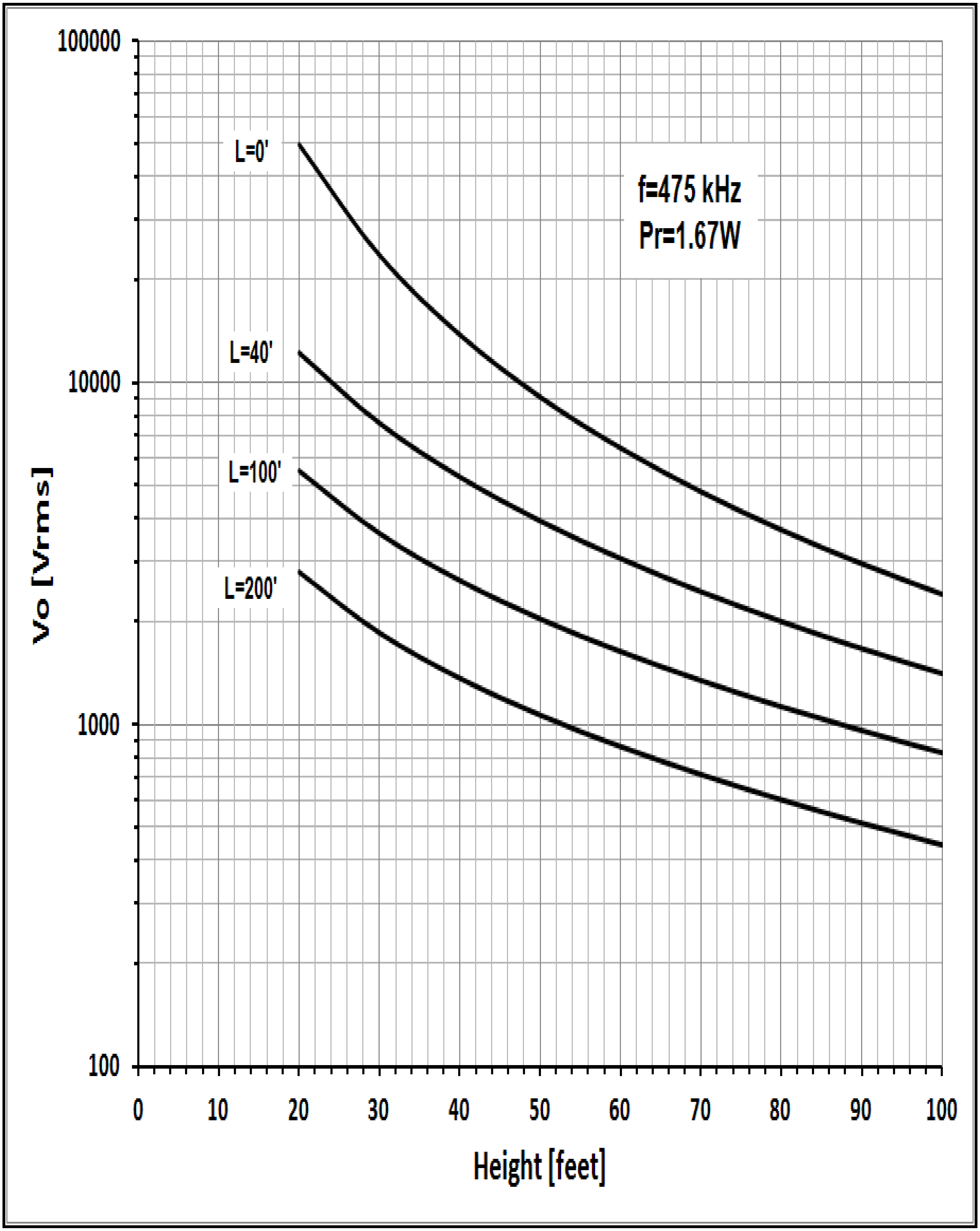

Figure 3.51 - Vo for Pr=1.67W at 475 kHz.

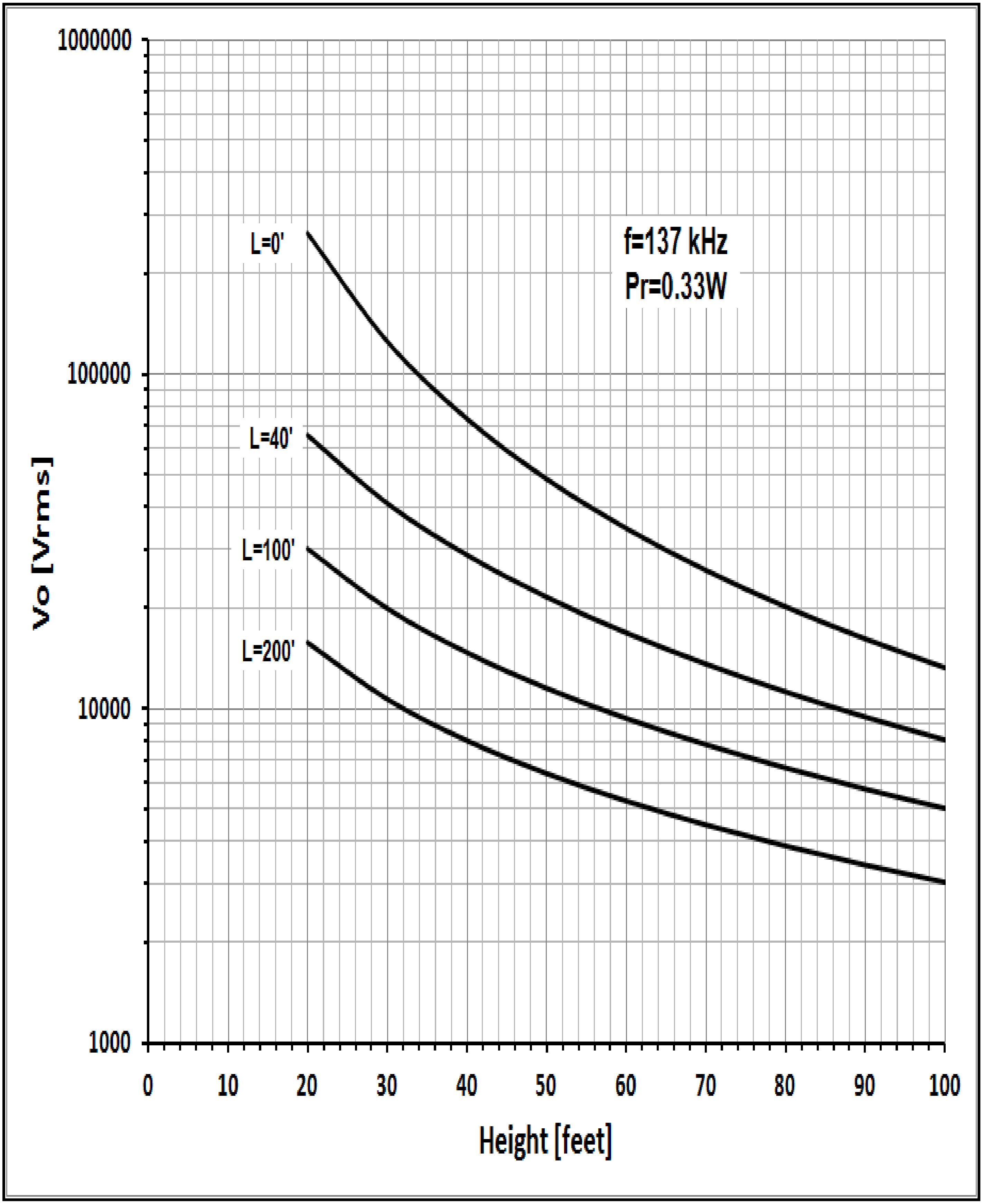

Figure 3.52 - Vo for Pr=0.33W at 137 kHz.

The input power to the antenna will include Pr, the radiated power, PL, the power dissipated in the loading inductor, Pg, the power lost the soil, plus conductor and leakage current losses. All of these loses are reduced substantially when Io and Vo are reduced. Pc was discussed in section 3.11, Pg is addressed in chapter 5 and PL is discussed in chapter 6. In general PL and Pg are the dominant losses, usually very much greater than Pr.

Summary

This chapter has shown many examples of top-loading arrangements, some simple, some more complex but all of them effective. The clear message throughout has been the great improvement in efficiency which capacitive top-loading can provide over a simple unloaded vertical. All the variations and discussion boil down to three practical points:

.... get as much wire as possible as high in the air as possible.....

...symmetry is not needed...

...even a small amount of top-loading is vastly better than none...

References

[1] Woodrow Smith, Antenna Manual, Editors & Engineers Ltd., 1948

[2] Grover, Fredrick, Methods, Formulas, and Tables for the Calculation of Antenna Capacity, NBS scientific paper 568, available on the NIST web site or use Google.

[3] Terman, Frederick E., Radio Engineers Handbook, McGraw-Hill Book Company, 1943

Chapter 4

Inductive Loading

4.0 Introduction

This chapter explores the use of inductive loading to increase the radiation resistance (Rr) and to enable excitation of a grounded tower. Chapter 3 demonstrated the utility of capacitive top-loading where the increase in Rr was primarily due to beneficial changes in the current distribution on the vertical. However, top-loading is not the only means for increasing Rr. We can move the tuning inductor or <u>even only a portion of it</u>, from the base up into the vertical. We can also move the feedpoint higher in the vertical. Multiple inductors can be useful in what referred to as "multiple-tuning", a technique using multiple inductors in multiple downleads to manipulate the feedpoint impedance and distribution of current between parallel wires.

4.1 Loading inductor location

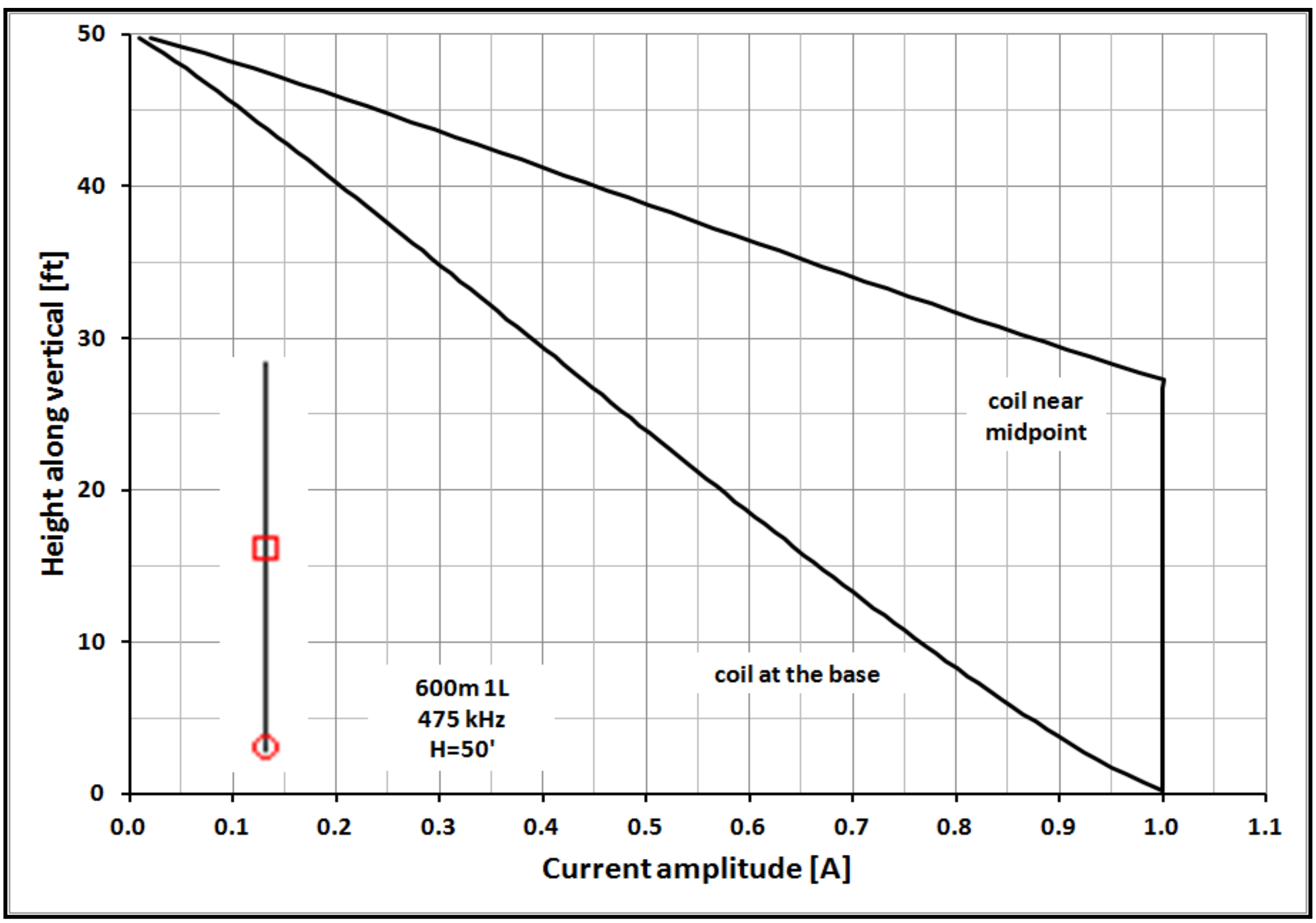

Figure 4.1 - Current distribution on a 50' vertical at 475 kHz.

In HF mobile verticals it has long been standard practice to move the loading inductor from the base up into the vertical to increase Rr[1]. We can do the same for LFMF verticals. Figure 4.1 compares the current distribution on a 50' vertical with the tuning inductor either at the base or near the midpoint. With the inductor near the midpoint the current below it remains essentially equal to Io. Increasing the current along the lower part of the vertical increases the Ampere-degree area A' (see section 3.3) which translates to increased Rr: 0.22Ω → 0.57Ω.

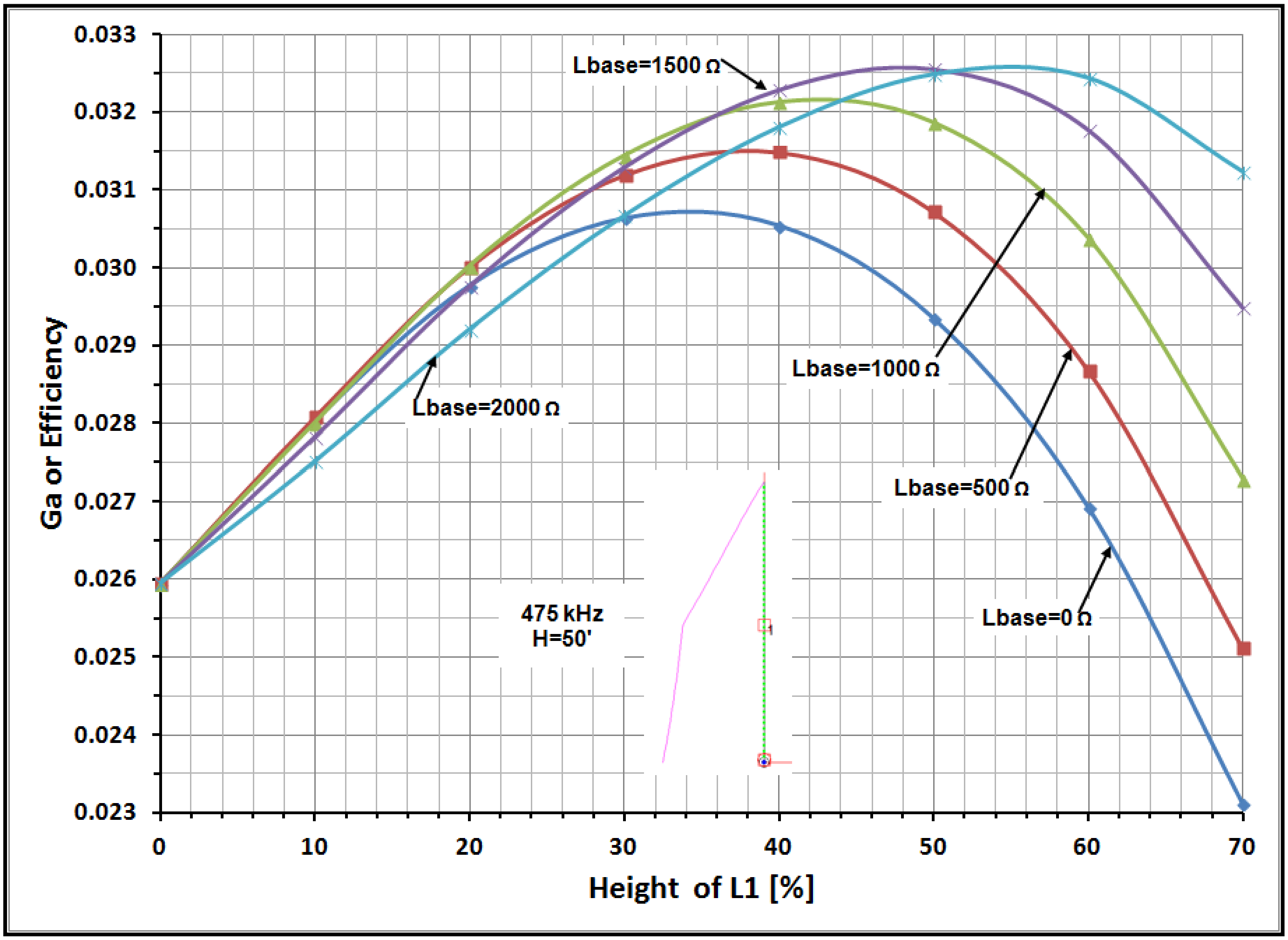

Figure 4.2 - Efficiency as a function of loading inductor location and value.

To keep the antenna resonant as the coil is moved up its impedance (XL) must be increased, 3411Ω → 6487Ω. Figures 4.2 and 4.3 show efficiency as the coil is moved higher. In this graph the horizontal axis represents the position of the loading inductor in percent of total height (H). The vertical axis is the efficiency in decimal form as a function of inductor placement. Traditionally the entire loading inductance is moved up. However, there are advantages to moving only a portion of the loading inductance up into the antenna and retaining the remainder (Lbase) at the base. In figures 4.2 and

4.3 the Lbase=0 contour represents the case where all of the inductance is moved up but there are also contours representing cases where Lbase remains substantial, from 500Ω to 2000Ω. Assuming the same QL for both inductors (QL=400) there can be some improvement in efficiency with divided loading (≈2%).

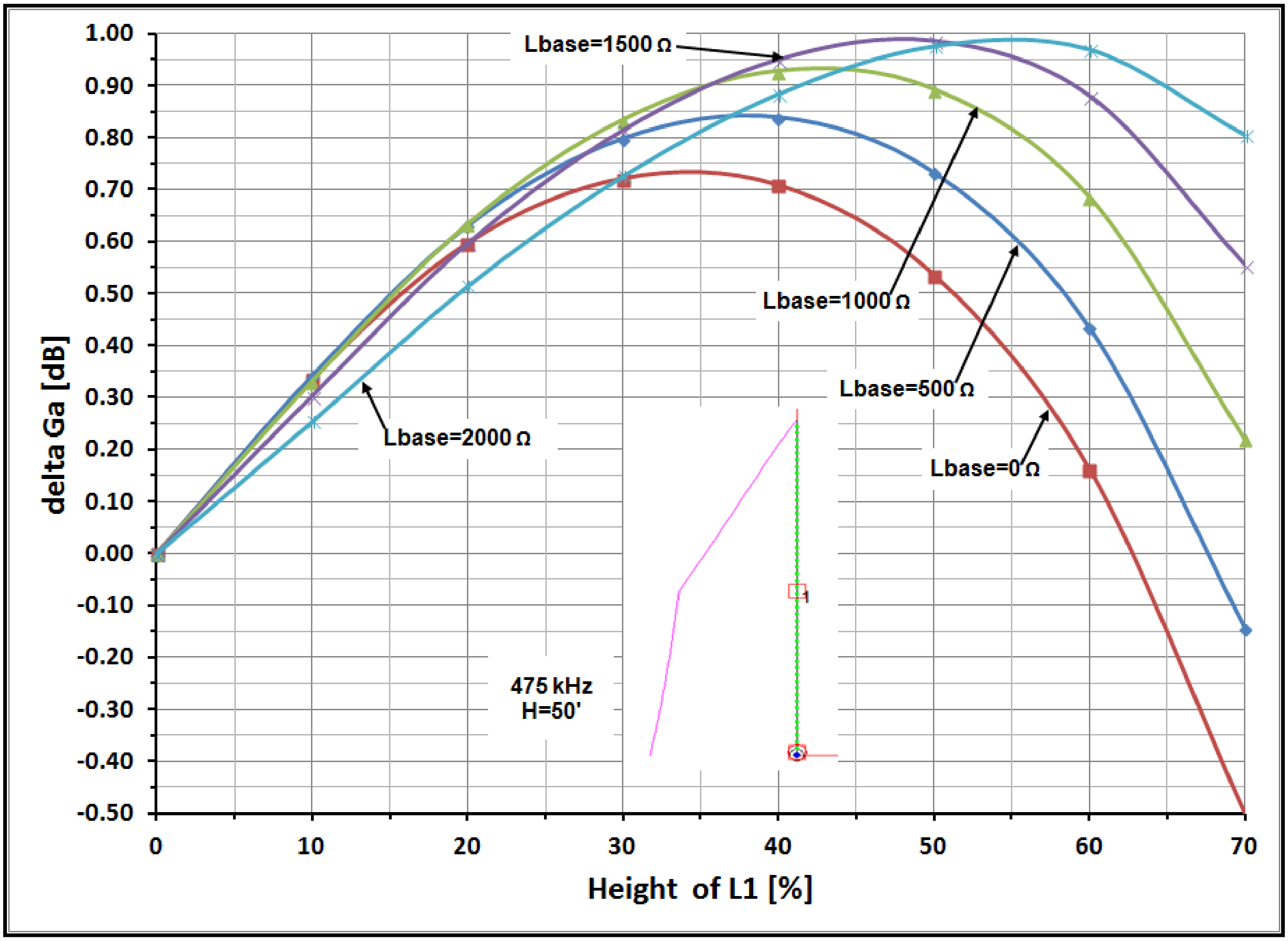

Figure 4.3 - Efficiency in dB.

Figure 4.3 converts the decimal efficiencies given in figure 4.2 to dB of signal improvement. Zero dB corresponds the case where all of the loading inductance is located at the base. For a given QL, RL will increase as the inductor is moved up. Despite this increase in RL moving the inductor up generally improves efficiency. The peak efficiency occurs for heights of 40 to 50% of H. How much does this increase our signal? For Lbase=0, i.e. we move all the inductance up, we can get about 0.74 dB of improvement at a height of ≈35%. By making Lbase=1500Ω we can pick up another 0.25 dB for a total improvement of almost 1 dB which is probably worth doing.

There is a simple trick for converting XL in ohms to L in µH: XL=2πfL, at 475 kHz $2\pi f_{MHz} \approx 3$ and at 137 kHz $2\pi f_{MHz} \approx 0.86$. For example at 475 kHz XL=6487Ω corresponds to 6487/3≈2,200µH or 2.2mH.

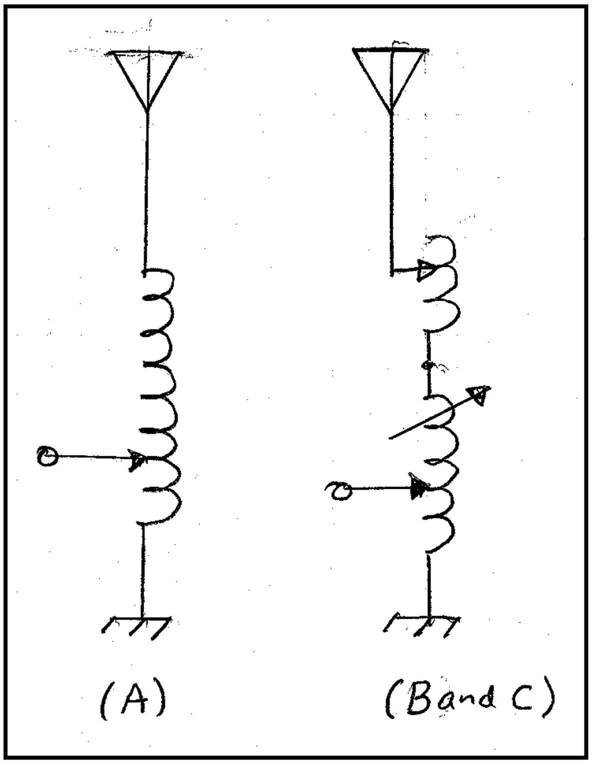

Figure 4.4 - Impedance matching with the base inductor

Even if a modest increase in signal is not compelling there are other reasons for using two inductors. Even when resonated it will still be necessary to match the feedpoint impedance to the feedline which can be done very simply by tapping the base inductance as shown in figure 4.4A.

A base inductor is also a convenient point to retune the antenna when necessary. Over the course of the seasons as the soil characteristics change, the tuning often shifts, primarily due to variations in effective loading capacitance as the soil conductivity changes with moisture content. Small heavily loaded verticals typically have very narrow bandwidths. In most cases some arrangement for adjusting the inductance will be needed. This can be readily done by using a variometer (figure 4.4B) (see chapter 6 for more details) or a separate small roller inductor in series. One additional advantage of not putting all the inductance up high is the reduced weight of the elevated inductor.

It is possible to use a long inductor for some or even all of the vertical. This is occasionally seen in mobile whips. EZNEC Pro v6 can model antennas constructed with a long helix. Figure 4.5 gives an example of a 50' vertical with a helix (coil) 24' long, 2' in diameter, with150 turns of #12 copper wire. The bottom of the coil is at 2' and the top of the coil at 26'. EZNEC gives an efficiency of ≈4.1% which is somewhat better than a single concentrated load at 0.35H (figure 4.2, QL=400).

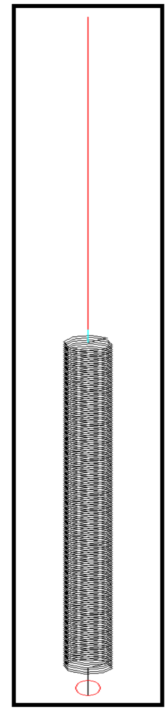

Figure 4.5 - A 475 kHz vertical with distributed loading inductor.→

4.2 Inductor location with top-loading

Due to windage considerations mobile verticals seldom have much capacitive top-loading but fixed station antennas have (or should have!) as much top-loading as practical.

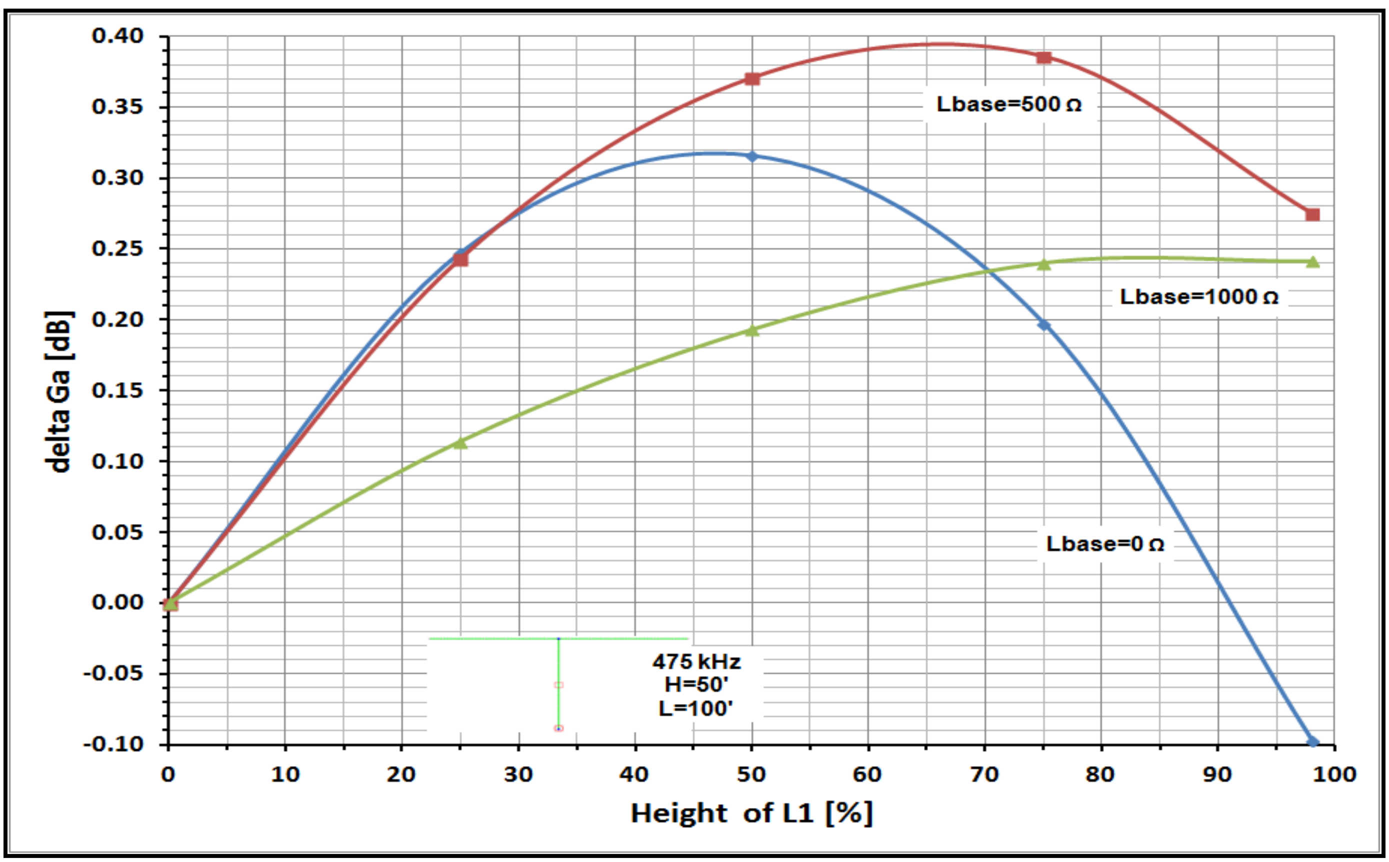

Figure 4.6 - Efficiency as a function of loading coil position.

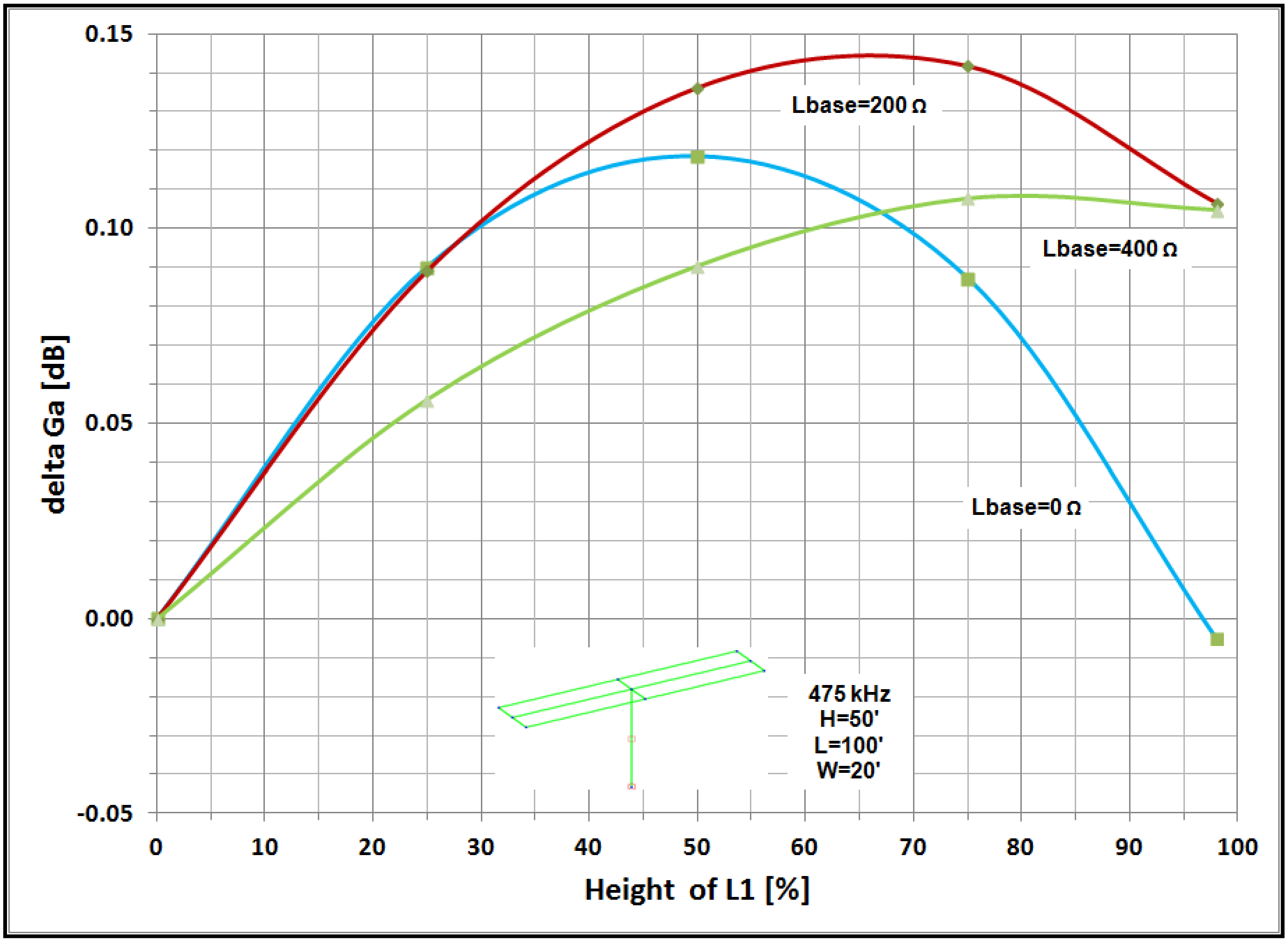

Figure 4.7 - Efficiency versus loading coil position with heavy top-loading.

Moving the inductor higher into a heavily top-loaded vertical has less effect on the current distribution. As a result efficiency improvements are much smaller. Figure 4.6 gives an example of a T antenna with H= 50' and a single 100' top-wire. As the inductor is moved up there is some improvement in the signal but not a lot, only 0.4 dB even when two inductors are used.

As shown in figure 4.7, when we have a much larger top-hat (three wires 100' long by 20' wide) the improvement from elevating the inductor is even smaller, <0.15 dB This small an improvement is not worth the hassle of mounting an inductor high in the antenna! The reason for the very small improvement can be seen in figure 4.8 which shows the current distribution for various loading inductor heights. Even without moving the inductor up, It/Io is almost 0.85 (It=current at the top of the vertical section). Moving the inductor up increases A' but not by very much.

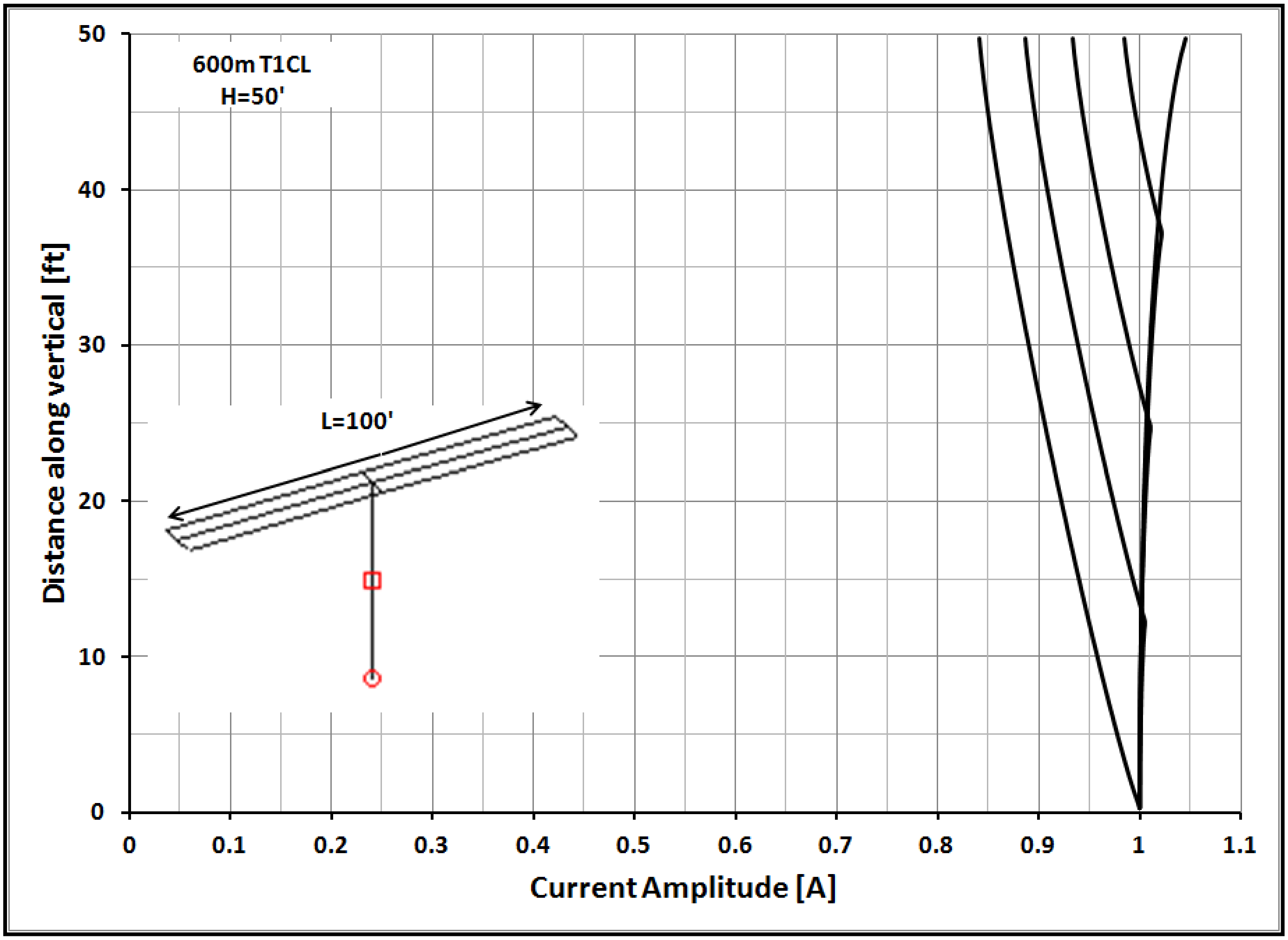

Figure 4.8 - Current distributions on a top-loaded vertical for various loading inductor heights at 475 kHz.

In heavily top-loaded verticals there appears to be little improvement in efficiency from elevating the loading inductor. On the other hand if the top-loading is less, It/Io ratio <0.4-0.5, and more top-loading is not practical then moving the coil up may help. This has to be evaluated on a case-by-case basis using modeling.

4.3 Grounded Tower Verticals

A grounded tower with attached HF antennas and associated cabling is sometimes available. For an LF-MF antenna the tower may be simply a support but it can also be a radiator. One way we might do this is shown in figure 4.9 where the loading inductor <u>and the feedpoint</u> have been moved to the top of the tower. The top-loading wires are insulated from the tower and connected to one end of the loading inductor. The other end of the inductor is connected to the top of the tower. A coaxial feedline runs up the tower with the shield connected to the top of the tower. The coax center conductor is connected to a tap on the loading inductor to provide a match. Although not shown, it is possible to have a mast with HF Yagis extending above the top of the tower which will add some additional capacitive loading. The downside of this scheme is that all the adjustments must be made at the top of the tower.

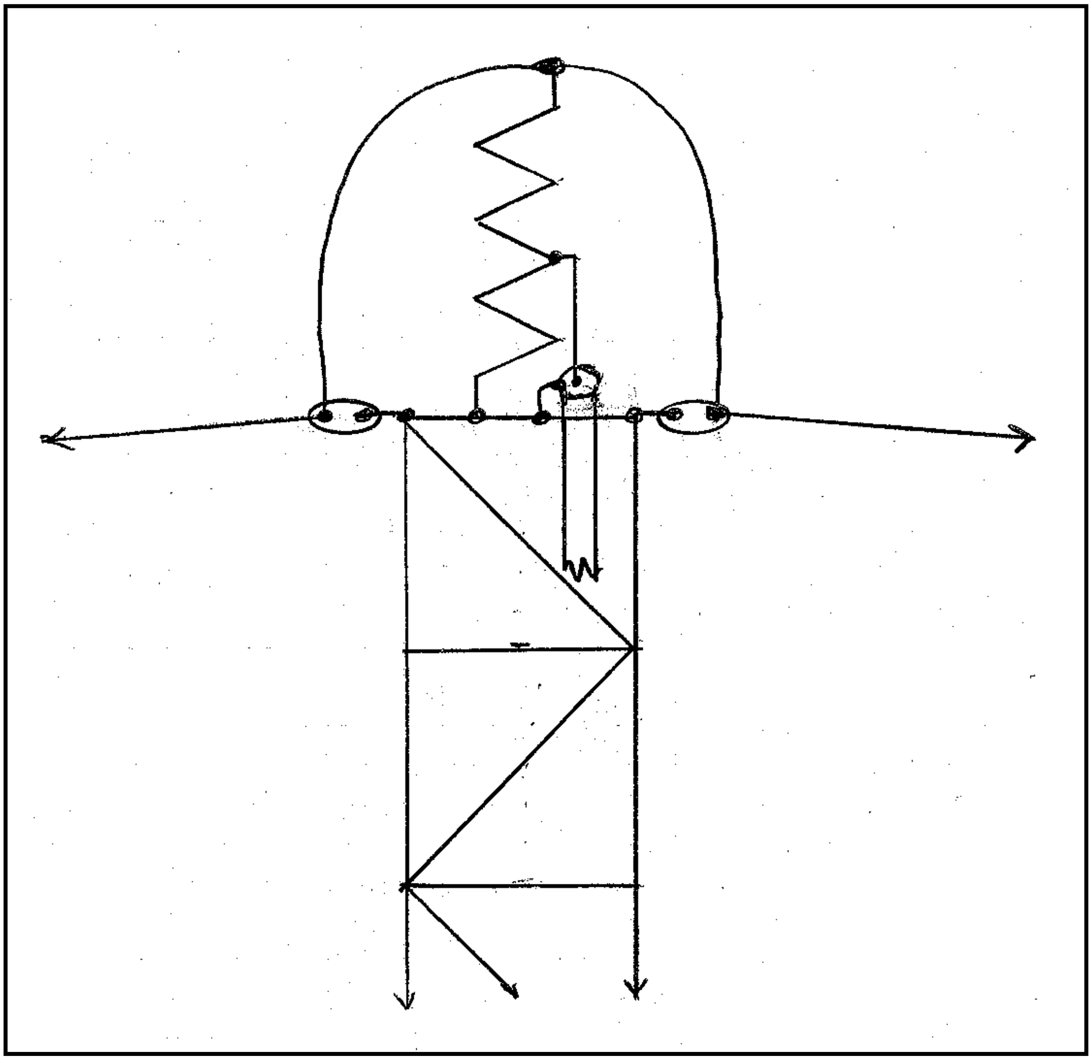

Figure 4.9 - Grounded tower, feedpoint and loading inductor at the top.

A common alternative for exciting a grounded tower often used on 80m and 160m is the shunt-fed tower shown in figure 4.10. Unfortunately this scheme works only if H>0.7 λ/4 at the operating frequency. At 475 kHz that would mean H>350', much taller than most amateur towers. The causes and cures for problems associated with this configuration are worth discussing in some detail because similar problems can arise whenever multiple downleads are used, which is quite common in LF-MF antennas.

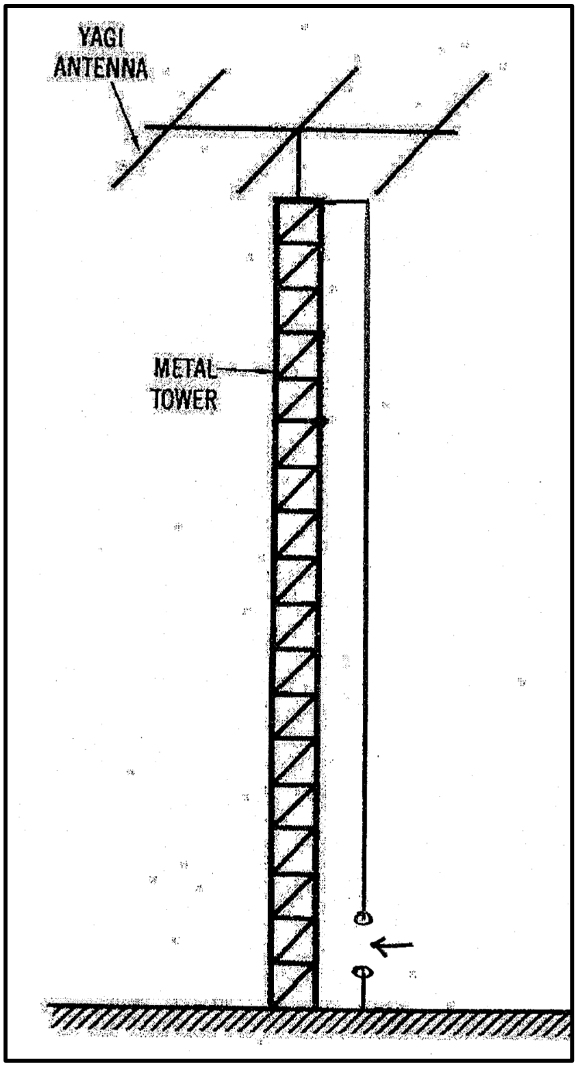

Figure 4.10 - Shunt fed tower.

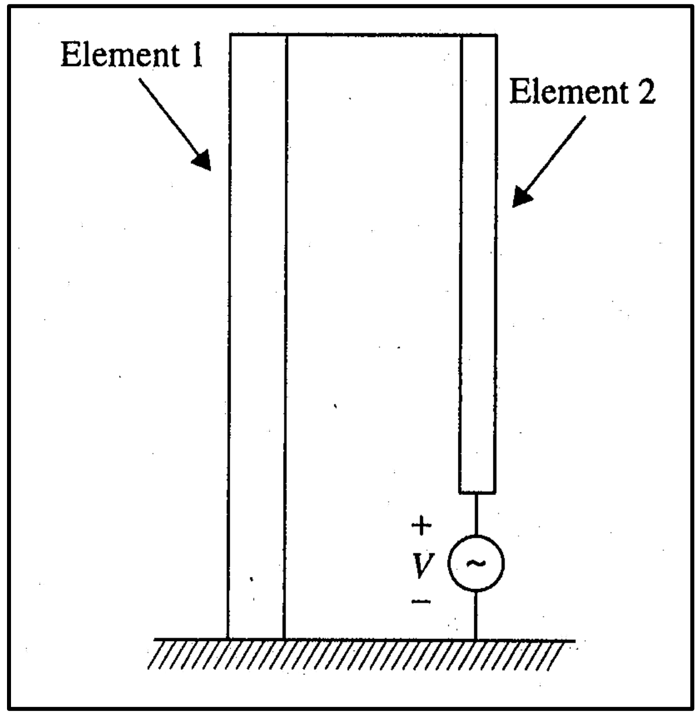

Figure 4.11 - Monopole antenna. From Raines[6]

The arrangement illustrated in figure 4.10 is usually considered to be an impedance matching scheme but the tower and the shunt wire is actually a member of a family of antennas called "folded monopoles", as shown in figure 4.11. One of the important properties of folded monopoles is that the feedpoint impedance can be a multiple of that for a single element vertical of the same height. For example, if two elements are used and both elements have the same diameter, the Zi will be 4X that for a single element. Even more elements can be added to further increase Ri. It is also possible to use elements of different diameters which can lead to arbitrary Ri ratios. All of this however, applies <u>only</u> when the height is close to λ/4. When we shorten the antenna to heights practical on 475 or 137 kHz the antenna behavior is quite different.

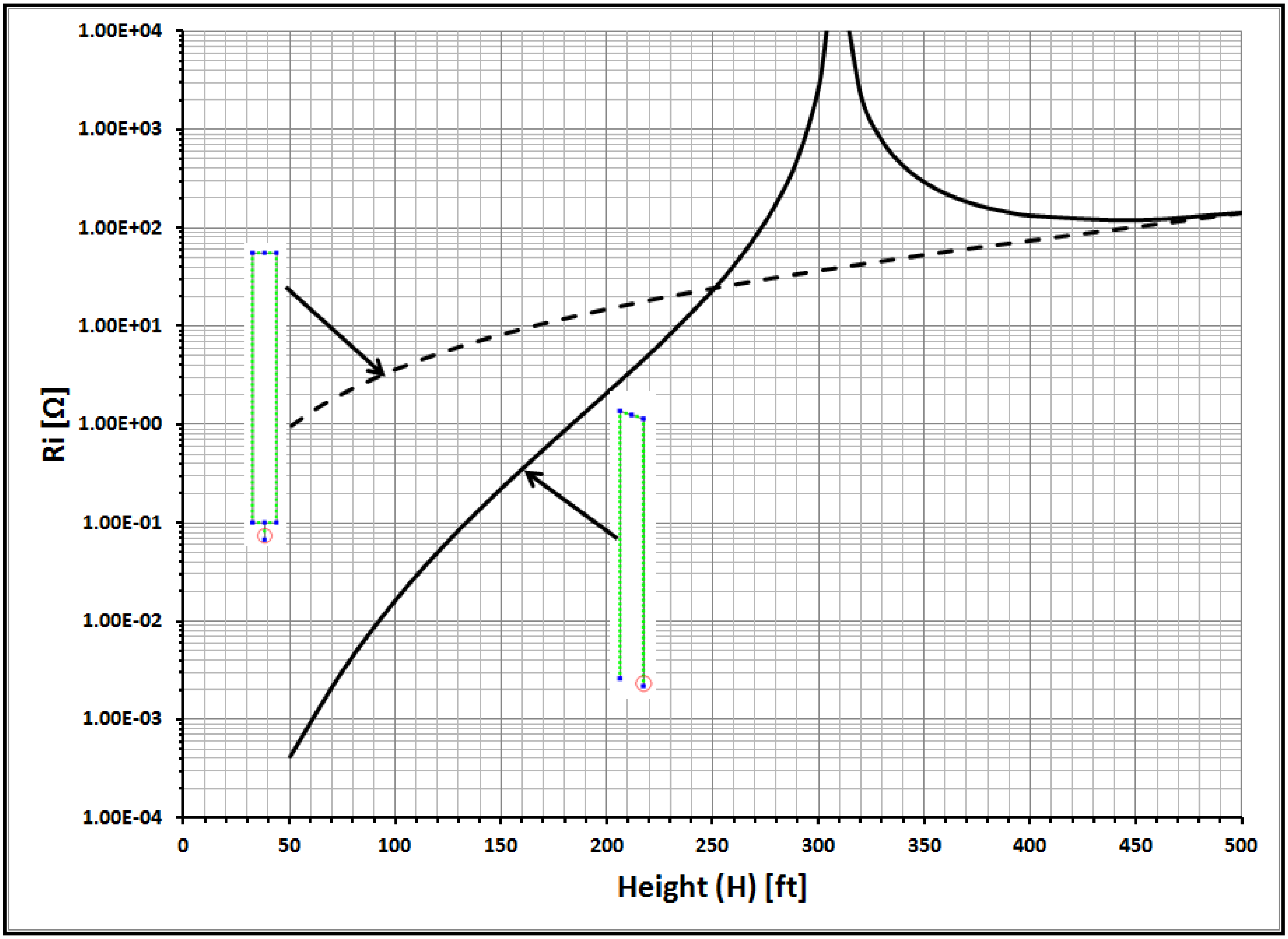

Figure 4.12 - Ri versus height for normal and folded monopoles.

Figure 4.12 compares the values of feedpoint Ri between a normal monopole (dashed line) and a folded monopole (solid line) at 475 kHz for a wide range of heights. Ri for the normal monopole has been multiplied by 4X for this comparison. A similar graph for Xi is given in figure 4.13. In both graphs we can see that for heights down to ≈400' Zi ≈ 4X as predicted but as we go to shorter heights there is a rapid divergence between the behavior of the two antennas. Our interest is primarily H<100' where the folded monopole impedance is very small compared to the normal monopole.

What's going on and can we fix it?

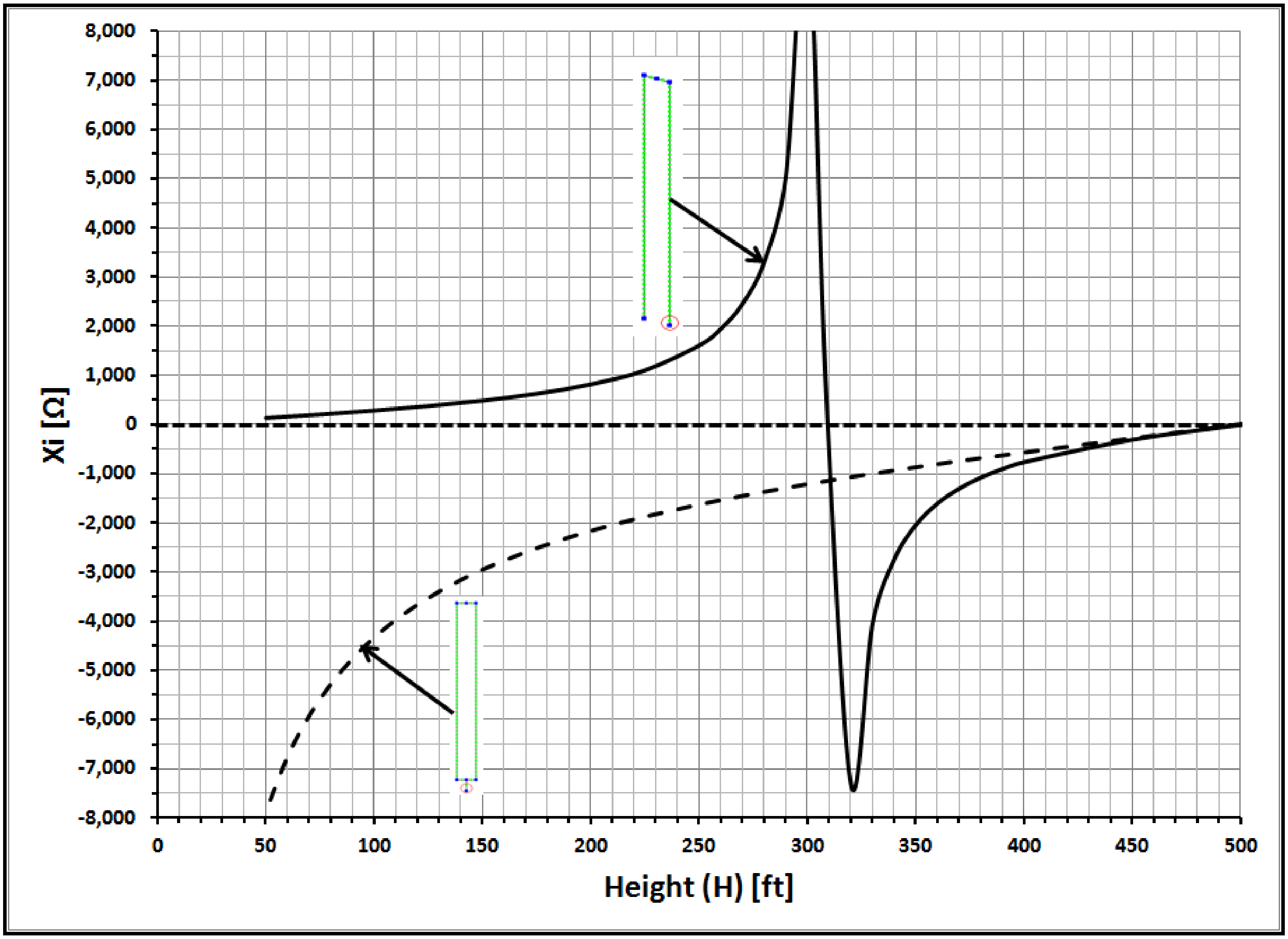

figure 4.13 - Xi versus height for normal and folded monopoles.

A folded monopole can be viewed as a superposition of a vertical and a transmission line shorted at the far end as shown in 4.14[5, 6, 7, 8]. There are two currents, a radiating "antenna current" =2X(Ia/2) and a circulating "transmission line current" $=I_T$. A lumped element equivalent circuit is shown where Ra+jXa represents the vertical and $+jX_T$ represents the inductance appearing across the feedpoint due to the transmission line. The value for jX_T can be found from:

$$jX_T = jZo\ Tan(H)$$

Where Zo is the characteristic impedance of the transmission line which depends on element spacing and diameters. H is the length in degrees or radians.

For the "monopole" Xc rises quickly as H is reduced (chapter 2) but the "transmission line" X_T falls rapidly as H is reduced. The result is shown in figure 4.13, at H≈310',

$Xa=X_T$ and we have a parallel resonance with very high input impedance. It can also be shown that at this point there are very large circulating currents leading to large losses.

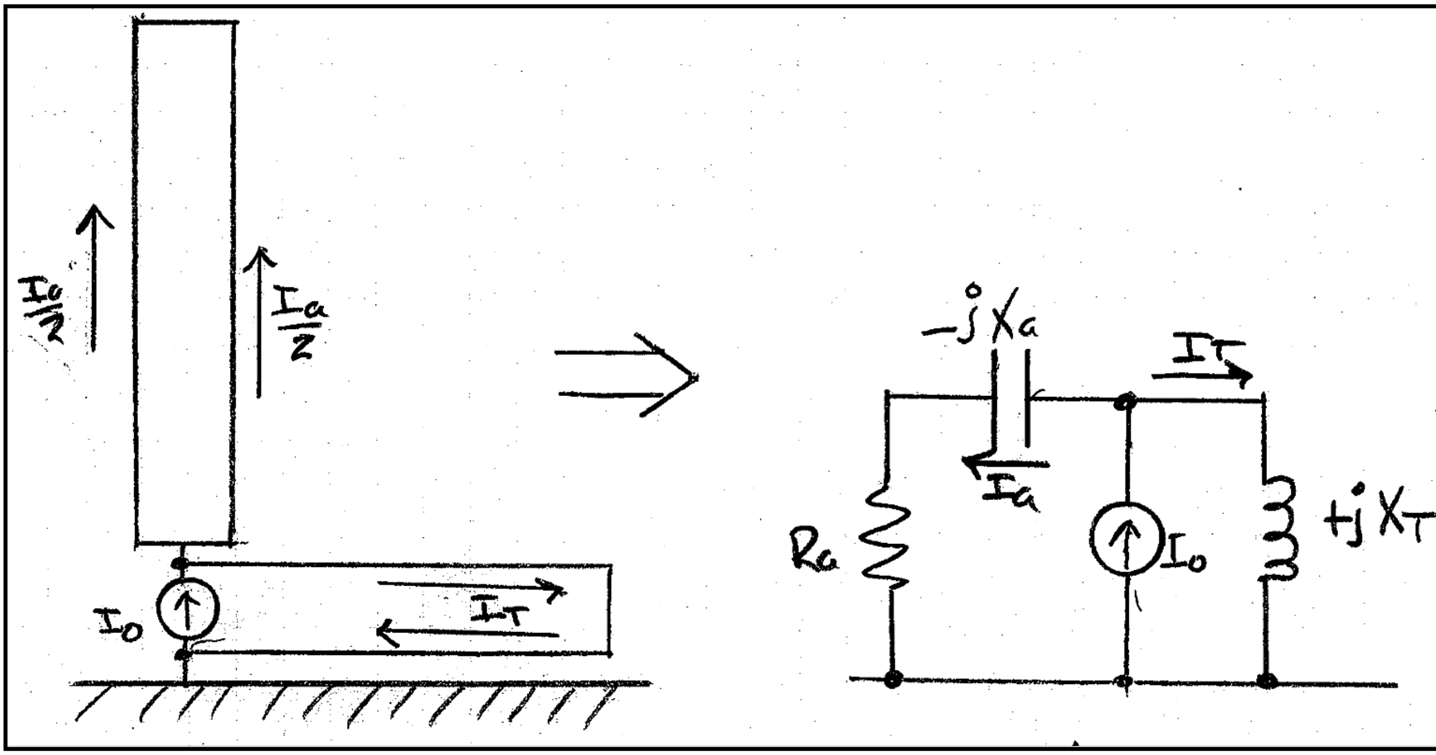

Figure 4.14 - Folded monopole vertical equivalent circuit.

Below the parallel resonant point the amplitude for X_T becomes much smaller that Xa and the antenna is basically just a tall narrow loop inductor with very little radiation!

Now we might ask if some top-loading would help? Figure 4.15 shows what happens when substantial top-loading is added (two parallel wires 200' long) . Certainly there is some improvement but the basic problem is still there, X_T still dominates Zi and there is a large circulating current in the "transmission line"

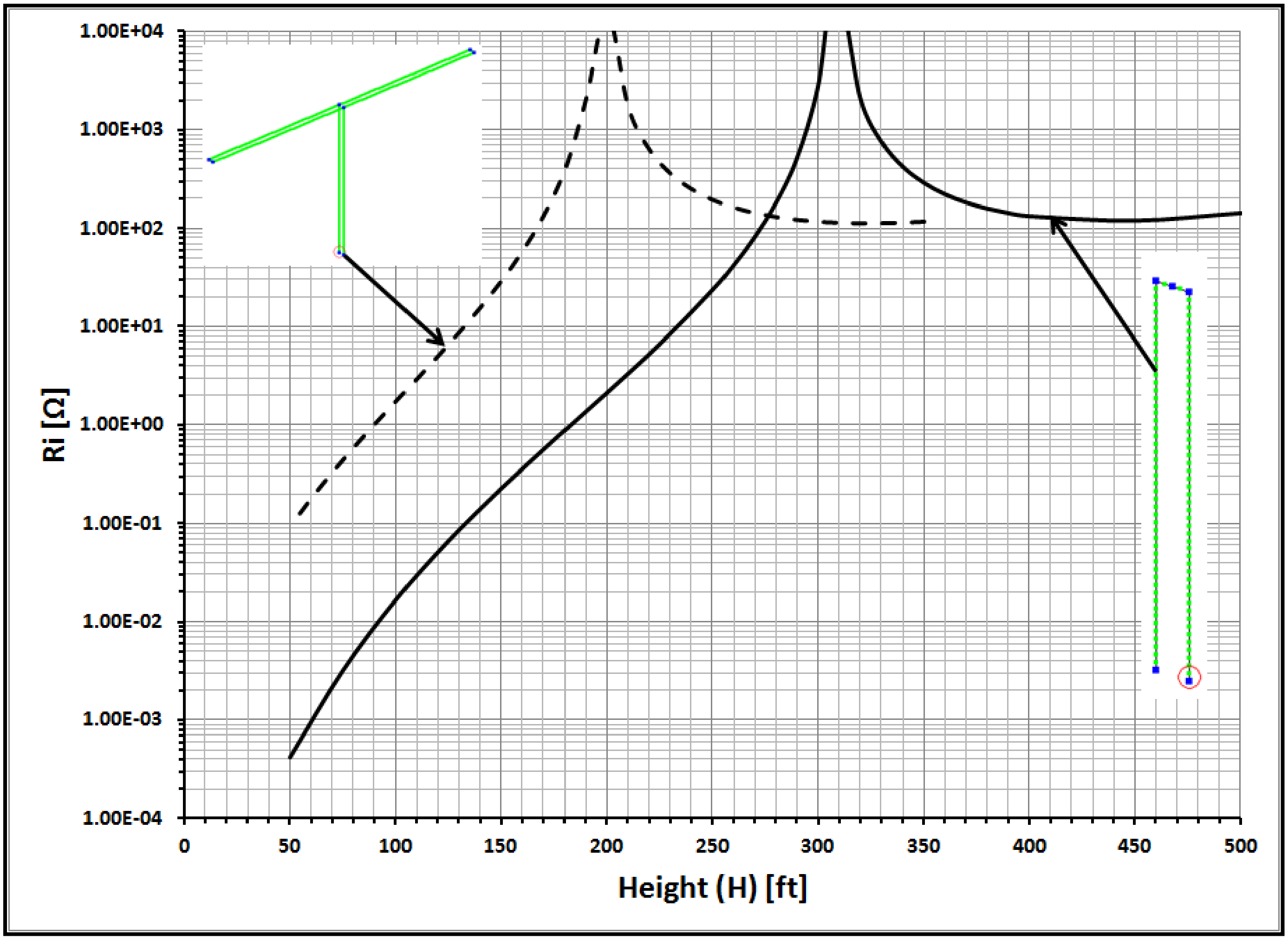

Figure 4.15 - Ri versus height for normal and top-loaded folded monopoles.

Figure 4.16 shows we might eliminate the effect of X_T by adding series inductance (XL). There are a couple of points at which we might add inductance: at the grounded end of element 1 or at the top of the antenna or both. Since our concern here is with a grounded tower we will place an inductor at the top of the antenna. Other possibilities are discussed in the section on multiple tuning. If XL>> X_T then the transmission line impedance is in effect constant and has a value of our choosing. Normally we would choose XL to resonate the antenna which means it's value will be large compared to X_T for typical amateur antenna heights. Figure 4.17 shows an application of this idea with part of the loading inductor at the top and the rest at ground level to make adjustment and matching convenient. Even though we've overcome a problem with short folded monopoles, we still have a very short vertical so in addition to the inductive loading we still need capacitive top-loading as shown to minimize the inductance. The top-hat wires are insulated from the tower as they were in figure 4.9.

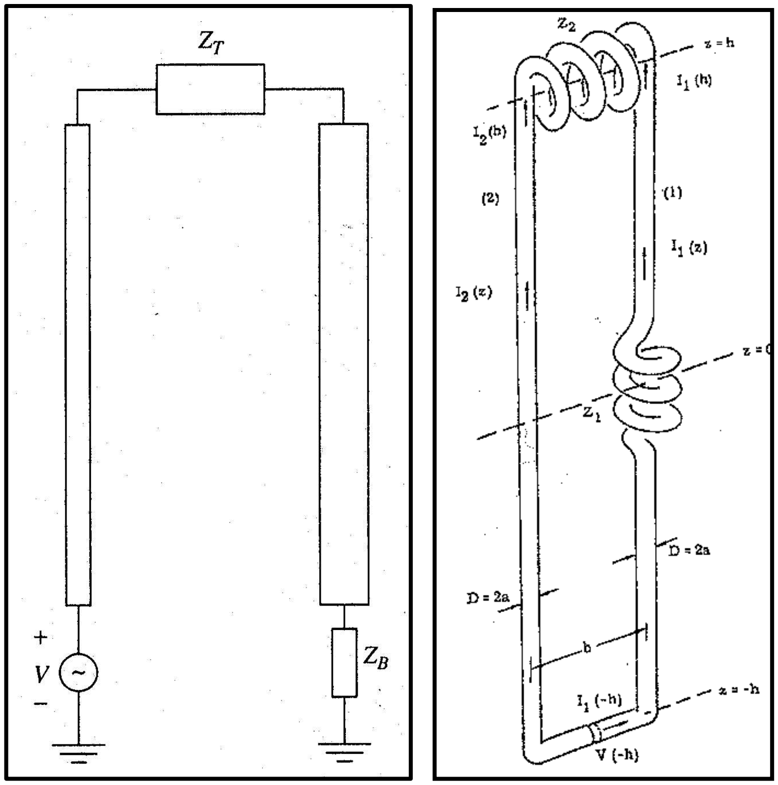

figure 4.16 - Alternate impedance placement. From Raines[tbd] & Harrison & King[5].

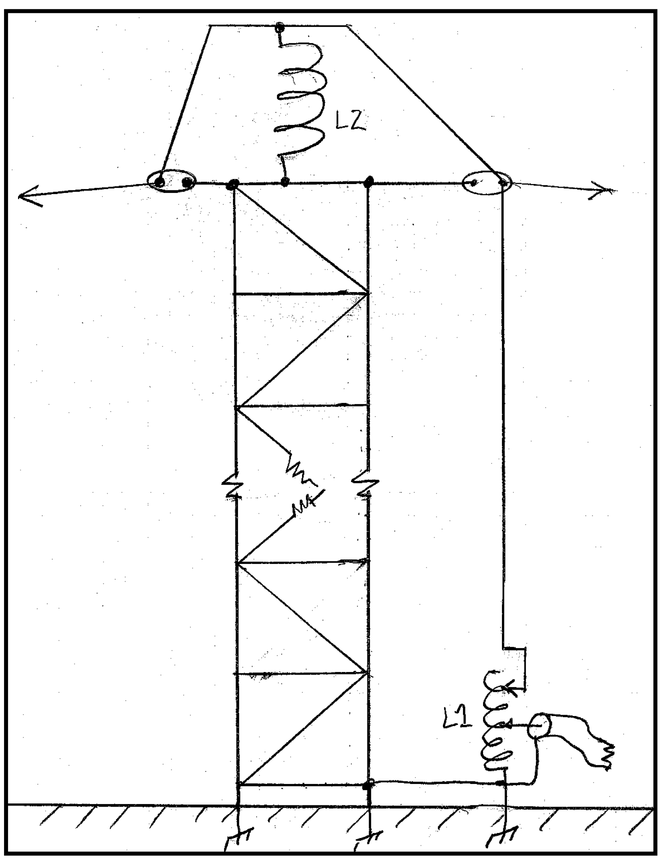

Figure 4.17 - Grounded tower feed scheme using two inductors.

4.4 Multiple Tuning

As suggested in figure 4.18, we can have multiple vertical elements, 1,2,...,m, n,... with inductors in series with each downlead/vertical.

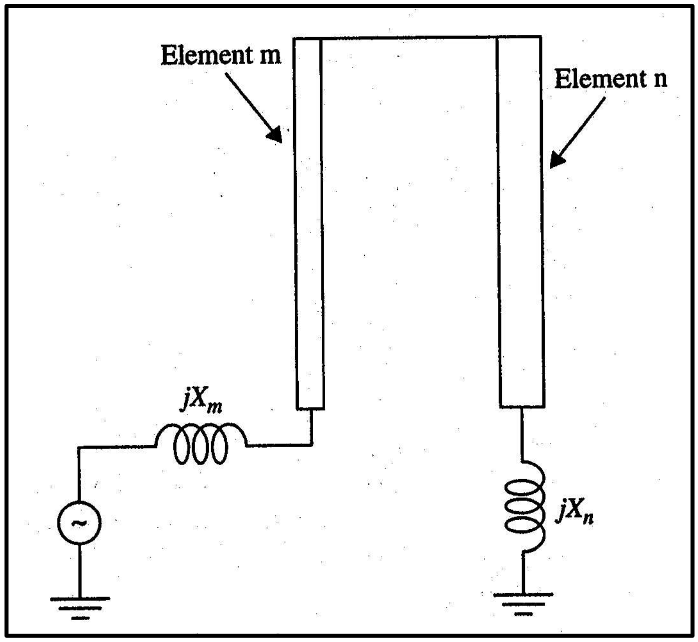

Figure 4.18 - Adding tuning inductors to a short folded monopole. From Raines[6]

Commercial LF antennas have long used multiple inductors to advantage. An example taken from Laport[2] is shown in figure 4.19.

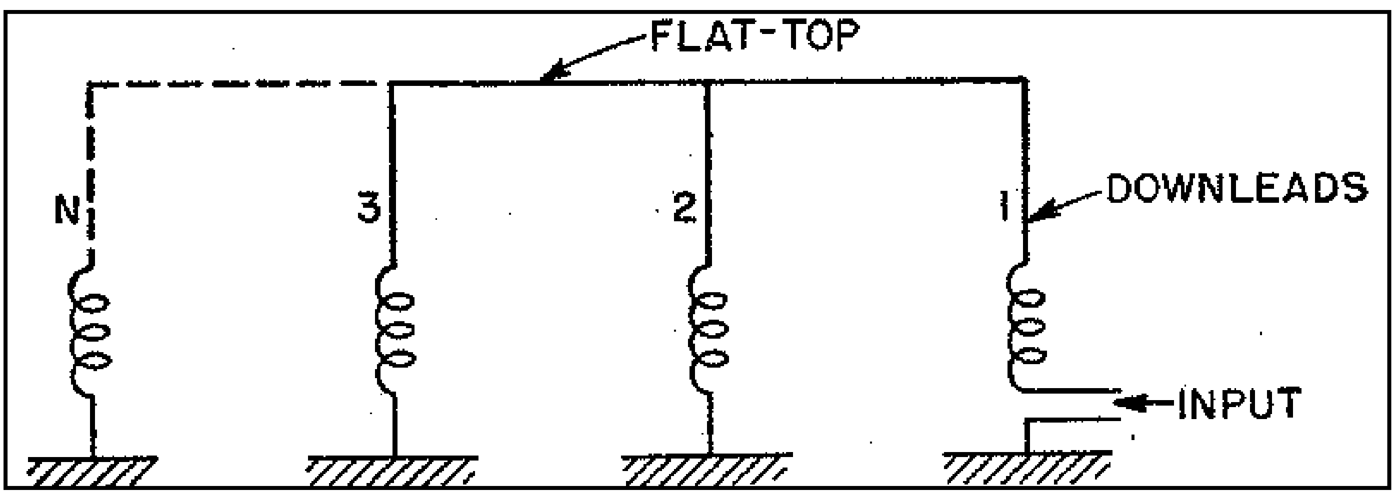

Figure 4.19 - Example of multiple tuning.

The following is a quotation from Laport:

"The most extreme conditions of low radiation resistance and high reactance are encountered at the lowest frequencies, and some extreme

measures are necessary to obtain acceptable radiation efficiencies. .... The most successful method of improving the radiation efficiency is that of multiple tuning. The antenna consists of a large elevated capacitance area with two or more down leads that are tuned individually as indicated in the figure. The total antenna current is thus divided equally among the down leads, each of which has its own ground system. The down lead currents are in phase, and because of their electrically small separation there is no observable effect on the radiation pattern, which is always nearly circular. Power is fed into the system through one of the down leads.

When arranged for multiple tuning, an antenna behaves as a number of smaller antennas in parallel, voltage being fed through the flat-top system. Thus, a system with triple tuning is essentially three antennas in parallel, one of which is fed directly by coupling to the transmitter and the other two at high potential (voltage feed) through their common flat-top. From a radiation standpoint, the same effect would be realized if the different portions of the antenna were not physically connected through their common flat-top but instead were separately fed from a common transmitter and feeder system in the manner of a directive antenna. Practically it is simpler to take advantage of the fact that almost all antennas for the lowest radio frequencies must of necessity employ flat-tops for capacitive loading and merely to add the extra down leads for multiple tuning. In that way, there is only one feed point, and the problems of power division, phasing, and impedance matching are automatically minimized. .....

If N represents the number of multiple tuning down leads carrying equal currents, the new radiation resistance Rrr is related to that for single tuning by the equation

$Rrr=RrN^2$. ...."

In a short LF-MF antenna this scheme can provide feedpoint impedances which are much easier to deal with. Laport goes on to suggest that, for equal QL in all inductors, the total inductor loss will be reduced with multiple tuning but this does not appear to be correct. Modeling shows that the losses are essentially the same. However, ground system losses with multiple grounds may be less.

As shown in chapter 2 (equation 2.8), a vertical can be viewed as a single wire transmission line with an average characteristic impedance[3] Za. The top-loading can be viewed as a voltage source which allows us to model the antenna as a transmission line with a voltage source at the top and a short-circuit termination at the bottom as shown in figure 4.20A.

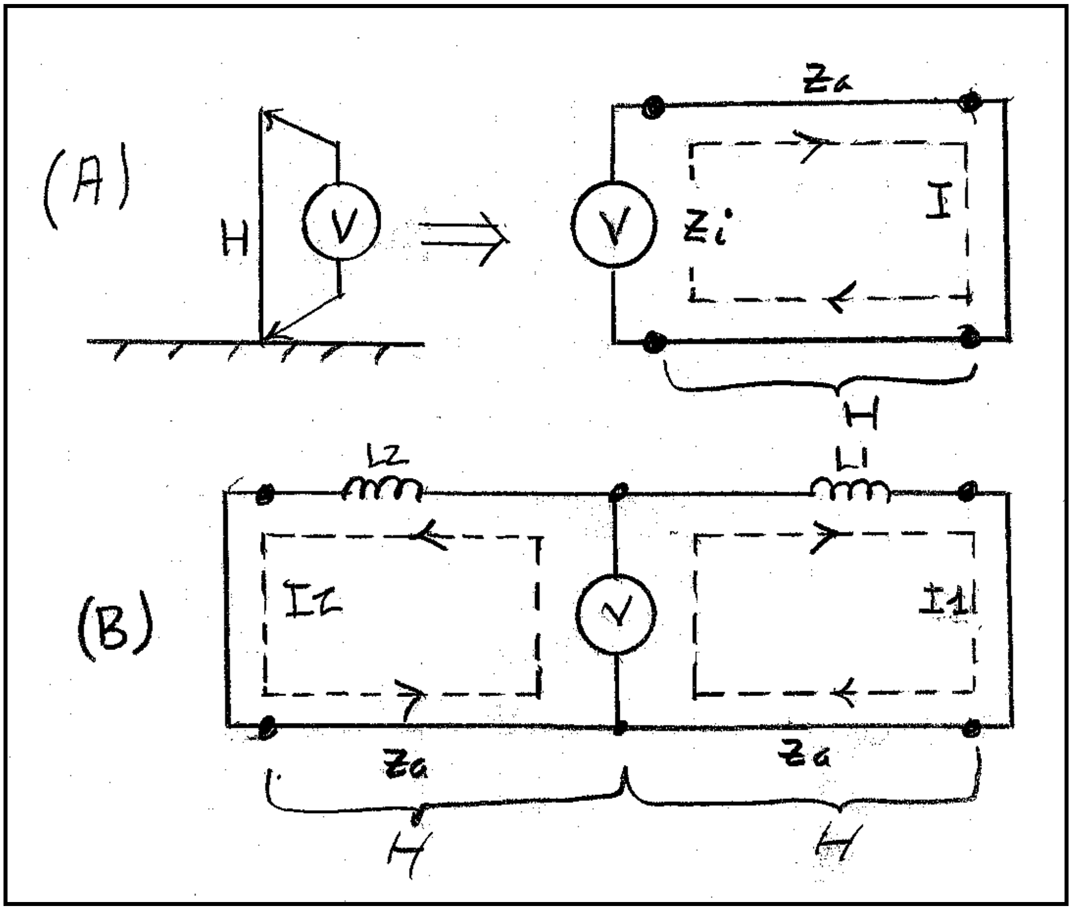

Figure 4.20 - Parallel down-lead equivalent circuit.

The input impedance of a S/C transmission line fed at the top is:

$$Zi = jZa\,Tan(H)$$

Figure 4.21 is a graph of Zi at 475 kHz as a function of H for #12 wire and 2" diameter tubing. H=500' is close to λ/4 resonance. From the graph we can see that Zi drops rapidly as the vertical is shortened below λ/4. At H=500', Zi≈10kΩ but at 100', Zi≈100Ω with Zi falling rapidly below H=100'. Given most amateur antennas will have H<100', Zi will be much smaller than the value of XL needed for resonance so that the impedance of the vertical(s) when excited at the top is almost entirely determined by the loading inductance. The multiple downleads shown in figure 4.19 are equivalent to multiple transmission lines in parallel as shown in figure 4.20B. To control the current distribution between the multiple downleads we insert inductors. The currents may all be the same as or different. Non-equal current distributions can be used to modify the feedpoint impedance, i.e. if you insert more inductance in the driven downlead, the current that download will be reduced and the feedpoint impedance increased. You will have to readjust the other inductances to re-resonate the antenna however!

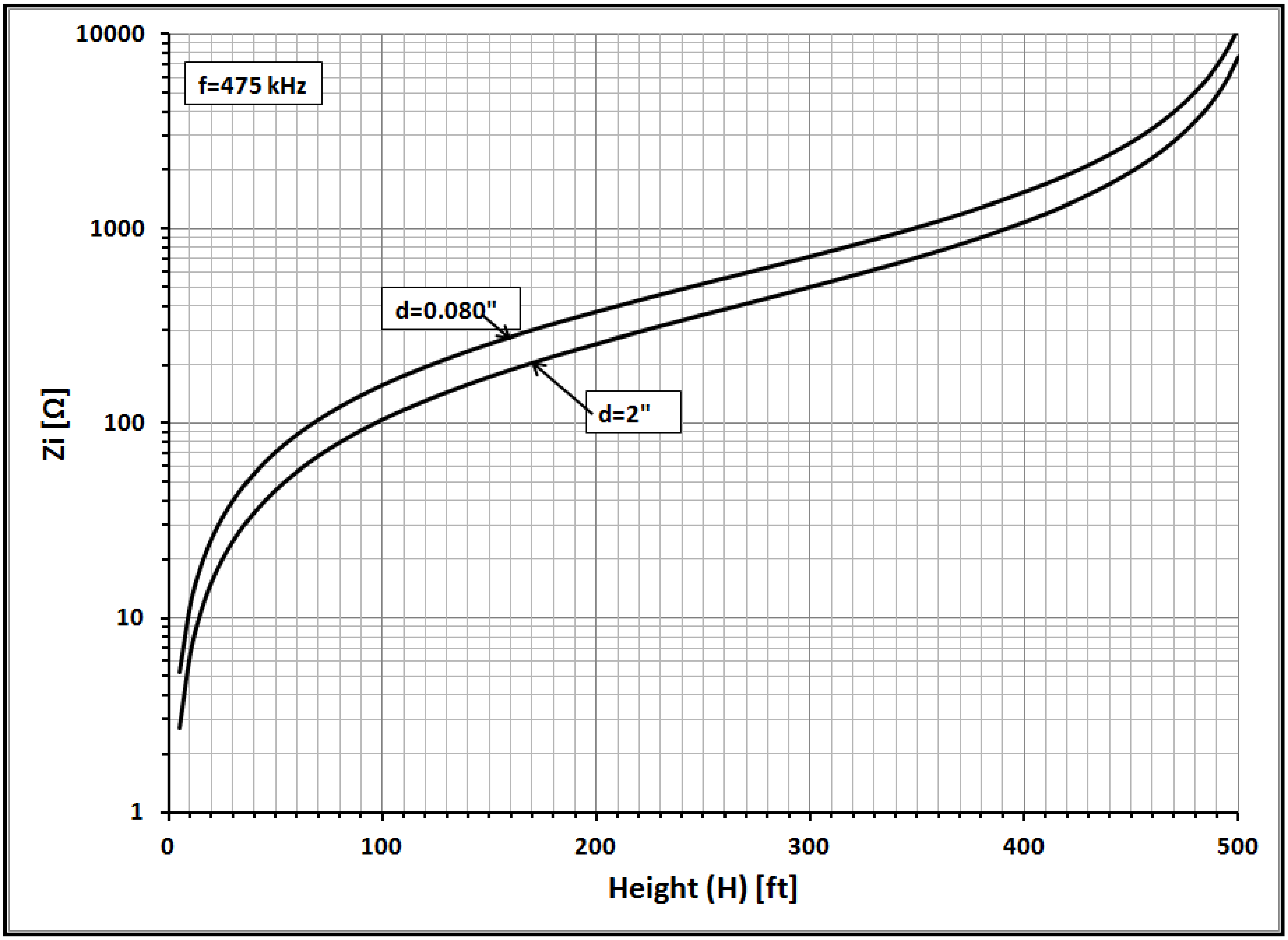

Figure 4.21 - Zi at 475 kHz versus height.

One other important point, Terman[4, page 841] has a comment on minimum top-hat capacitance which applies to multiple tuning:

> ***"...the flat-top capacitance should be considerably greater than the capacity of the vertical downlead....."***

There has to be enough capacitance so that the impedance of the "voltage source" (i.e. the top-loading) is low compared to Zi.

4.5 Multiple tuning examples

As a reference point we can start with the T antenna shown in figure 4.22.

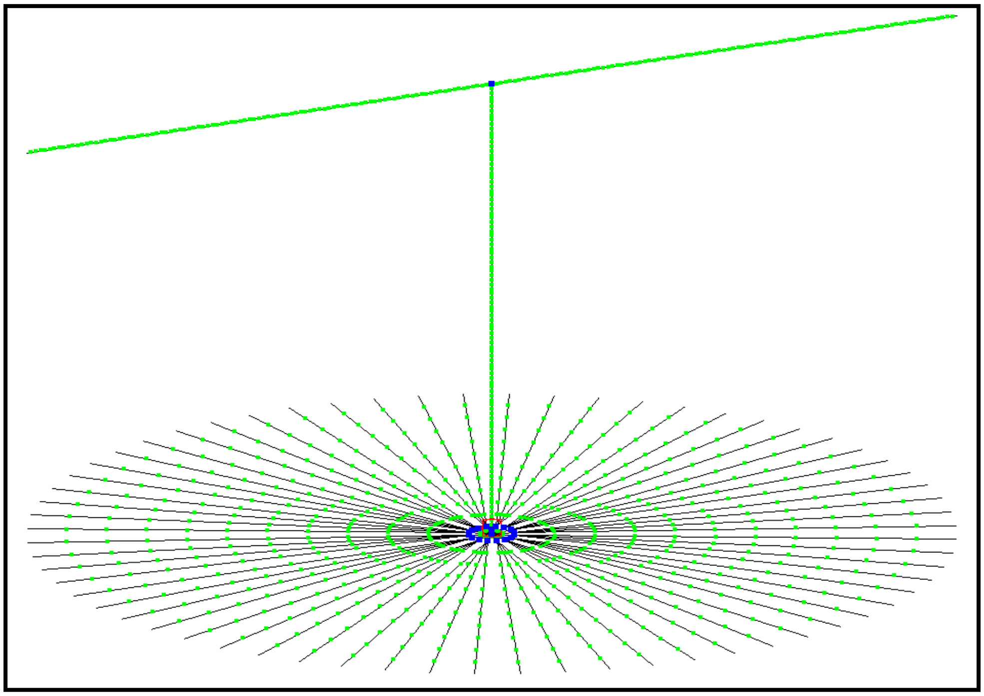

Figure 4.22 - T-antenna example.

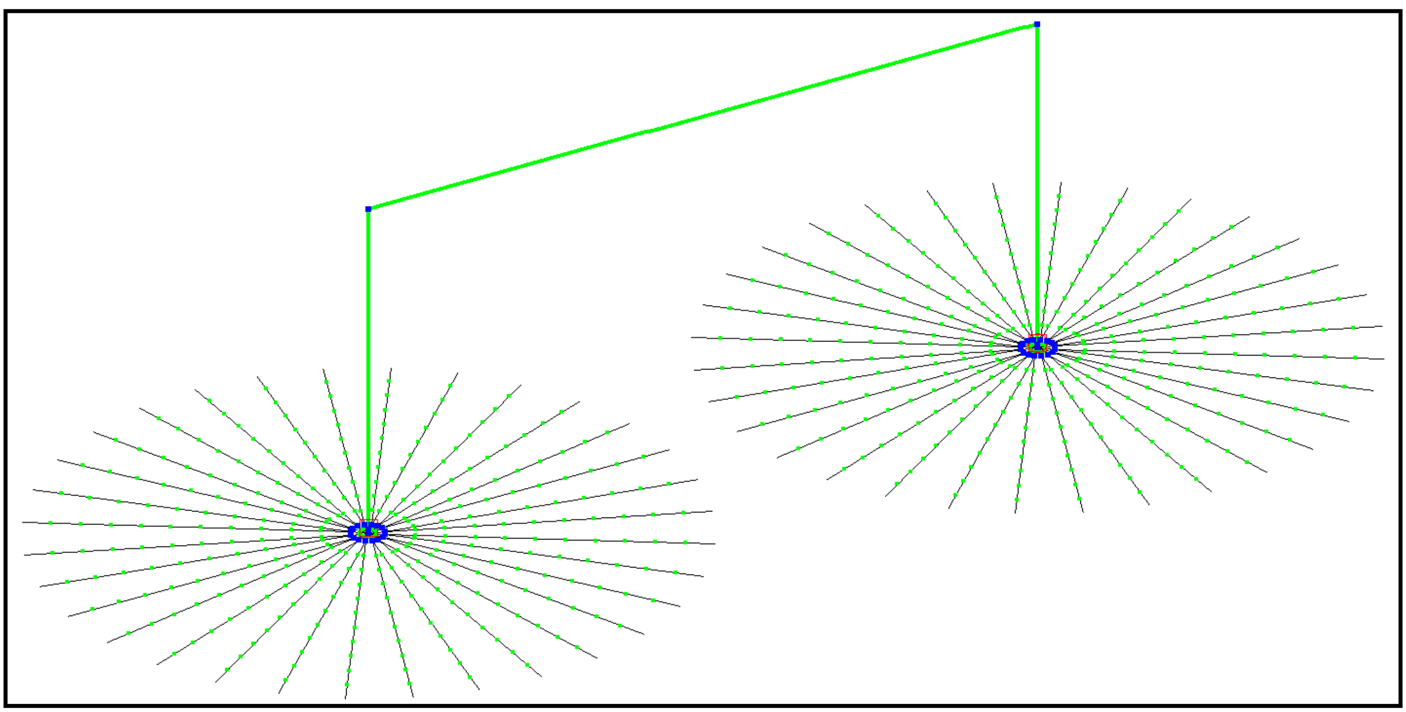

Figure 4.23 - Antenna with two downleads and a loading inductor at the base of each download.

The antenna and the radials are #12 copper wire. H=50' and each top-wire is 50' long. There are sixty four 45' radials buried 6" in average soil (0.005/13). For resonance at 475 kHz, XL=1239Ω (≈410µH). Including copper (Rc), RL and soil losses (Rg), the feedpoint resistance Ri is ≈6.21Ω and the radiation efficiency is ≈10.0% or -10dB. Suppose we use the same top-wire (100') but introduce multiple tuning with two downleads, one at each end as shown in figure 4.23.

For each download H=50' and the top-wire is 2x50'=100'. The same total amount of wire is used in the new ground system, i.e. each downlead has thirty two 45' radials. For resonance at 475 kHz, XL1=XL2= 1857Ω (≈620 uH). Ri≈16.3Ω and the efficiency is about 11.9% or -9.24 dB which represents a signal improvement of +0.76 dB. As predicted going from one to two downleads the current in each downlead is Io/2 and Ri is increased by a factor of four. There is some improvement in efficiency, ≈2%. It should also be noted that the antenna in figure 4.23 forms a half-loop. In regions subject to ice storms it is possible to inject a line frequency current at the base of the driven element and add a ground wire between the bases to complete an AC heater circuit.

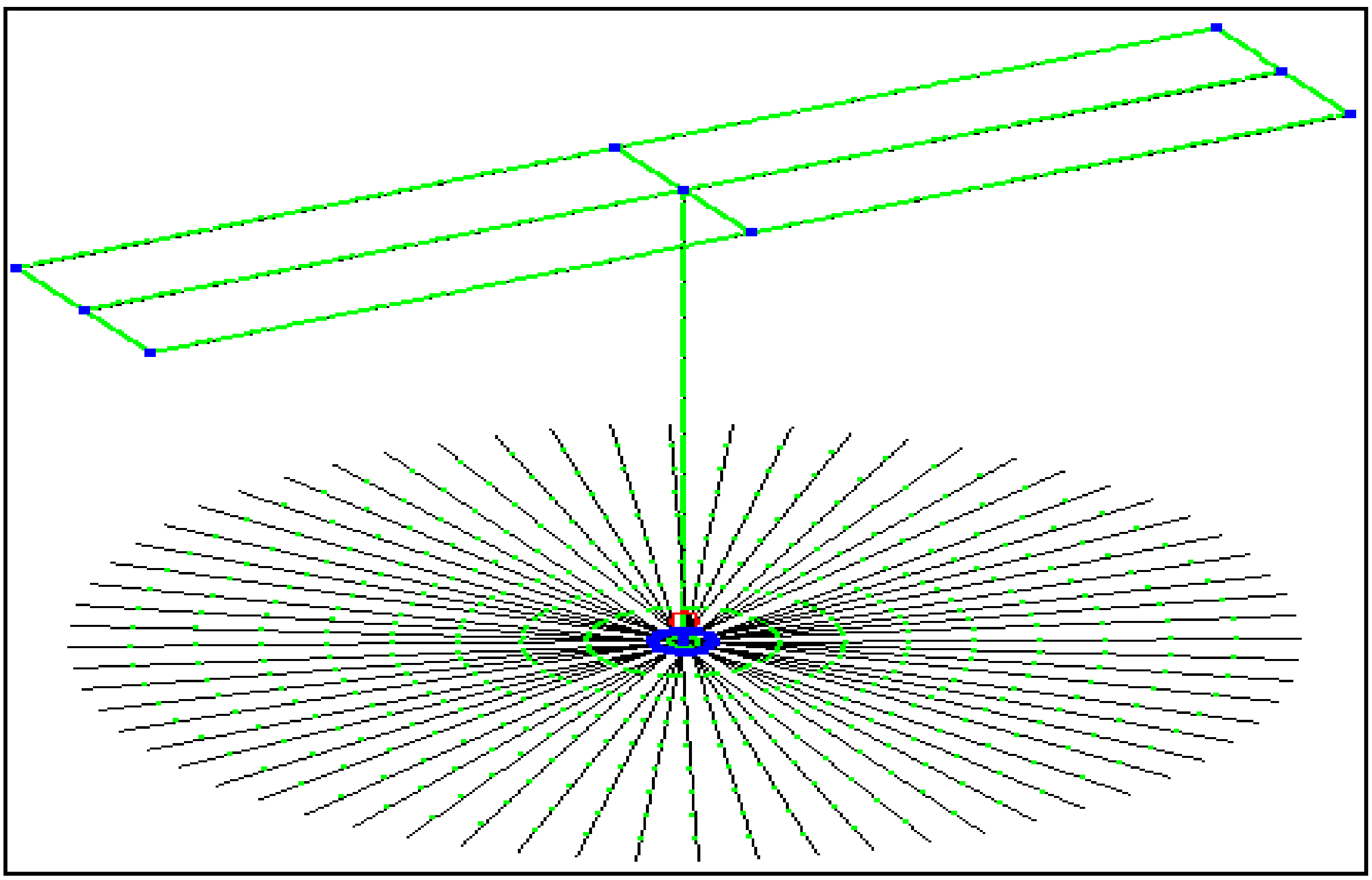

Figure 4.24 - Increased top-loading.

We could have used more top-loading (figure 4.24) to increase efficiency. This increases the efficiency to 15.1% or +1.8 dB which is almost 1 dB better than the multiple tuning example in figure 4.23. Of course we could also combine multiple tuning with the increased top-loading as shown in figure 4.25. The efficiency is now 18.6% or -7.3 dB. In this example multiple tuning increases the signal by another dB. Multiple tuning can increase efficiency but usually only modestly.

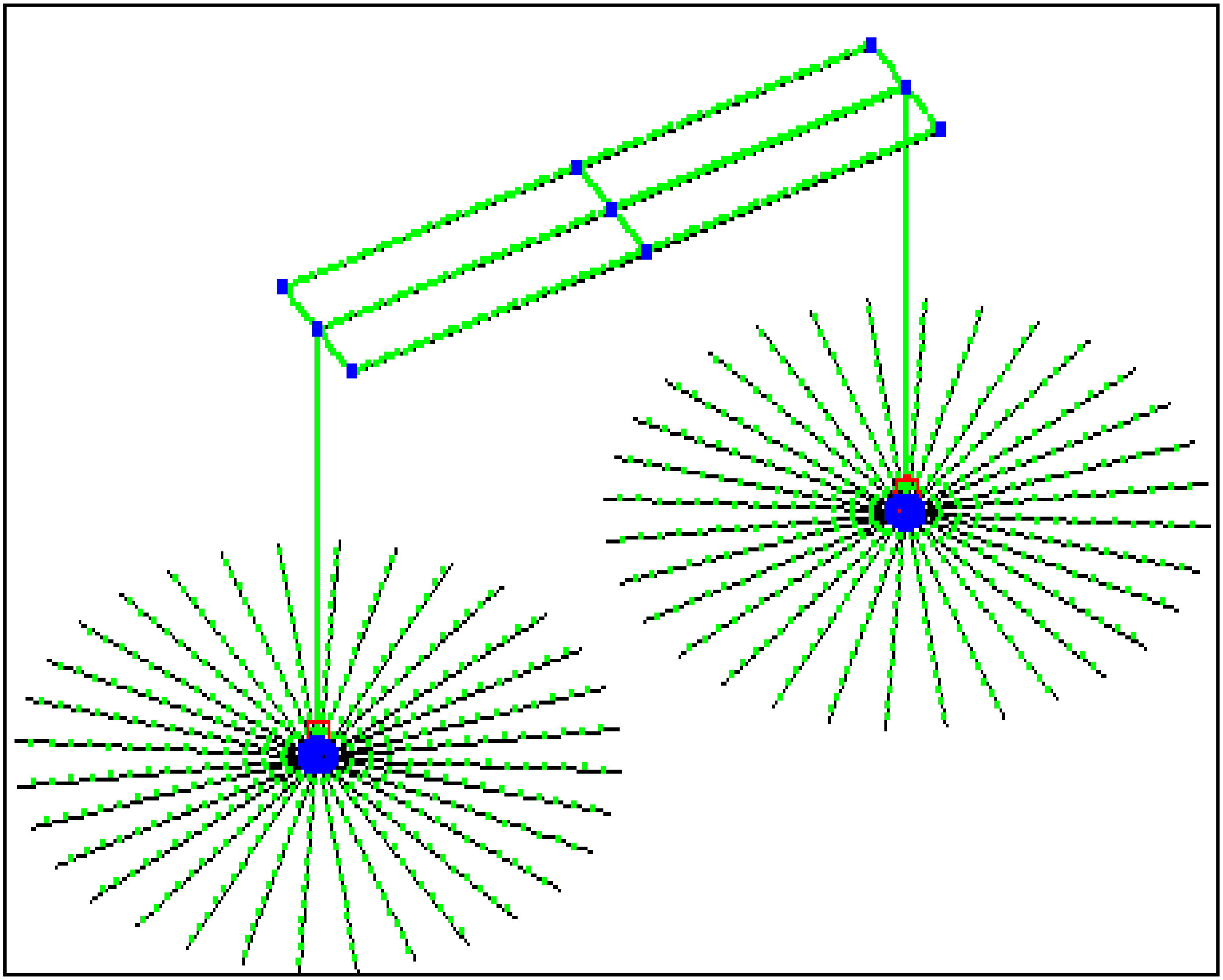

Figure 4.25 - Multiple tuning applied to figure 4.24.

4.6 Loop antennas

Ground systems are a nuisance and sometimes impractical. We might consider using a transmitting loop like that shown in figure 4.26.

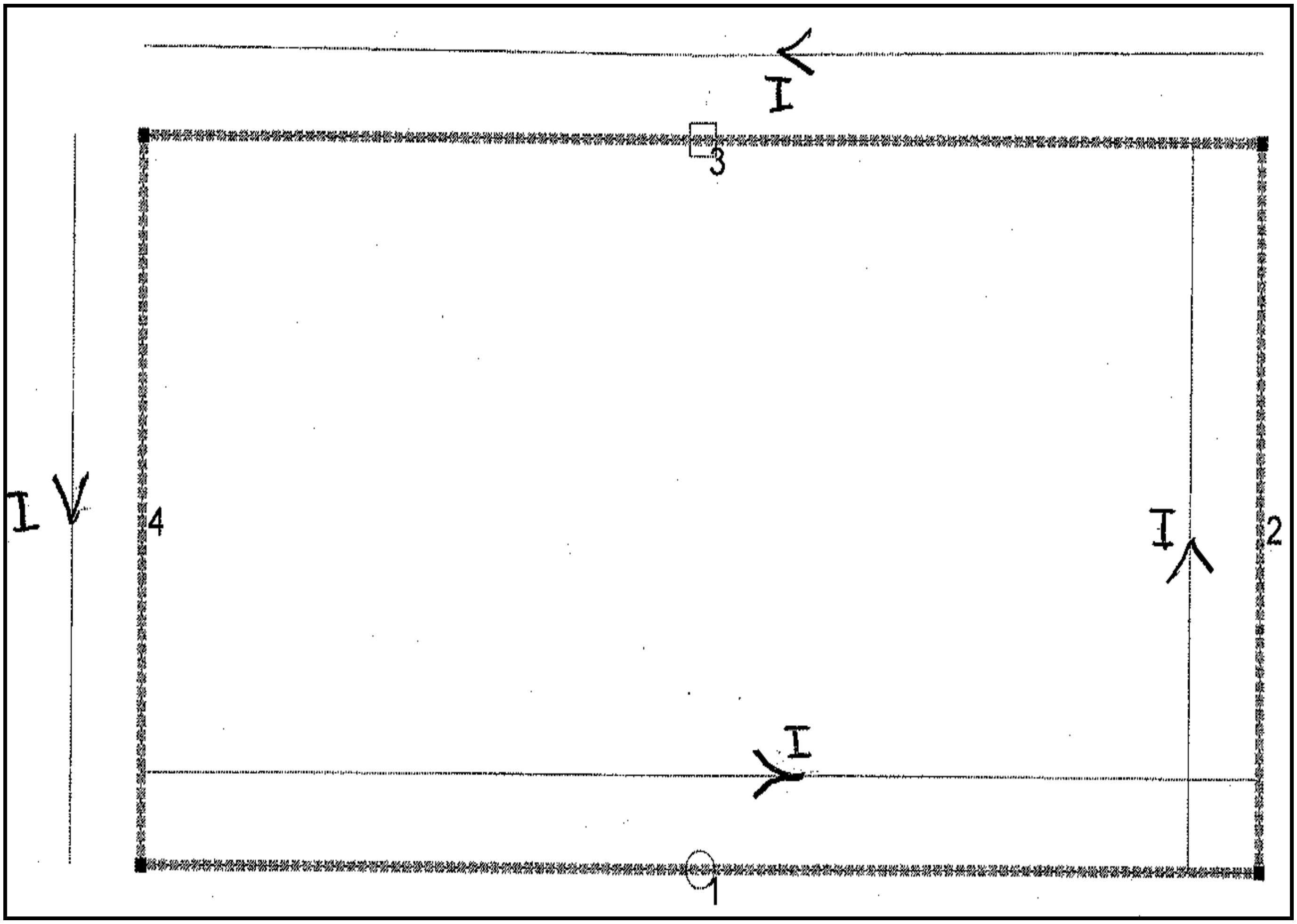

Figure 4.26 - Loop antenna.

In this example the horizontal wires are 100' long and the vertical wires 50', all #12 copper wire. The bottom wire is 8' above average ground (0.005/13) and f=475 kHz. The antenna is resonated with a capacitive load at the center of the upper wire (3) where Xc=537Ω. As is typical for small loops the current amplitude around the loop varies only +/- 5% with very little phase difference and for the values given, the radiation efficiency will be ≈1.8%. John Andrews, W1TAG, WE2XGR/3, has used a similar loop made with RG8 coax (diameter≈0.3"). This increases the efficiency to ≈3.1%. Not great but if 100W of input power is available then the maximum EIRP can be reached. Even if we used super-conducting wire the efficiency would still be limited to ≈4.2% due to near-field ground losses, a factor often overlooked in transmitting loops. Poorer soil would mean even lower efficiency. These efficiencies are not very encouraging but then transmitting loop antennas are not known for their efficiency! This antenna has a directive pattern with a mix of vertical and horizontal radiation shown in figure 4.27.

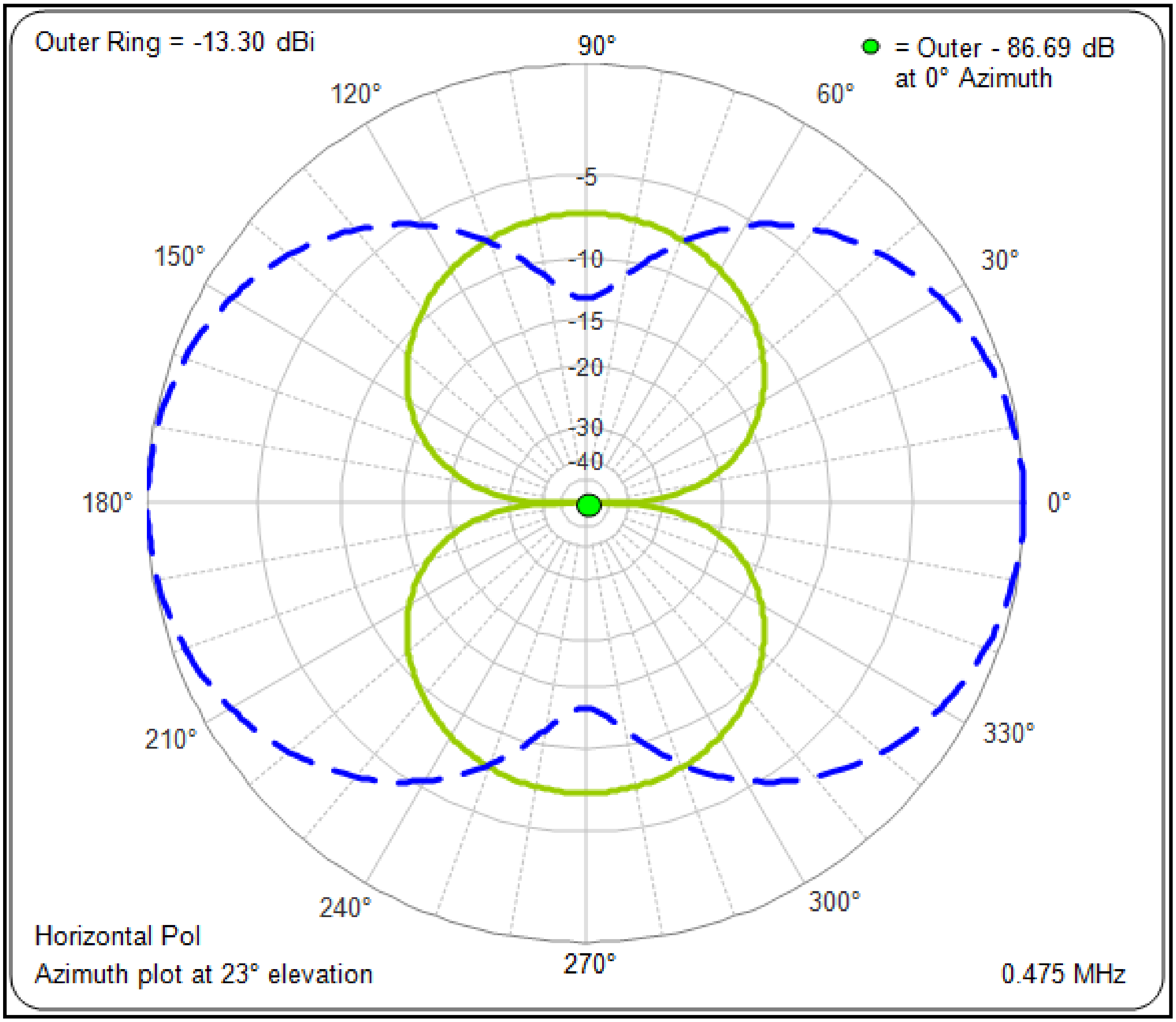

Figure 4.27 - V and H radiation pattern at 23° elevation.

This discussion raises the question "can we use inductors to improve transmitting loop efficiency?" Suppose we take the antenna in figure 4.23 and elevate it 8' above ground and replace the ground system with a single wire as shown in figure 4.28. The loads are equal, XL1=XL2= 4443Ω for resonance at 475 kHz. The antenna has the same dimensions and construction as the loop in figure 4.26. However, the current pattern is very different! The currents in horizontal wires are out of phase, with a null at the center of the horizontal wires. The currents in the downleads are now in-phase.

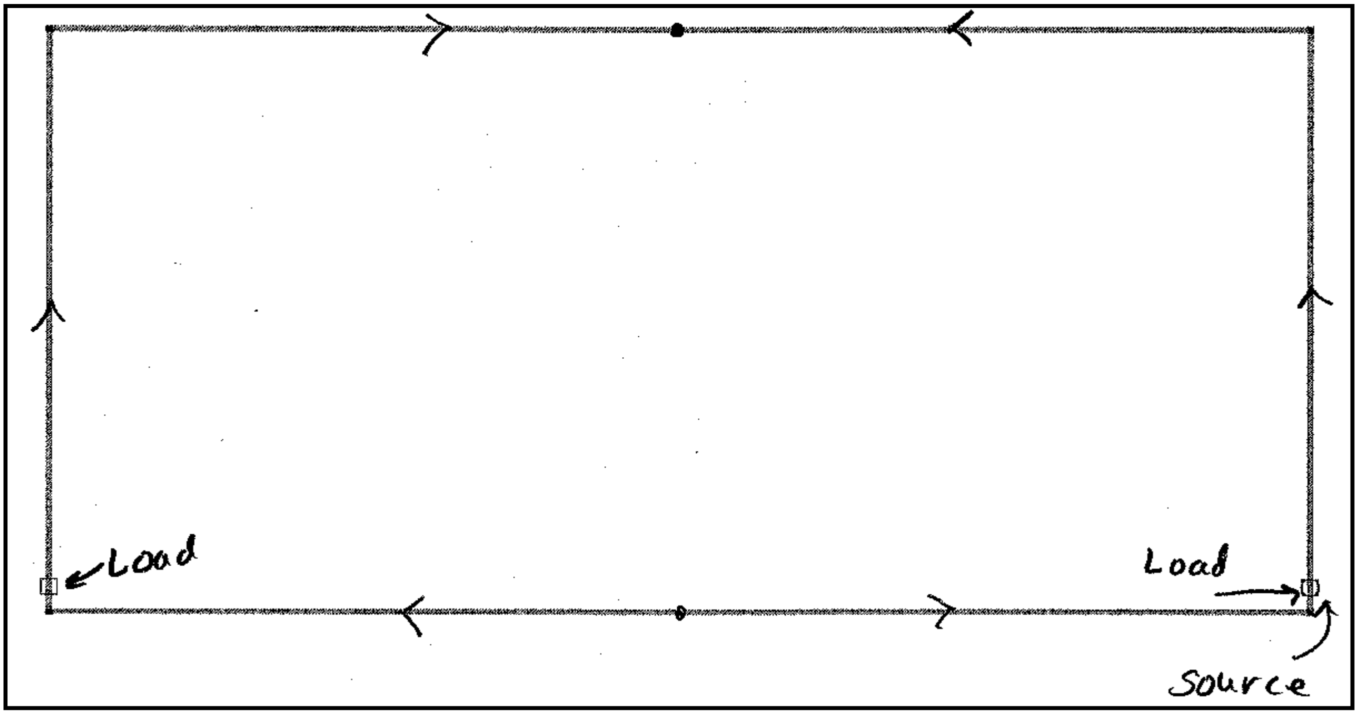

Figure 4.28 - Loop currents with multiple tuning inductors.

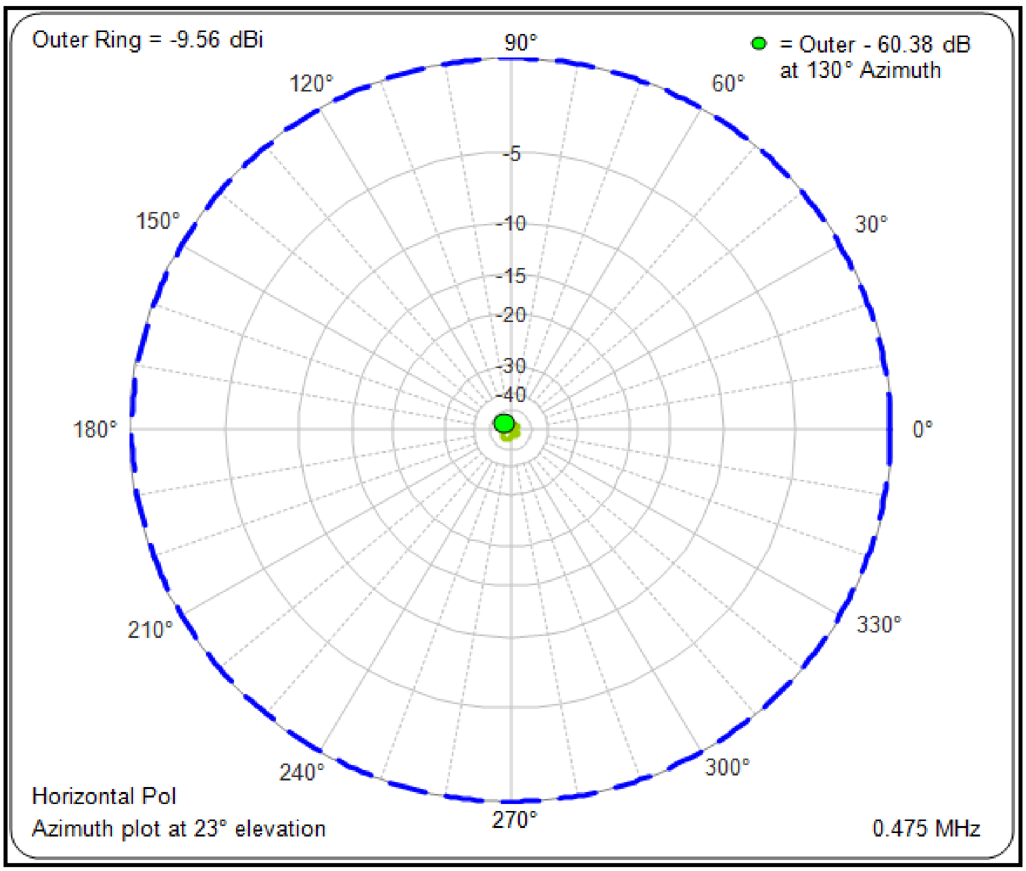

Figure 4.29 - Directional pattern with multiple tuning.

This results in a very different radiation pattern as shown in figure 4.29. The radiation is almost all vertically polarized and uniform in all directions. The efficiency including conductor and soil losses has been increased from 1.8% to 8.9%, almost 5X!

> ***It is important to recognize that by adding two inductors to a small transmitting loop (0.14λ circumference) we have transformed the current distribution making it a very different antenna!***

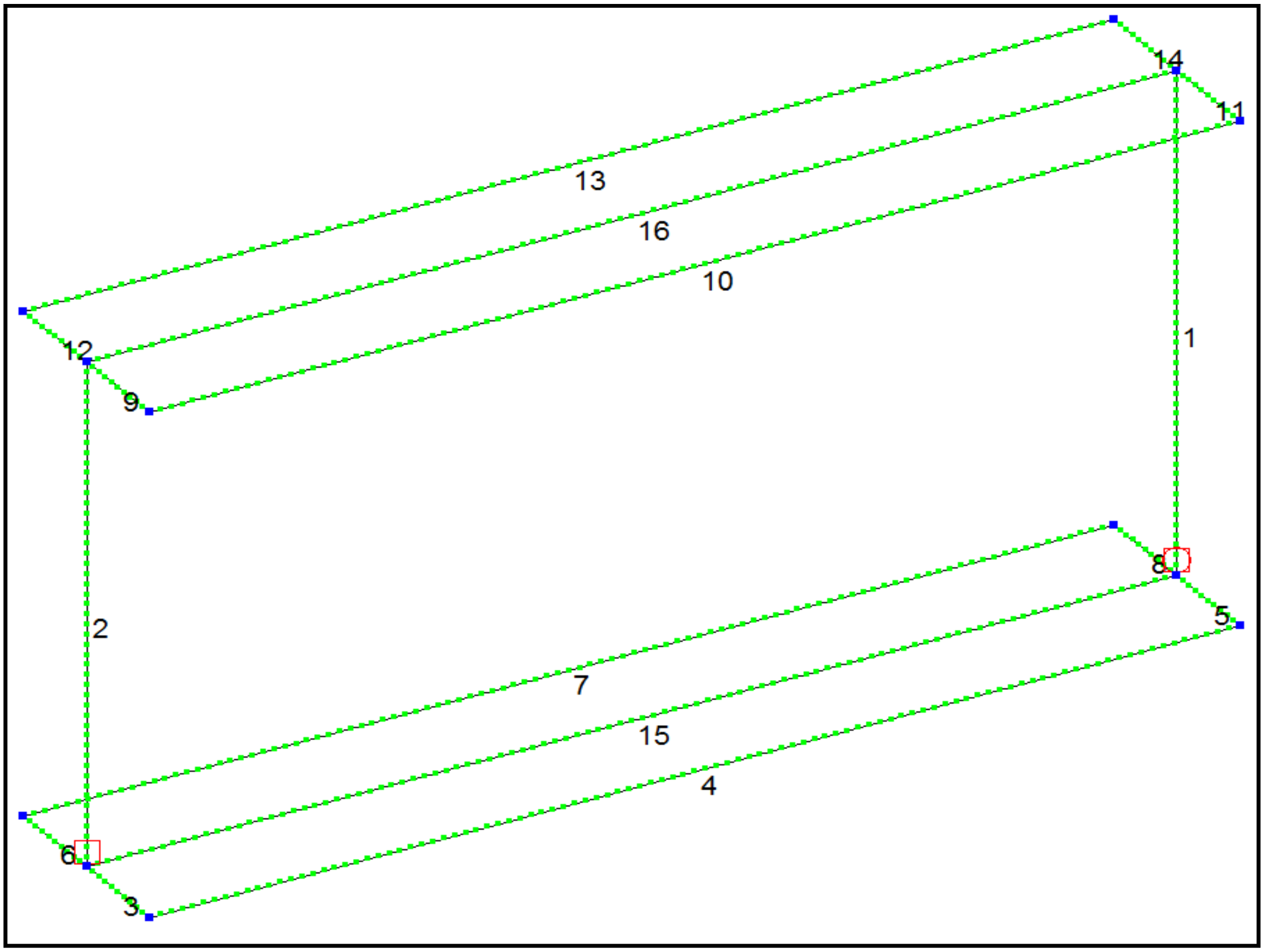

Figure 4.30 - Adding top and bottom capacitive loading.

Returning to figure 4.28, 8.9% is a significant improvement over the conventional loop but still no great shakes. In the case of the simple loop, conductor loss is important but in the multi-tuned loop the losses are dominated by loading inductor RL. We know how to reduce RL: add capacitive loading! Figure 4.30 gives an example. The efficiency of this configuration is ≈12.8%, far higher than the simple loop. It's still not as good as antennas with ground systems (figures 4.24 and 4.25) but then there is no ground system. A fair trade perhaps?

4.7 Effect of feedpoint location

Although it's very convenient to feed a vertical directly at the base we don't have to. In short LF/MF verticals we can modify the current distribution by placing the feedpoint some distance up the vertical.

In a full size λ/4 vertical we can ground the base and place the source at any point along the conductor. You could for example insert an insulator at some elevated point with the coaxial feedline inside the vertical up to the insulator with the shield connected to the lower section of the antenna at that point and the center conductor connected to the upper section. The base of the vertical is grounded. When you do this the antenna is still resonant but, depending on the placement of the insulator, the feedpoint impedance can now be 50Ω or whatever you wish. This is a common trick at HF to improve the SWR without a matching network. However:

> When you do this in a λ/4 vertical there is no detectable change in the current distribution along the vertical, i.e. A' is not changed.

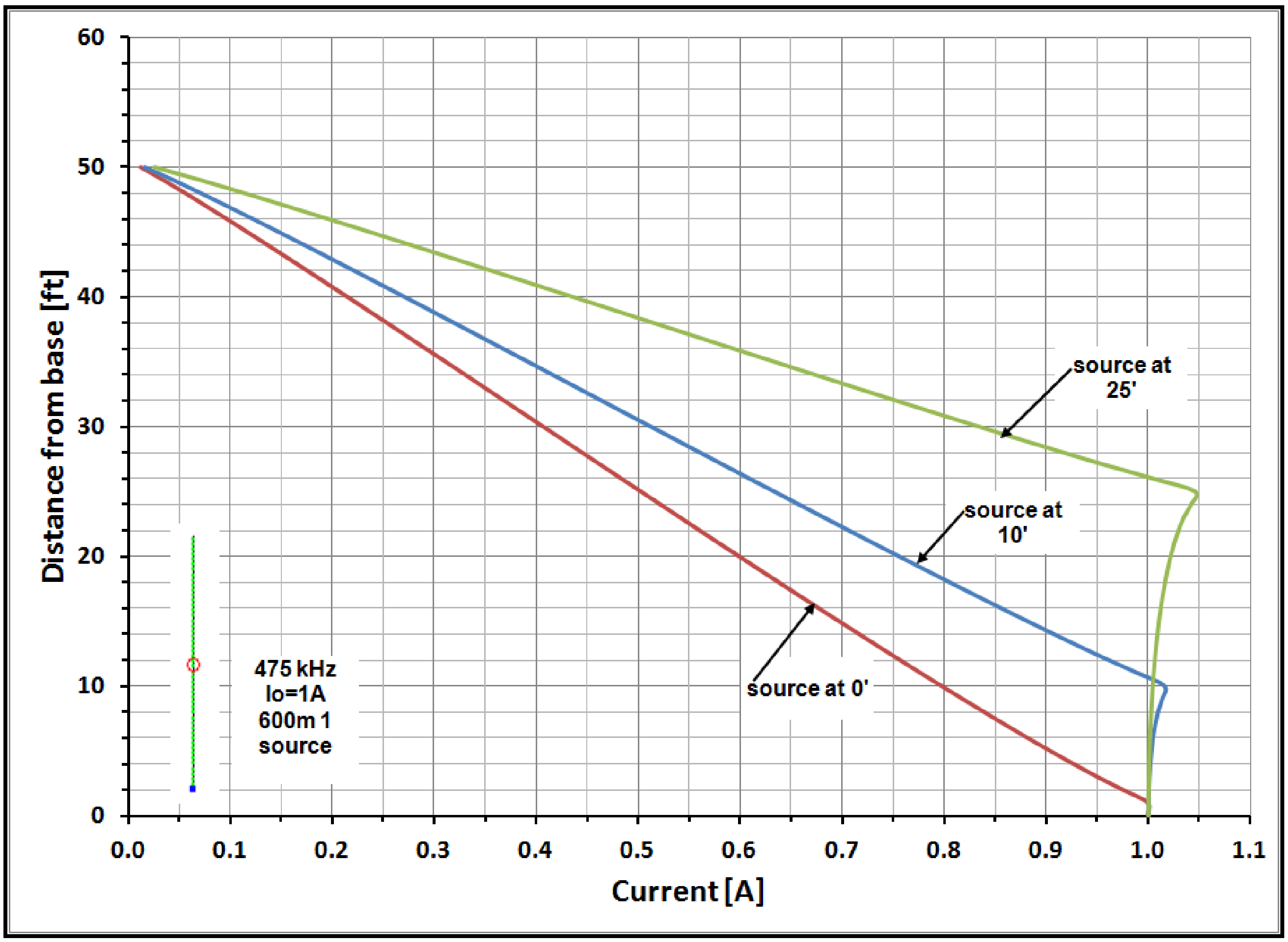

Figure 4.31 - Effect on current distribution of different feedpoint locations.

When you play this same game with a short vertical, say H=50' at 475 kHz, the current distribution and A' changes greatly! As shown in figure 4.31, the current below the feedpoint is nearly constant. Elevating the feedpoint increases A' which implies an increase in Rr. While the increase in Rr from moving the tuning inductor up into the antenna has long been known I've never see the effect of moving the source discussed.

This effect is interesting but it does not appear to be particularly useful for LF/MF transmitting antennas because there will be a large tuning reactance which can be moved to increase Rr. If we try to leave the tuning inductor at the base and only elevate the feedpoint then we have the problem of decoupling the feedline at the base. If the base inductor is wound with coaxial line then the base inductor can provide the decoupling but that complicates the tuning inductor. However, this scheme does allow us to obtain higher Rr without having to elevate the loading inductor.

While not particularly useful for transmitting verticals an elevated feedpoint can be very useful when short verticals (e-probes) are used as elements in a receiving array. These verticals are normally not resonated with loading inductors but simply connected to the inputs of high impedance amplifiers. Moving the feedpoint up into the e-probe increases A' which increases the effective height (h) which increases the terminal voltage for a given E-field intensity. It should also be kept in mind that adding top-loading to an e-probe will also increase the effective height.

Summary

This chapter has shown several variations for the placement and number of loading inductors. Once the antenna height and top-loading have been maximized the following points were made:

...loading inductor arrangements can significantly improve the efficiency...

...and make it possible to excite a grounded tower to act as radiating part of the antenna...

...for a receiving e-probe vertical, moving the feedpoint up into the vertical can increase the received signal....

References

[1] ARRL Antenna Book, any edition, chapter on mobile antennas.

[2] Laport, Edmund, Radio Antenna Engineering, McGraw-Hill, 1952. You can find this one free on-line by Googling Edmund Laport.

[3] Schelkunoff and Friis, Antennas, Theory and Practice, page 426

[4] Terman, Frederick E., Radio Engineers Handbook, McGraw-Hill Book Company, 1943

[5] Harrison and King, Folded Dipoles and Loops, IRE AP transactions, March 1961, pp. 171-187

[6] Raines, Jeremy, Folded Unipole Antennas, McGraw-Hill, 2007

[7] Guertler, Rudolf, Impedance Transformation in Folded Dipoles, IRE Proceedings, September 1950, pp. 1042-1047

[8] Leonhard, Mattuck and Pote', Folded Unipole Antennas, IRE AP Transactions, July 1955, pp. 111-116

Chapter 5

LFMF Ground Systems

5.0 Why do we need a ground system?

The discussion of efficiency in earlier chapters has focused on loss introduced by the tuning inductor which is reasonable given that RL often represents a major loss in short antennas. While much can done to reduce Xi and increase Rr, there are practical limits and at some point we have to start thinking about reducing other losses. A substantial portion of the power supplied to the antenna may be absorbed in the soil near the base. To reduce this loss a ground system is used.

This chapter begins with some basic definitions and then moves on to examples and practical questions like:

- what type of ground system?
- how many radials?
- radial lengths?
- what performance can we expect?
- optimum use of a fixed amount of wire?

A range of examples have been chosen to provide general guidance but none should be taken as exact numerical descriptions for all cases. You will have to do some measurements, modeling and/or calculations to arrive at the best solution for your unique situation.

5.1 Choices for ground systems

Ground systems can take several forms:

1. A radial wire fan lying on the ground surface or buried a few inches.
2. A rectangular grid of wires
3. Single or multiple ground rods.
4. Elevated wires in the form of a counterpoise or "capacitive" ground.
5. Combinations of the above. Limited only by imagination!

The choice of ground system will be dictated by the operating wavelength, available space, soil mechanical characteristics (i.e. sandy loam or tree stumps and boulders), available resources, etc. Because of the much longer wavelengths at LFMF and significant differences in soil electrical characteristics between LFMF and HF (shown in chapter 1), the ground systems may be significantly different from what we are accustomed to at HF.

Most of information given here is derived from NEC modeling simply because it's far easier to generate that way but in the end we have to have experimental verification. References [1] through [15] provide this and in general the correlation is excellent. As a practical matter the accuracy of NEC modeling is limited by how well the actual soil electrical characteristics are known. References [9, 15] show how to measure these characteristics but we must keep in mind that soil characteristics vary widely with moisture content with changes of season. We have to assume the worst case when modeling. At HF seasonal variation is usually not noticeable especially if there is an extensive ground system. But with the short, heavily top-loaded verticals and limited ground systems typical on 2200m and 630m, the seasonal change in soil conductivity can significantly detune the antenna. This is mostly a capacitive loading effect as the soil conductivity increases or decreases.

Since the mid 1930's the generally accepted "ideal" for a ground system has been the broadcast (BC) system using 120 0.4λ radials, typically #8 Copperweld buried 12-18". This system originated with the work of Brown, Lewis and Epstein[16] and other work by Brown[17-21] published in the 1930's. The value of this work was immediately recognized and a standard design was adopted by the BC industry and sanctified with FCC regulations. While this was certainly seminal work of great value, it's adoption as a standard had the effect of making ground system design appear to be a "perfected art" and attention shifted to other problems! At 630m and especially at 2200m the "standard" ground system is not even remotely practical for amateurs! We have to be more inventive and accept compromise.

For this discussion the ground system radii have been limited to ≤150' but even that is a bit large, so many examples have smaller ground systems.

5.3 Feedpoint equivalent circuit

Figure 5.1 is an equivalent circuit representing the resistive part of an antenna's feedpoint impedance (Ri). Rr is the radiation resistance representing the radiated power (Pr) and Io is the current at the feedpoint. Rg accounts for the power dissipated in the soil (Pg) and RL represents the tuning inductor loss. RL is a physical resistor arising from the series resistance (RL=XL/QL) of the inductor, but Rr and Rg are <u>not</u> lumped physical resistors. Rc represents conductor loss and Rmisc other losses.

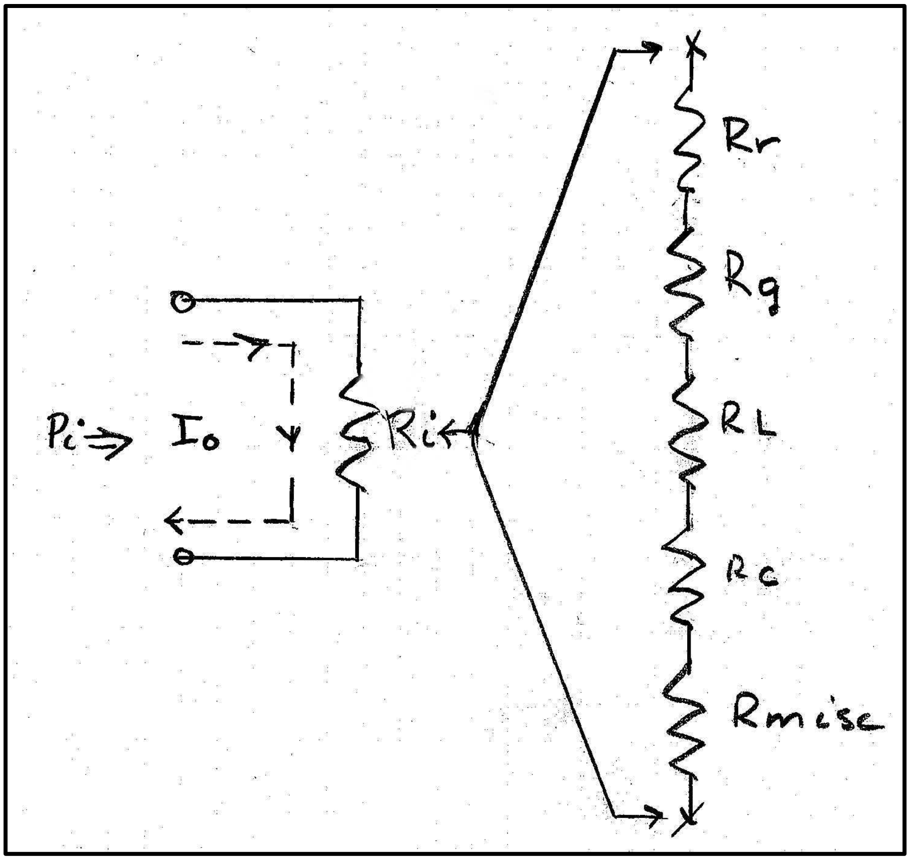

Figure 5.1 - Equivalent circuit for the resistive part of the feedpoint impedance.

In earlier chapters we've seen how dependent Rr is on specific details of the antenna: i.e. dimensions and loading. Earlier examples determined Rr with perfect ground but Rr can also be a function of soil electrical characteristics and ground system design. This effect is prominent at HF though significantly less at LFMF[22].

Rg depends on frequency, soil electrical characteristics, details of the ground system and <u>the antenna associated with the ground system</u>. If we modify the antenna, even without changing the ground system or soil, Rg will change. The reason for the change in Rg is that the soil loss depends on the EM field intensities close to the antenna. The field intensity changes when the antenna is altered which in turn changes soil loss and Rg. For example, the field intensities are directly proportional to Io, when we add top-loading Rr increases which means we must reduce Io to stay within the Pr limits. Reduced Io results in less power dissipation in the soil and an altered value for Rg.

5.4 Definitions for Pr, Pg, Rr and Rg

Figure 5.2 illustrates "Pr" and "Pg". The dashed line represents a virtual hemispheric surface, with radius r, enclosing the antenna. Pr is defined as the total power radiated through the surface of the hemisphere. Pg is defined as the power passing through the bottom into the soil, which is the ground surface, and dissipated in the soil. r=λ/2 is usually chosen because it is approximately the outer boundary of the reactive near-field for verticals with a height of λ/8-λ/4. For shorter antennas r can be somewhat smaller[22].

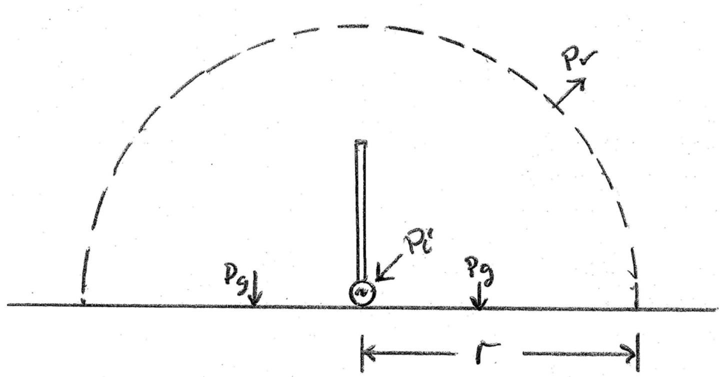

Figure 5.2 - Pr and Pg.

Rr and Rg are defined in terms of Pr and Pg:

$$\mathbf{Rr} \equiv \frac{\mathbf{P_r}}{\mathbf{I_o^2}}\ \mathbf{\Omega} \quad (5.1) \qquad \mathbf{Rg} \equiv \frac{\mathbf{P_g}}{\mathbf{I_o^2}}\ \mathbf{\Omega} \quad (5.2)$$

5.5 Efficiency with ground losses

In the following discussion Ri at the feedpoint is assumed to be the sum of Rr + Rg + RL. Conductor and other losses will be omitted, not because these are unimportant, but the interest here is in RL/Rr and Rg/Rr. We can state efficiency η In terms of Rr, RL and Rg:

$$\eta = \frac{1}{1+\frac{R_L}{R_r}+\frac{R_g}{R_r}} \qquad (5.3)$$

5.6 Simple advice

The instructions for an adequate ground system can be boiled down to:

1) Use at least 50 radials. Note, we are not talking about 0.25λ radials! Most backyards will only have room for 30-40' radials. Where possible the radials should be somewhat longer than the height of the vertical

2) In the case of a very large top-hat, the radials should extend out to 1.25X the top-hat radius if possible.

3) When a large number of radials are used the wire size is not important. The wire needs to be strong enough to be installed and survive in its environment.

4) Almost any metal can be used for the radials but the usual choice is insulated copper house wiring because it is usually cheaper than the same wire bare. For an elevated system #17 aluminum electric fence wire can be used. However, lying on the surface or buried, aluminum wire may degrade quickly

5) If the radials are lying on the surface, use lots of staples to keep them close to the ground so mowing or other traffic will not damage them.

6) Use at least one ground stake at the base for safety.

5.7 Ground system for an unloaded vertical

Figure 5.3 shows a typical buried radial wire ground system.

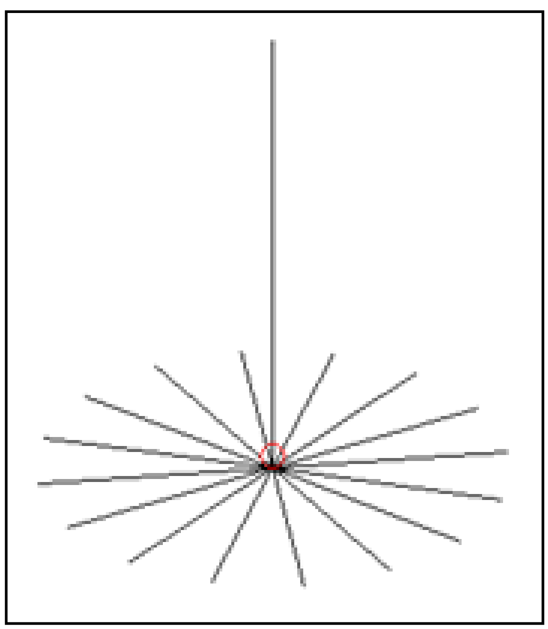

Figure 5.3 -Vertical with a buried wire radial ground system

Typical urban lots have a width of 50-60' and a depth of roughly 100-120'. Usually most of the lot will be occupied by the house and front yard setbacks so in the end at most a 50'X50' area is available for the ground system. For those lucky enough to have more property longer radials can be used so for the following discussion radial lengths from 25' to 150' will be used. Unless noted otherwise average soil (0.005/13) is assumed along with bare #12 radial wire buried 12". Insulation on the wires and larger or smaller wire sizes generally have only small effects. Radial wires can also be lying on the ground surface with similar effectiveness.

Figure 5.4 is an example of how efficiency varies as a function of radial length for a given number of radials (32 in this case). For H=20' the maximum usable radial length is about 30'. Making the radials longer has very little effect. The reason is that in a vertical this short Xi is very large which means a large loading inductor (XL=Xi) and consequently large RL. For this short a vertical RL is large compared to Rr and it's also becomes large compared to Rg as radial length approaches 30', i.e. Rg falls as the radial lengths are increased so that it becomes small compared to RL.

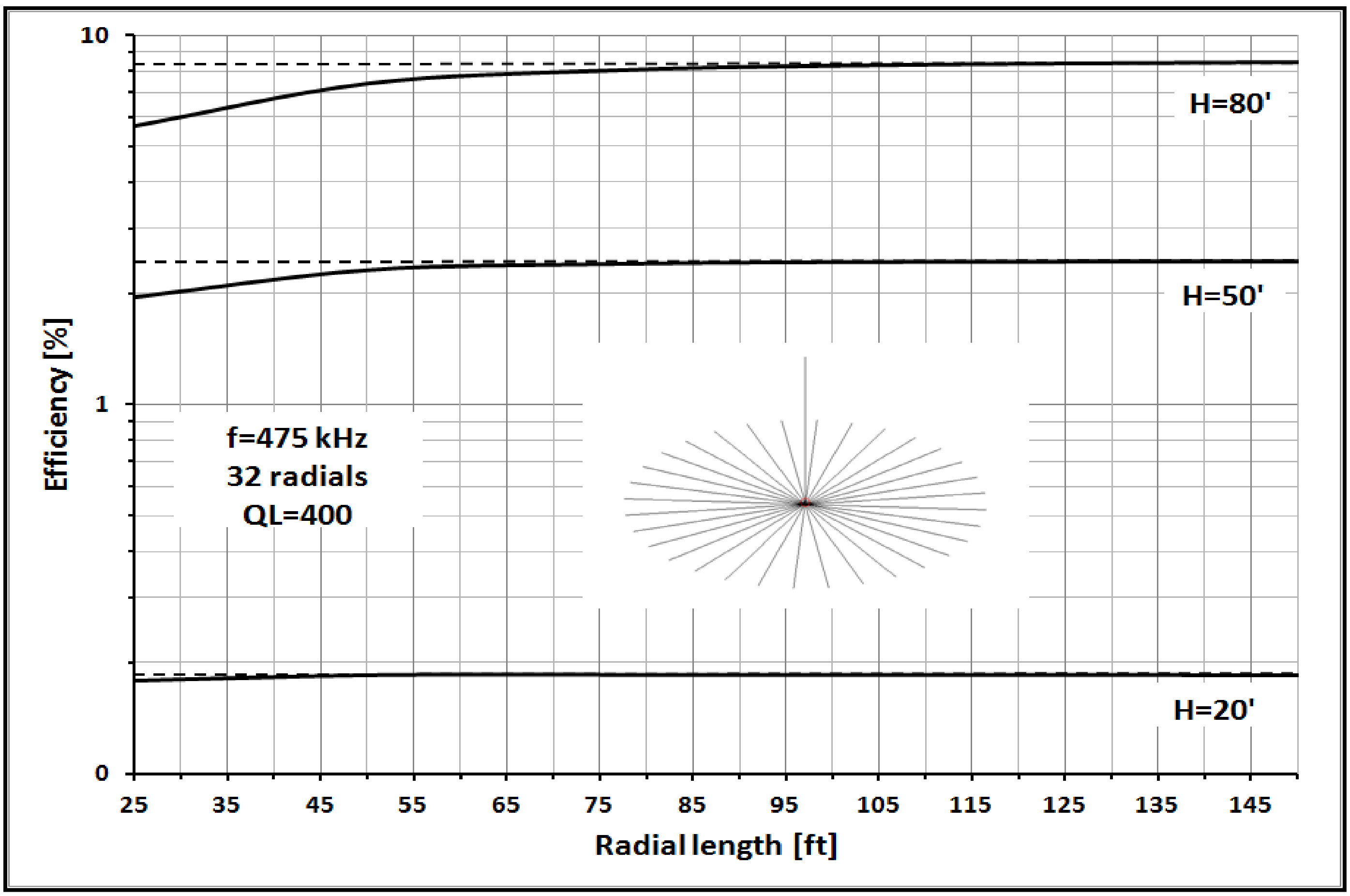

Figure 5.4 - Efficiency as a function of H and radial length.

As we increase H the efficiency rises quickly because RL is reduced and the maximum usable radial length expands to 50' or more. There is a further increase in efficiency and usable radial length when we let H=80' but even with that height the efficiency is not very good. As shown in chapter 3 even a small amount of capacitive top-loading can greatly reduce Xi and improve efficiency. In addition, because the loading inductance will be much larger on 2200m, the dominance of the RL will be even more pronounced. For those reasons the use of at least some top-loading will be assumed in the following discussion.

5.8 Ground systems for urban lots

In this and following sections T antennas with H=20', 50' or 80' and a top-wire length =100' will be used to illustrate the interplay between efficiency, radial length, radial number, QL and H. It should be kept in mind that although the following examples use a single top-wire for loading, any loading structure which gives similar value for Xt will produce the same result. As emphasized in chapter 3, it's the amount of loading (Xt) not the shape that matters. This means the conclusions we'll draw from the following graphs can apply to different antennas with similar heights.

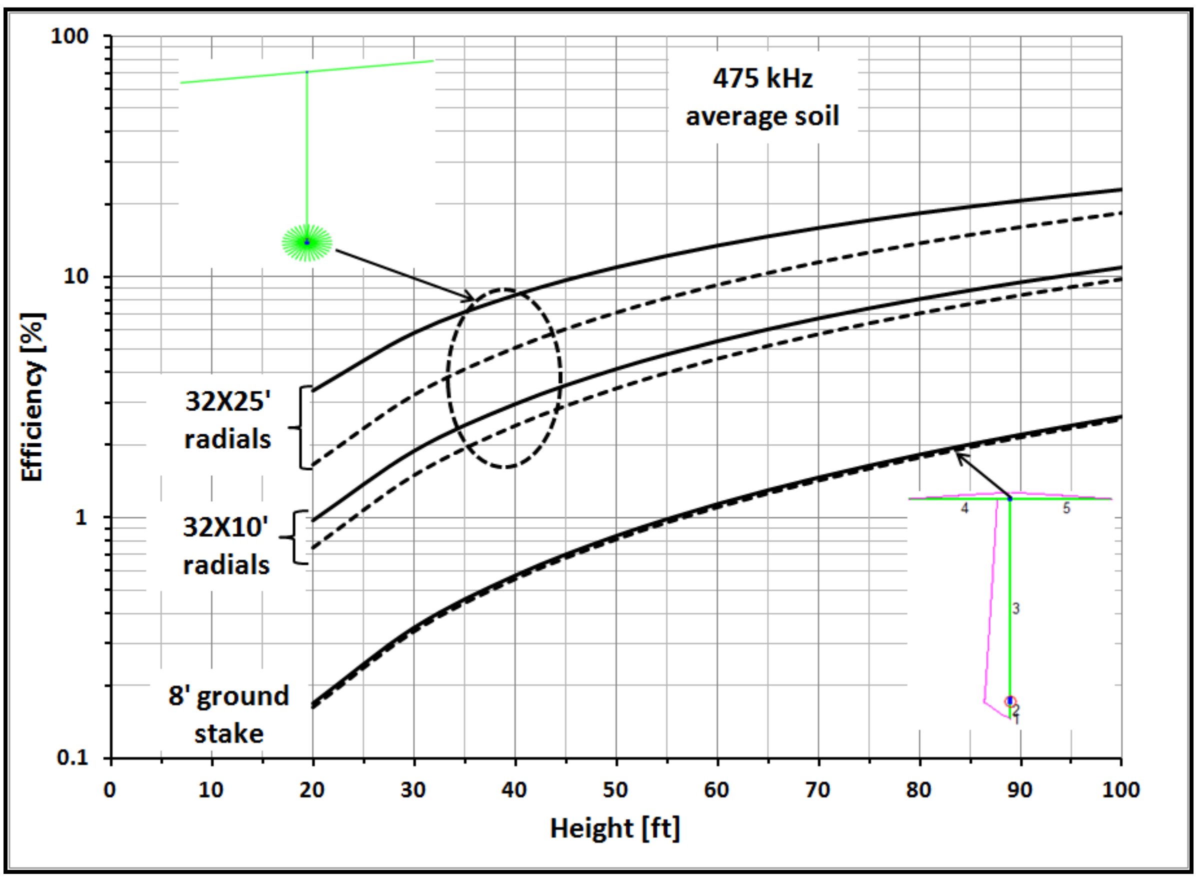

Figure 5.5 - Small ground system comparison.

The simplest ground system would be a single ground stake. From there we can add various numbers of radials of lengths up to 25'. It is also possible to have a radial system with ground stakes at the far ends of the radials.

Figure 5.5 gives comparisons between three ground systems: a single 8'X5/8" ground rod or stake, thirty two 10' radials and thirty two 25' radials. The solid lines represent the efficiency without the tuning inductor loss and the dashed lines represent the efficiency when QL=400. With a the single ground rod the ground loss (Rg) is so large that RL doesn't matter very much. As soon as we add even the small radial system (32X10') the efficiency increases by almost an order of magnitude and expanding the radial lengths to 25' yields another factor of ≈3X in efficiency. This is due to reductions in Rg with longer radials.

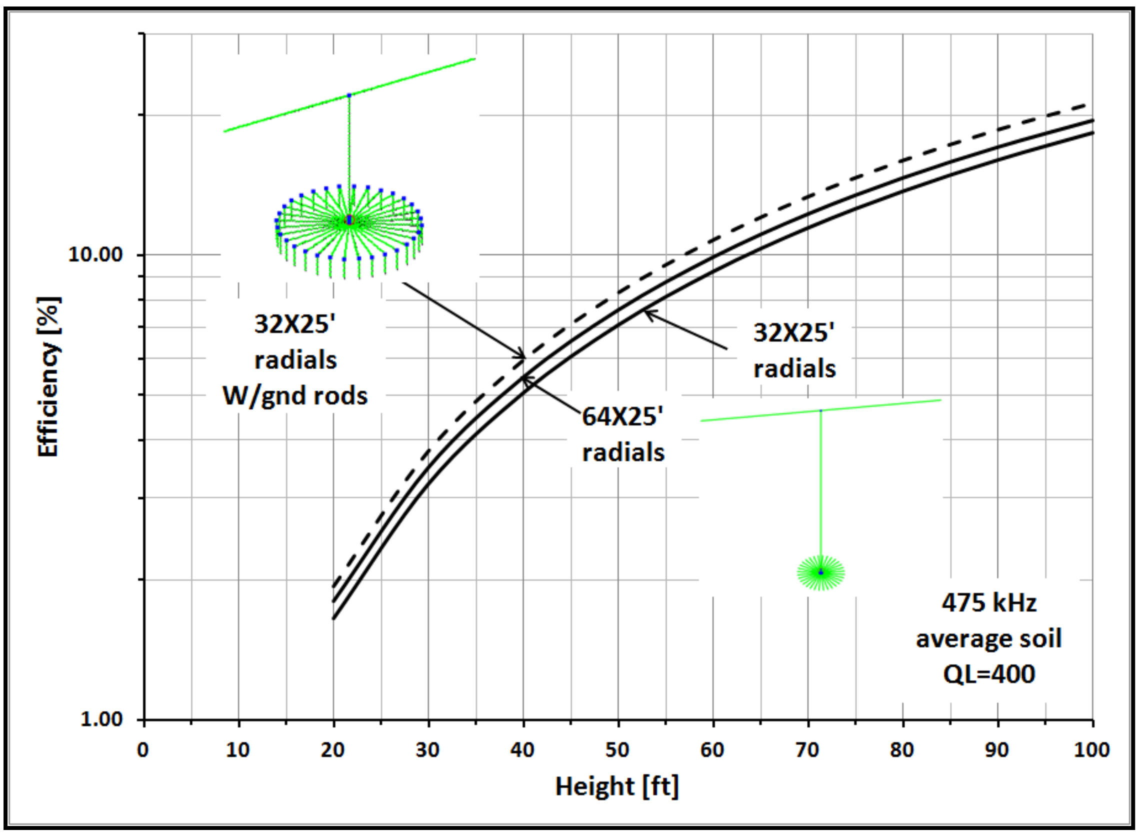

Figure 5.6 - Increasing radial number and adding ground stakes.

As shown in figure 5.6, increasing the radial number to 64 improves the efficiency but not by a lot. The efficiency tapers off because the Rg/Rr term becomes small compared to the RL/Rr term. Another option when radial lengths are restricted by available space is to add ground stakes at the ends as indicated in figure 5.24. This yields more improvement in efficiency than doubling the radial number but represents significant cost and labor for which only 1% or so is gained!

5.9 Larger ground systems

Figure 5.7 shows efficiency as a function of radial length with the radial number as a parameter. The dashed lines are straight-line approximations for the curves which allow us to identify the point of vanishing returns, i.e. the point at which increasing the radial length begins to provide less improvement. On some of the graphs this point is labeled the "knee". Many of the graphs to follow will have these dashed lines. Two points to notice, first increasing the radial number is very helpful, in fact we could have gone to 128 radials and still have had some useful improvement. Of course every time you double the number of radials you double the amount of wire used! Second, the efficiency improves rapidly up to lengths of 60-70' before leveling out. By 150' there isn't much point in making the radials longer. Independent of radial number (for this example!) the "knee" corresponds to a radial length of ≈65'-70' which is a bit more than H (50'). This the reason for comment 1 in section 5.6. Note that 64 radials are only marginally better than 32.

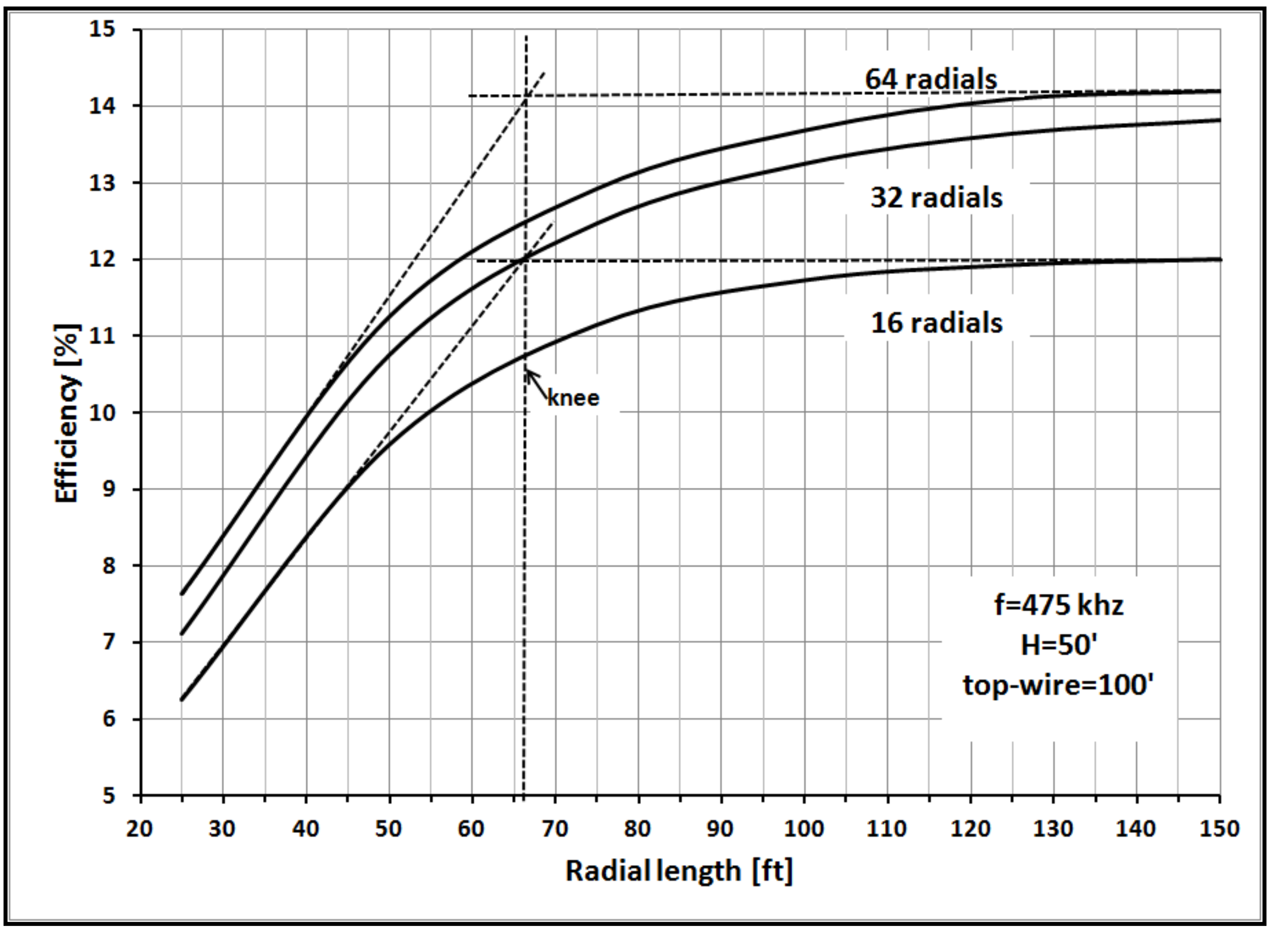

Figure 5.7 - Efficiency versus radial length.

Figure 5.8 graphs the same data in a different way, efficiency as a function of radial number for different radial lengths. Independent of radial length, the knee is about 30 radials. Above the knee it's better to use longer radials rather than more numerous radials! This behavior is quite different from what is normally seen in HF verticals. At HF the radial lengths are typically 0.125λ or longer but at 475 kHz 50' is only 0.024λ and the E and H fields are quite different as shown in appendix TBD. Ground system optimization is somewhat different at LFMF.

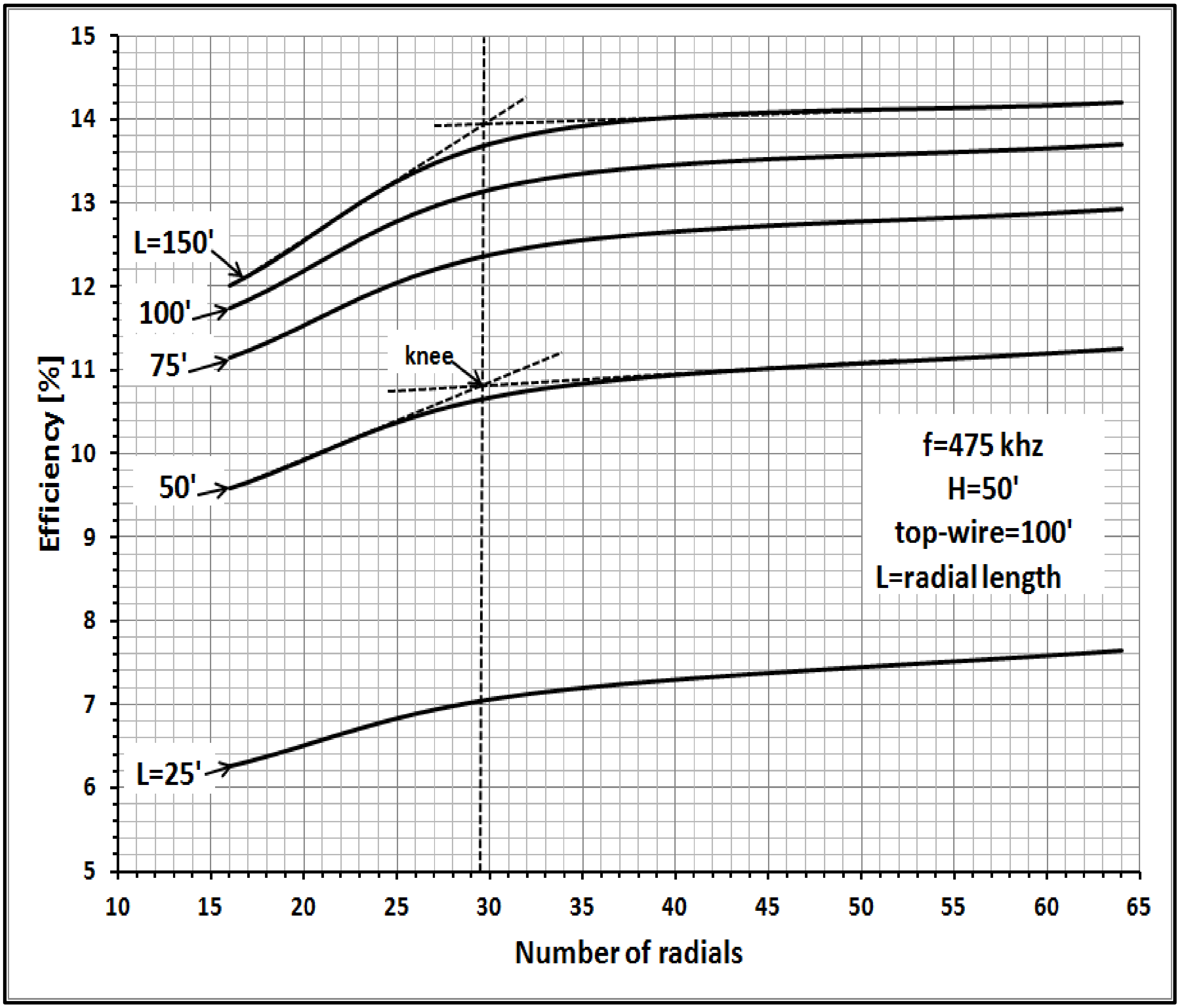

Figure 5.8 -efficiency versus radial number for different radial lengths.

One frequently asked question is, "If I have a limited amount of wire available for radials how should I divide it up?" In other words, "should I use a few long radials or many short radials?" We can re-plot the data in figure 5.8 to answer that question as shown in figure 5.9. Here we see that for H=20', 32X50' radials would be the best use of the wire. At H=50' or 80' either 16X100' or 32X50' radials would work pretty much the same. Given that 50' radials take up 1/4 the area of the 100' radials, 32X50' radials would seem to be a good choice in those cases also. It is interesting to note 32 radials is close to the knee value for the radials in figure 5.8, although this should not come as a great surprise since all three of these graphs are using the same data graphed in different ways.

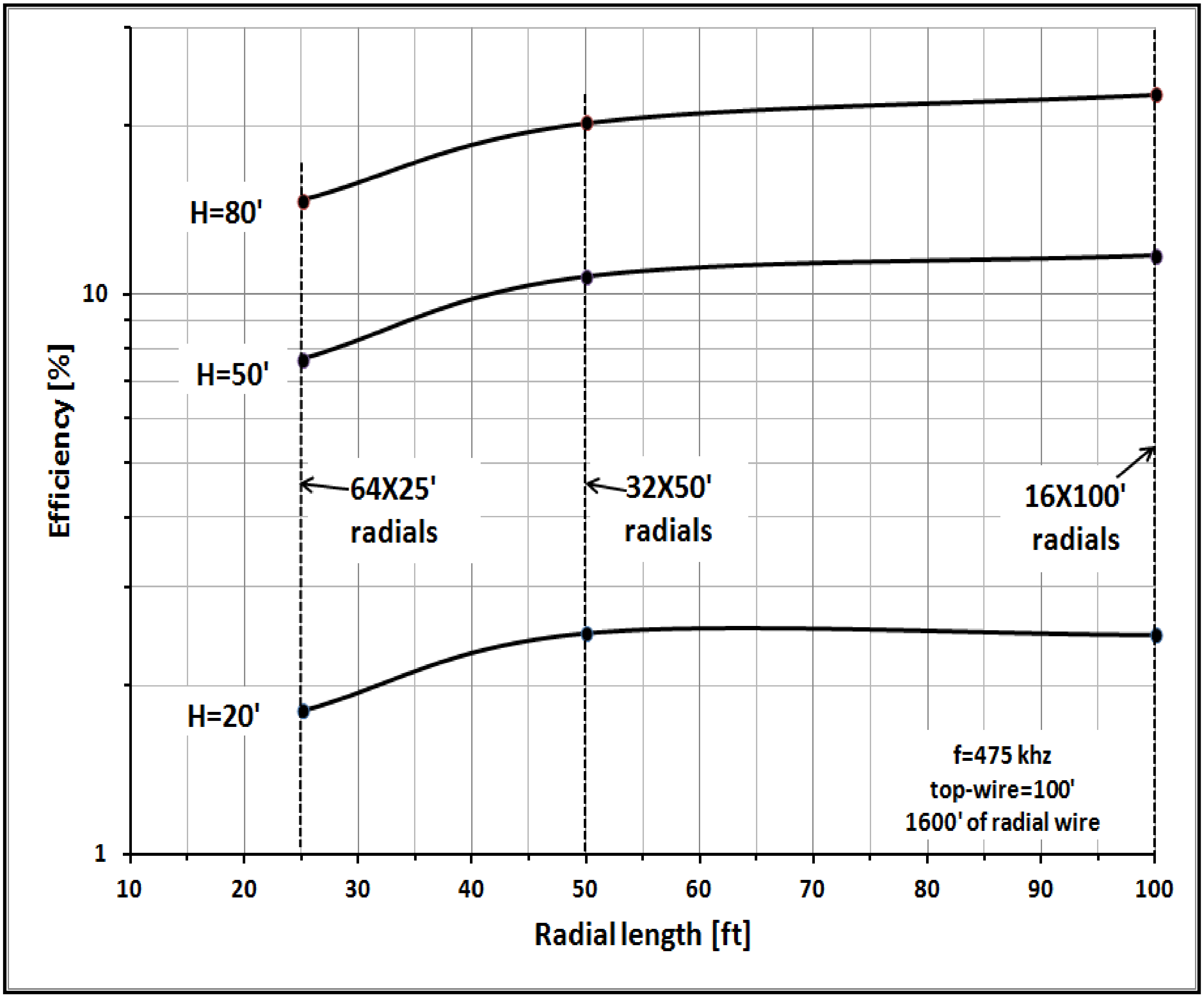

Figure 5.9 - Efficiency versus radial length for various heights.

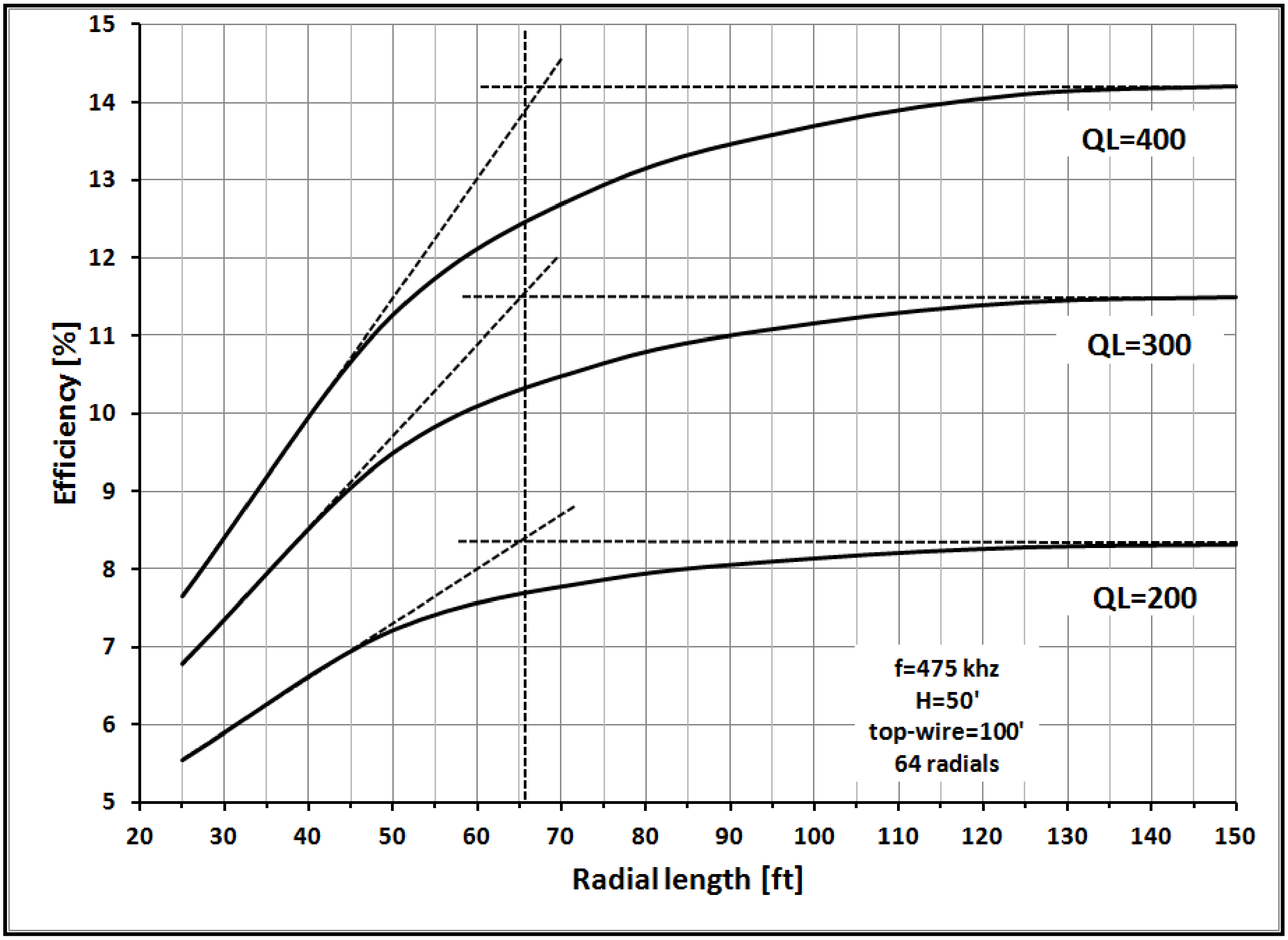

Figure 5.10 -Efficiency versus radial length versus QL.

Up to this point QL has been assumed to be 400. At 475 kHz that's practical with some modest effort. However, as shown in chapter 6, much higher QL is possible but requires careful inductor design and construction. Figure 5.10 illustrates what happens to the efficiency when QL is lower, as it easily could be. This graph shows how critical QL is for efficiency. For radial lengths < 70' both QL and radial length play a role but as we make the radials longer the accompanying reduction in Rg/Rr gets smaller and RL dominates. This again reminds us that if we really want higher efficiency we need to increase height and/or Xt. With more top-loading, RL is smaller and longer radials can be used to advantage.

Figures 5.11 through 5.13 are another way to show the relationship between H, radial length and radial number.

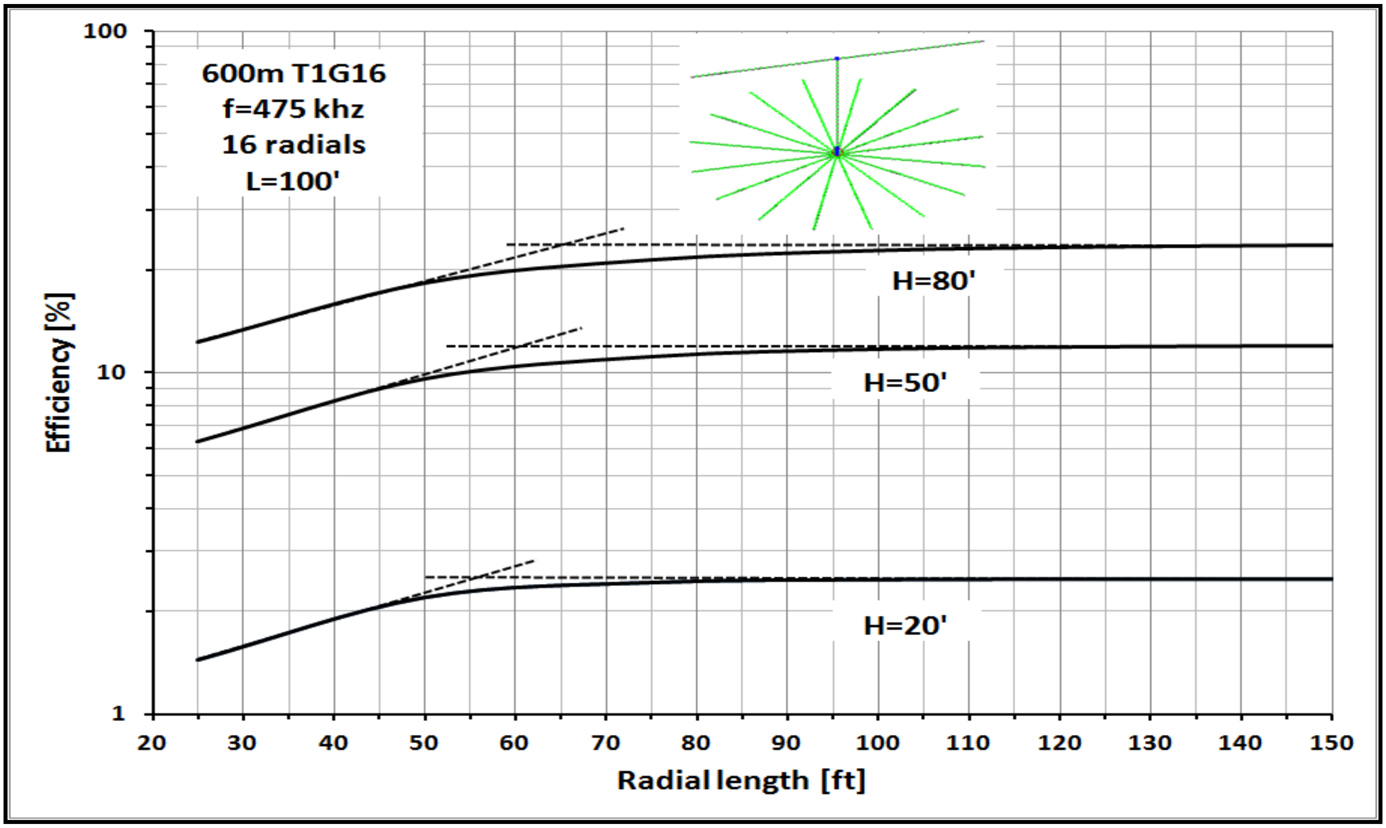

Figure 5.11-Efficiency versus radial length, 16 radials.

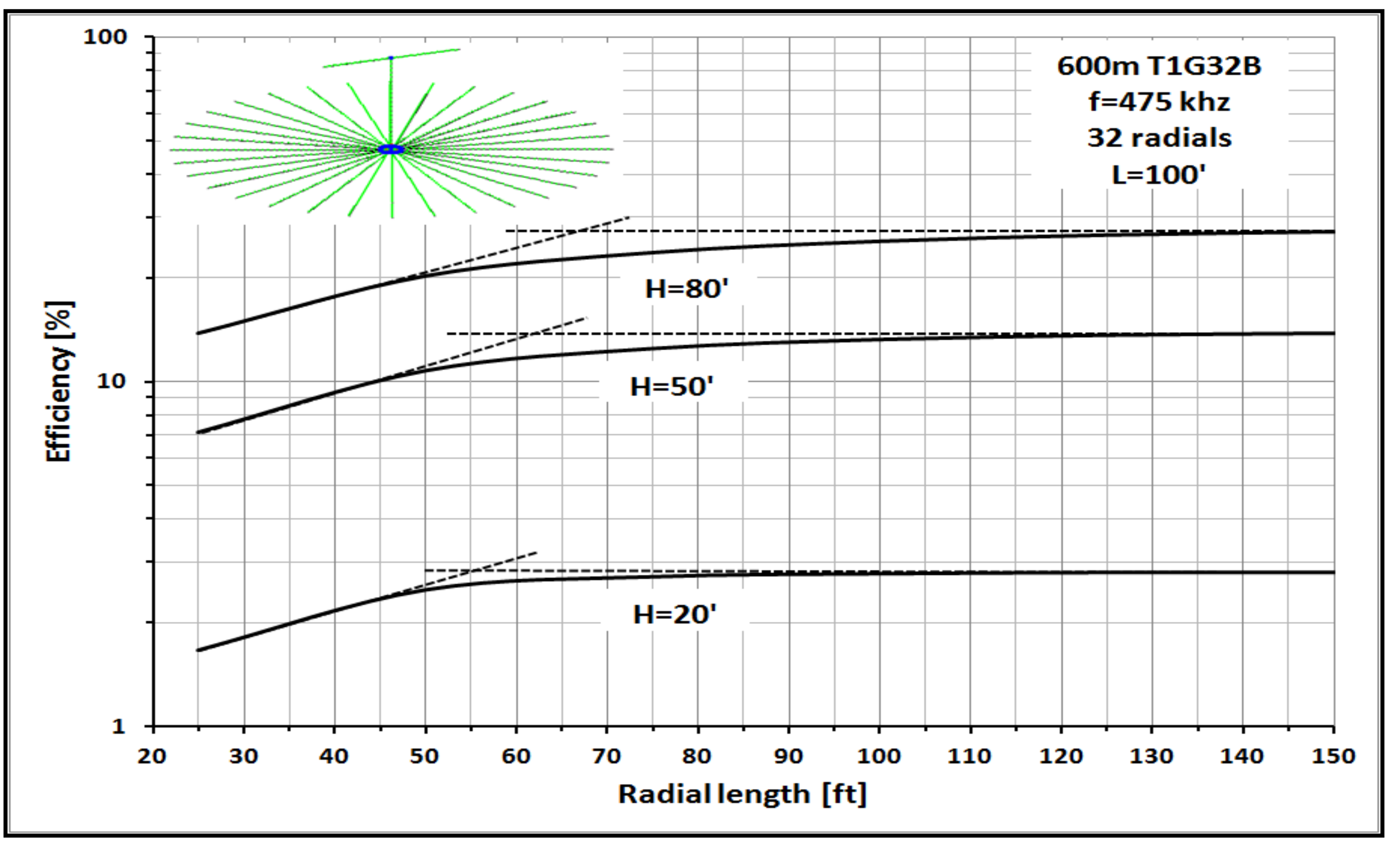

Figure 5.12 - Efficiency versus radial length, 32 radials.

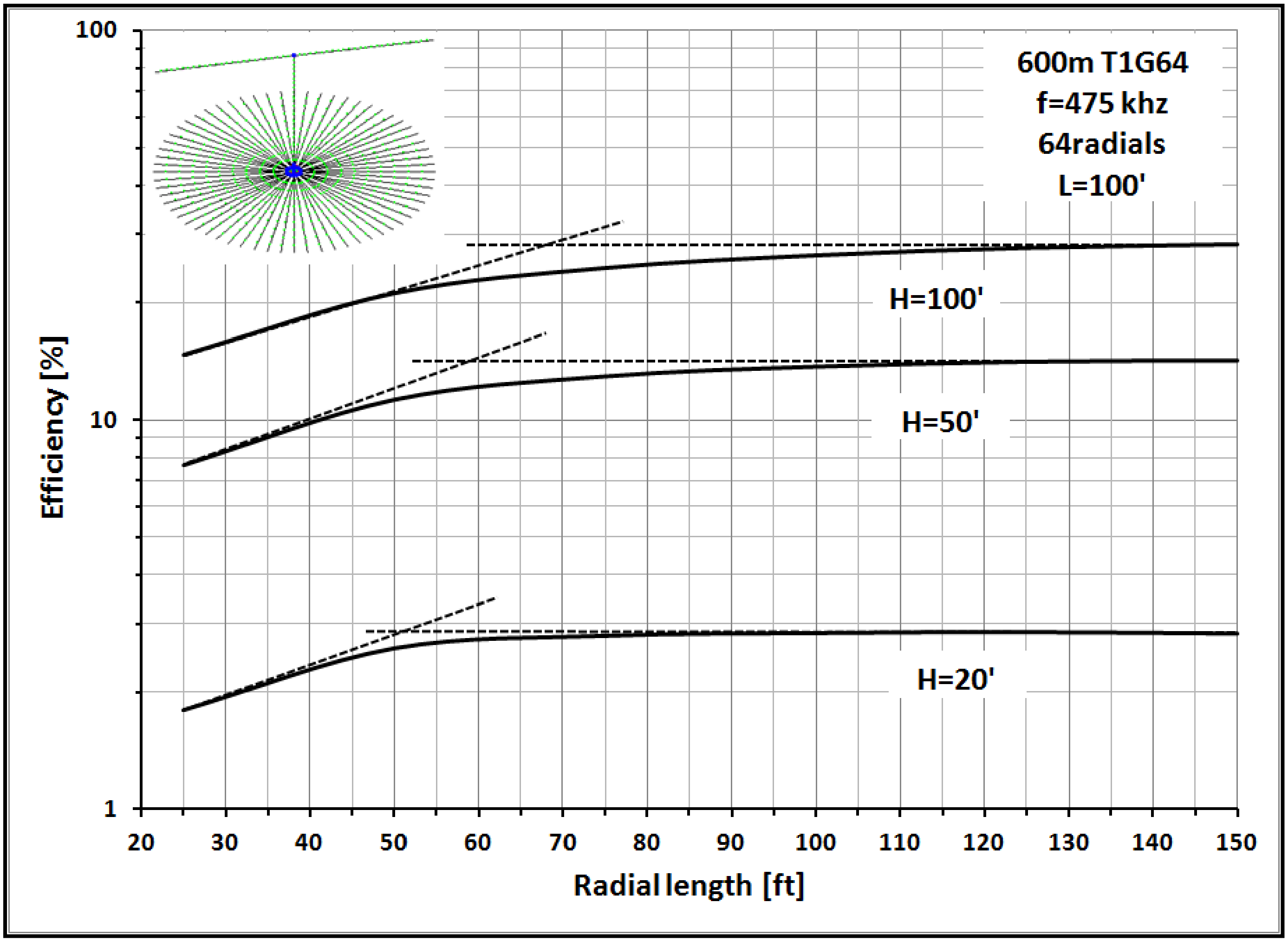

Figure 5.13 - Efficiency versus radial length, 64 radials.

The earlier suggestion that radial length is related to height has long been part of amateur antenna lore. The idea is that with a 1/4-wave antenna you use 1/-4wave radials and with an 1/8-wave vertical 1/8-wave radials, etc. To explore this idea I modeled 1/4-wave and 1/8-wave verticals at 1.8 MHz over average soil (0.005 S/m, Er=13). The radial lengths were stepped in the sequence 1/8, 1/4, 3/8, 1/2-wavelength. The number of radials was stepped in the sequence 4, 8, 16, 32, 64 and 128. At each point I recorded the average gain (Ga) and used this as a measure of relative efficiency between different radial configurations. The radials were buried 3" below the ground surface.

The results are shown in figures 5.14 and 5.15. Note that the vertical axis is "improvement" in dB when going from four 1/8-wave radials to more and/or longer radials. The gain for four 1/8-wave radials was used as the reference and set to 0 dB. This nicely illustrates what you might "gain" by adding more wire, in different ways, to the radial fan.

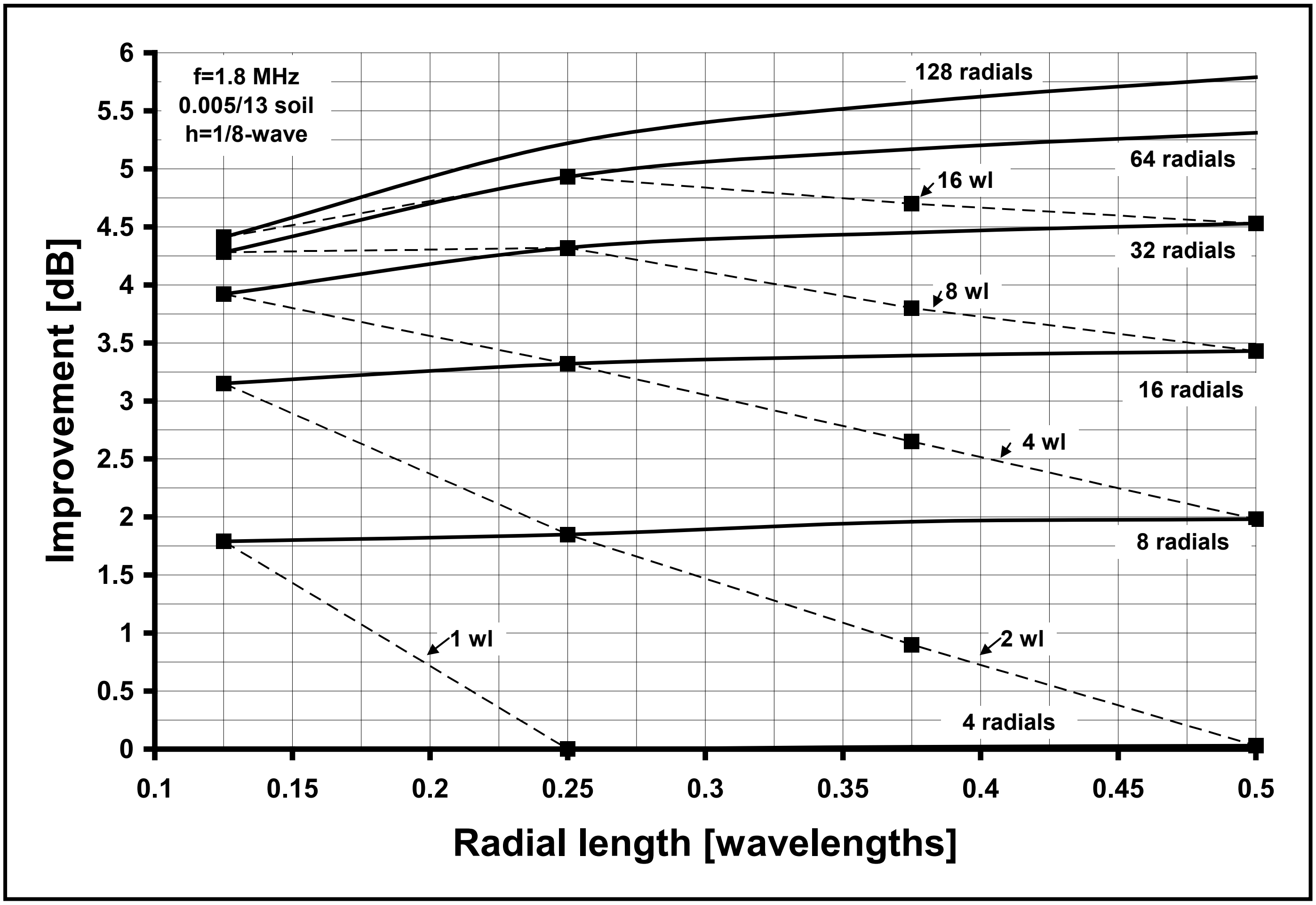

Figure 5.14, Signal improvement for various radial configurations: 1/8-wave vertical.

Note that the solid lines represent constant radial numbers of different lengths. The dashed lines connect points of common total radial wire length: 1, 2, 4, 8 and 16 wavelengths. For example, 4 radials 1/2-wave long represent a wire total of 2-wavelengths. Eight 1/4-wave and sixteen 1/8-wave radials also total 2-wavelengths, etc. 16 wavelengths at 1.8 MHz is almost 9,000' of wire, which is a substantial ground system (64 1/4-wave radials).

These graphs show that <u>how</u> you add wire to the radial system matters as well as <u>how much</u> wire you add. As we can see from the graphs when only a few radials are used (4 to 8 radials), making them longer is waste. In fact[2] you can actually lose in the case of ground surface radials. This is by no means a first look into optimum radial lengths, Stanley[23], Sommer[24] and Christman[25] have all written on this subject.

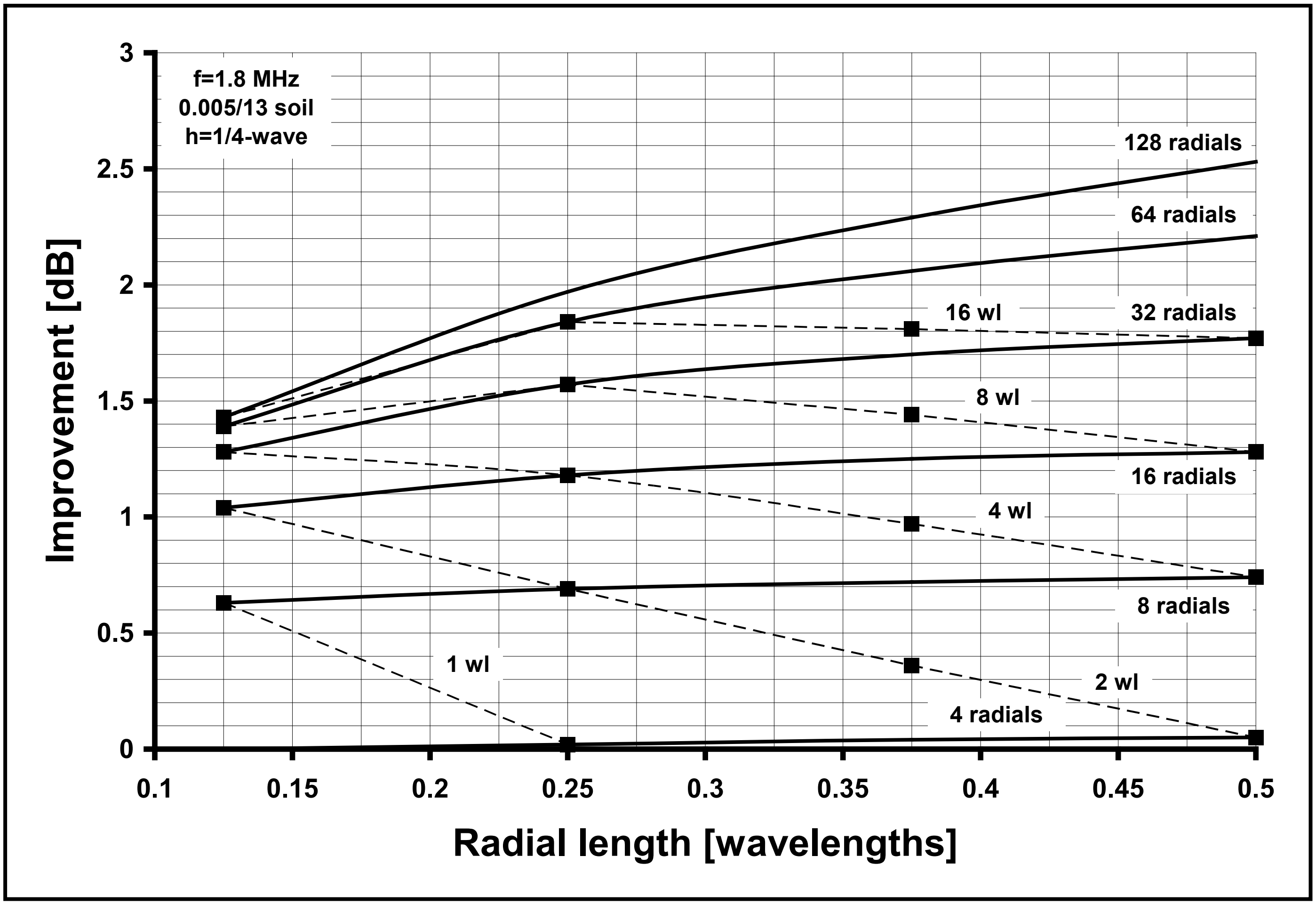

Figure 15, signal improvement for various radial combinations: 1/4-wave vertical.

Referring to figure 5.14 (the 1/8-wave vertical), if your total wire length is limited to four wavelengths, the gain improvement goes from 2 dB with 8 radials to 3.3 dB with 16 radials and 3.9 dB with 32 radials. Obviously you're much better off to using thirty two 1/8-wave radials as apposed to a smaller number of longer radials. When you increase the wire length to 8 wavelengths then it's a wash whether you use either thirty two 1/4-wave or sixty four 1/8-wave radials. The choice becomes one of convenience in laying out the radial field. If you don't have room for the 1/4-wave radials then the larger number of 1/8-wave radials will work just as well. When you go up to 16 wavelengths of wire then sixty four 1/4-wave radials will give you about 0.6 dB improvement over 128 1/8-wave radials.

When we look at the gain improvement in figure 5.15 (the 1/4-wave vertical), we see similar behavior except that when we are using 8 wavelengths of wire there is a clear advantage to go from 1/8-wave to 1/4-wave radial lengths. 1/4-wave radials also work best when 16 wavelengths of wire are available. If we go up to 32 wavelengths of wire (about 18,000'!) then radial lengths of 3/8-wavelength are best.

These graphs shed some light on a long standing rule of thumb: "the radials should be the same length as the height of the vertical". In the case of the 1/8-wave vertical this seems to be true up to at least 8 wavelengths of total radial wire. Beyond this, longer radials become a more effective use of the wire. In the case of the 1/4-wave vertical, for small amounts of wire, 1/8-wave radials are best but as we make more wire available the 1/4-wave radials are superior. The break point in radial number where you shift from 1/8-wave to 1/4-wave lengths is lower for the 1/4-wave vertical than the 1/8-wave vertical.

The physics of this seem fairly clear. Once you have greatly reduced the losses near the base of the antenna, adding more close in copper doesn't buy much. At some point it's time to put the copper further out and reduce more distant losses, which may be smaller but still significant. The difference in break point (in terms of radial number) between the two antennas stems from differences in the field intensities around the two antennas. For the same power, the fields near the base of the 1/8-wave vertical will be much higher than those for the 1/4-wave vertical (see appendix TBD) so we need to put more effort into reducing the close-in power losses.

The forgoing optimization was for relatively tall antennas. Figure 5.16 shows an a short top-loaded vertical for 630m.

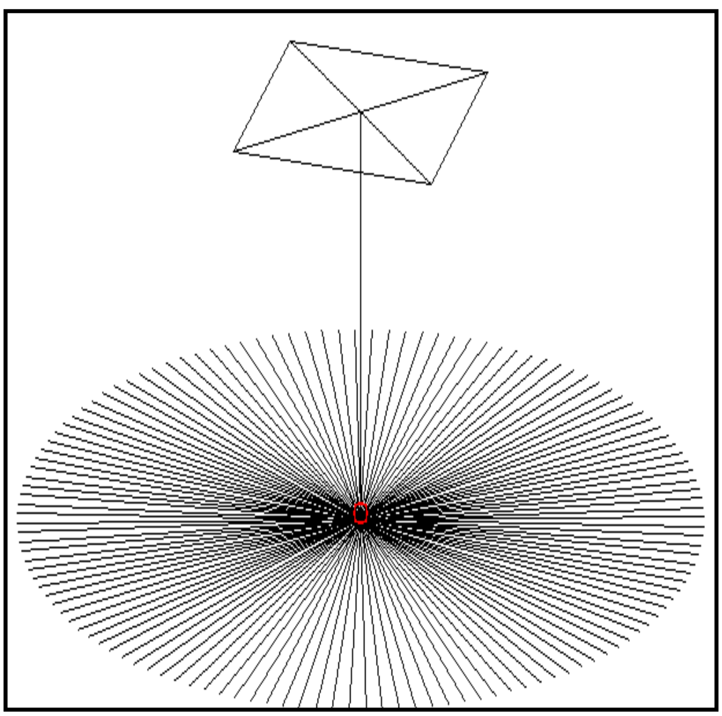

Figure 5.16 - typical 630m antenna.

The vertical is 15.24m high (50', 0.024λ) with 7.62m (25', 0.012λ) radial arms in the hat. The usual practice for very short verticals is to have a dense ground system which extends some distance beyond the edge of the top-hat and/or a bit longer than the height of the vertical.

5.10 Elevated ground systems

In many cases the ground under and near the antenna may not be suitable for a buried radial system. Systems with elevated wires are well known at HF, i.e. ground-plane verticals, but these systems typically use radials with lengths close to λ/4. For amateur installations this will not be possible but all is not lost. In the early days of radio ground systems it was recognized that an elevated system called a "counterpoise" or "capacitive ground", with dimensions significantly smaller than λ/4, could be very effective. Figure 5.17 shows an example of a counterpoise.

Here is an interesting quotation from Laport[26] regarding counterpoises:

> ***"From the earliest days of radio the merits of the counterpoise as a low-loss ground system have been recognized because of the way in that the current densities in the ground are more or less uniformly distributed over the area of the counterpoise. It is inconvenient structurally to use very extensive counterpoise systems, and this is the principle reason that has limited their application. The size of the counterpoise depends upon the frequency. It should have sufficient capacitance to have a relatively low reactance at the working frequency so as to minimize the counterpoise potentials with respect to ground. The potential existing on the counterpoise may be a physical hazard that may also be objectionable."***

Rectangular counterpoises, some with a coarse rectangular mesh, were common. A rather grand radial-wire counterpoise is shown in Figure 5.18. Amateurs also used counterpoises. Figure 5.19 is a sketch of the antenna used for the initial transatlantic tests by amateurs (1BCG) in 1921-22[27,28]. The operating frequency for the tests was about 1.3 MHz (230m). At 1.3 MHz λ/4 = 189' so the 60' radius of the counterpoise corresponds to ≈0.08λ.

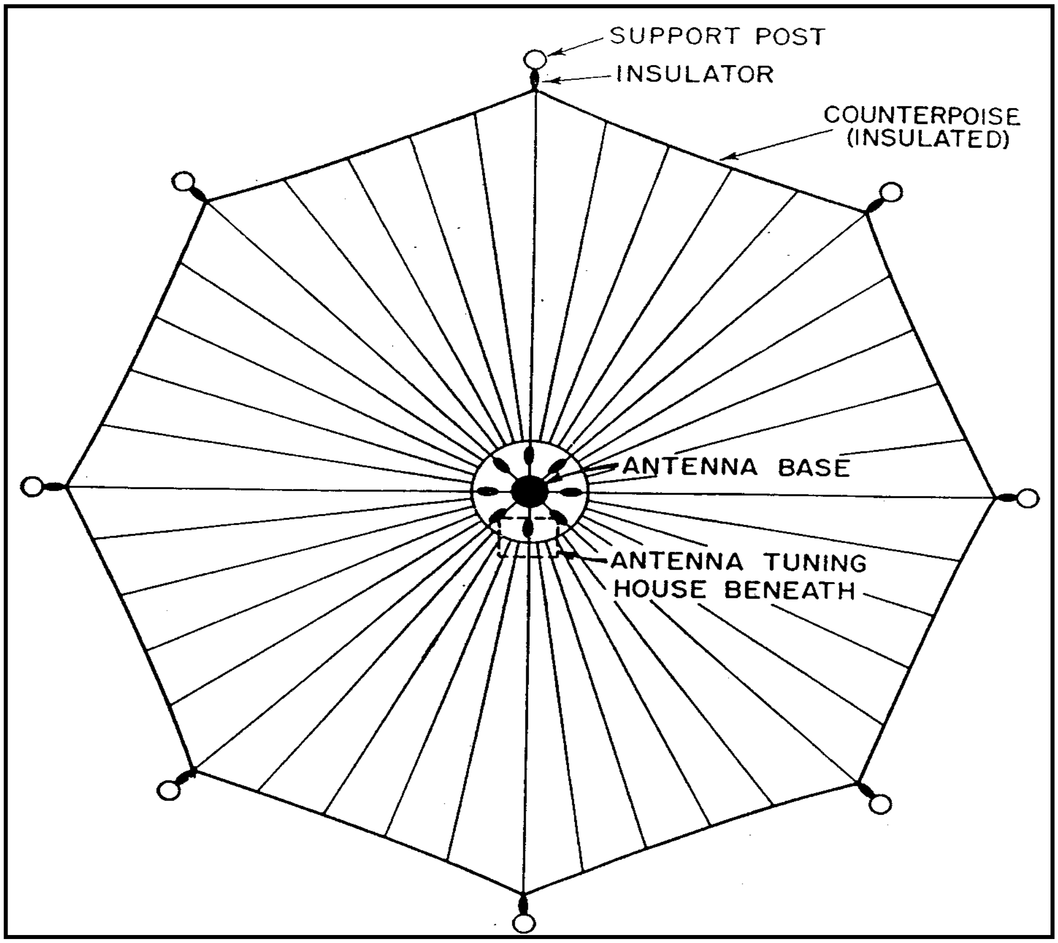

Figure 5.17 - A typical counterpoise ground system. Figure from Laport[26].

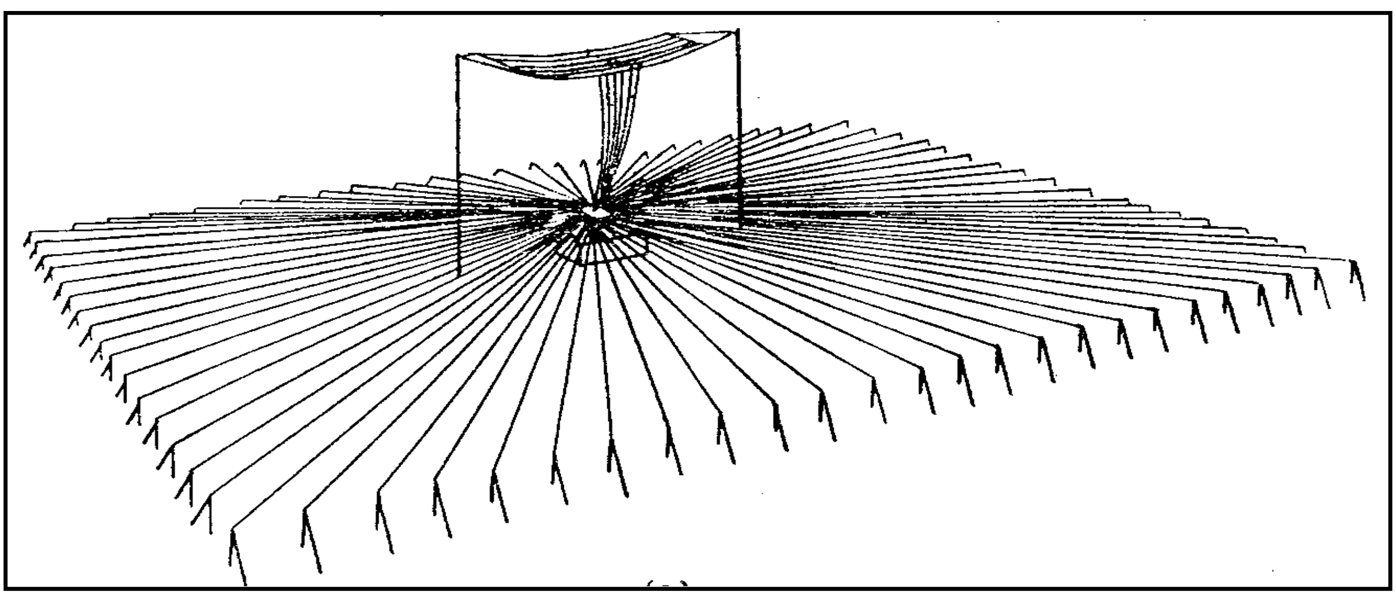

Figure 5.18- A very large LF elevated ground system. From [29].

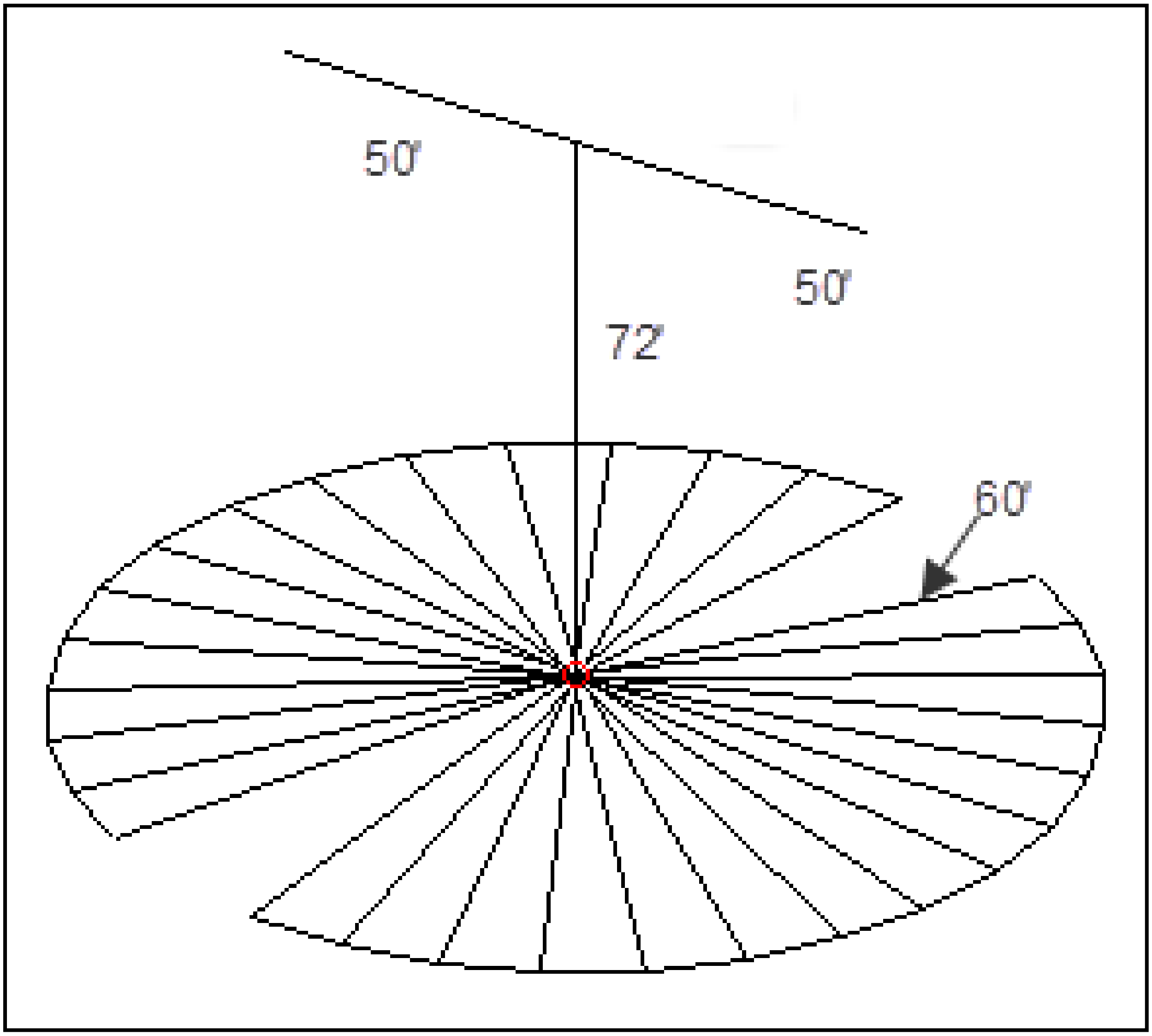

Figure 5.19 - EZNEC model of the 1BCG antenna.

Note that in all these examples a large number of radials were used.

We could replace the buried radial systems shown earlier with a counterpoise. For safety reasons the counterpoise will have to be at least 8' off the ground so that there is no danger of casual contact with the high potentials on the counterpoise while transmitting. For the moment we'll keep the height of the top of the vertical at 50' and assume a 16-wire, 25' radius umbrella with a skirt. This means that the total length of the vertical will be reduced from 50' to 42'. Figure 5.20 shows a comparison between 64 radial buried radial systems and 16 radial counterpoises (with skirt wires) as the radius of the ground system is varied from 10' to 50'. Note, in figures 5.20-5.24 the solid lines are for the counterpoise and the dashed lines are for the buried wire system.

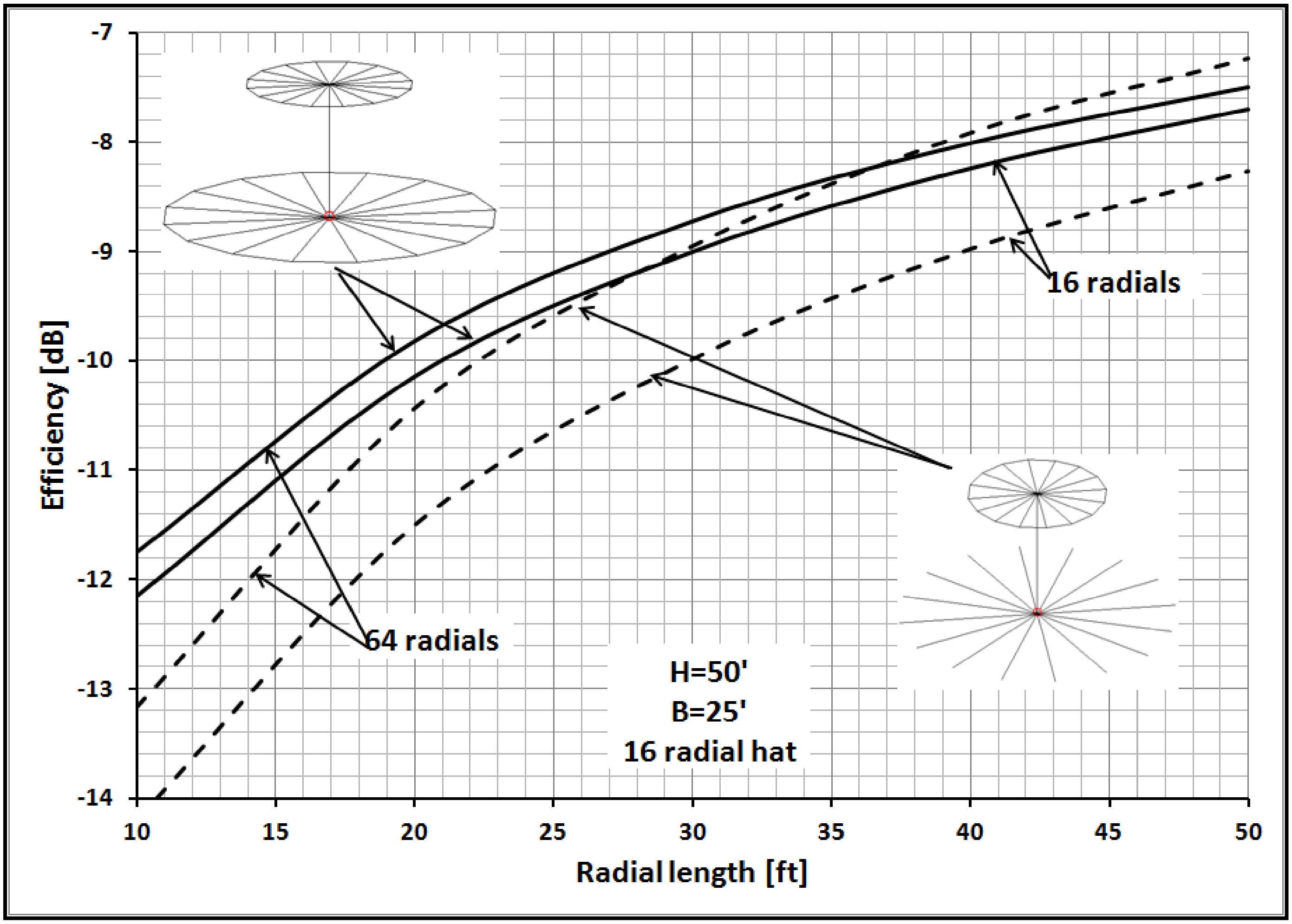

Figure 5.20 - A comparison between buried ground systems and counterpoises.

With 16 radials the counterpoise is superior for radii less than 100'. With 64 wires the counterpoise is better out to about 35' after which the buried system is better. That's great but we have to remember that the counterpoise is a large and very visible elevated structure while the buried or ground surface system is out of sight. In addition the counterpoise will require posts to support the ends, insulators at the outer periphery, a larger value for the loading inductor and an isolation choke for the feedline.

This is just for one example. To generalize to other configurations it's useful to look in more detail at what's happening to Rr and the loss components RL and Rg as we vary the ground system radius. Figures 5.21 through 5.23 show comparisons between a buried system and a counterpoise of Rr, RL and Rg as a function of radial length. Note, the top is constant at 50', for the counterpoise H=42' but for the buried system H=50'.

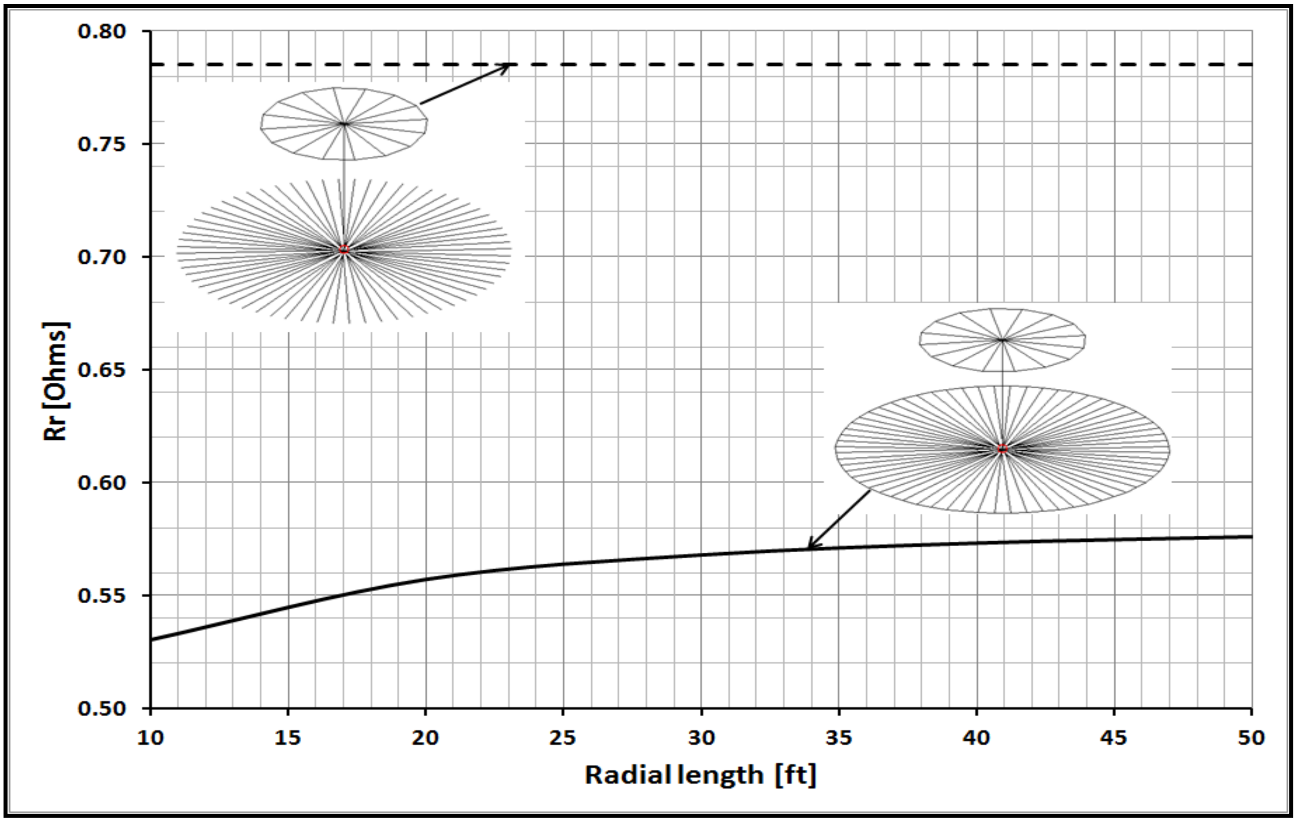

Figure 5.21 - Comparison for Rr between buried and counterpoise ground systems.

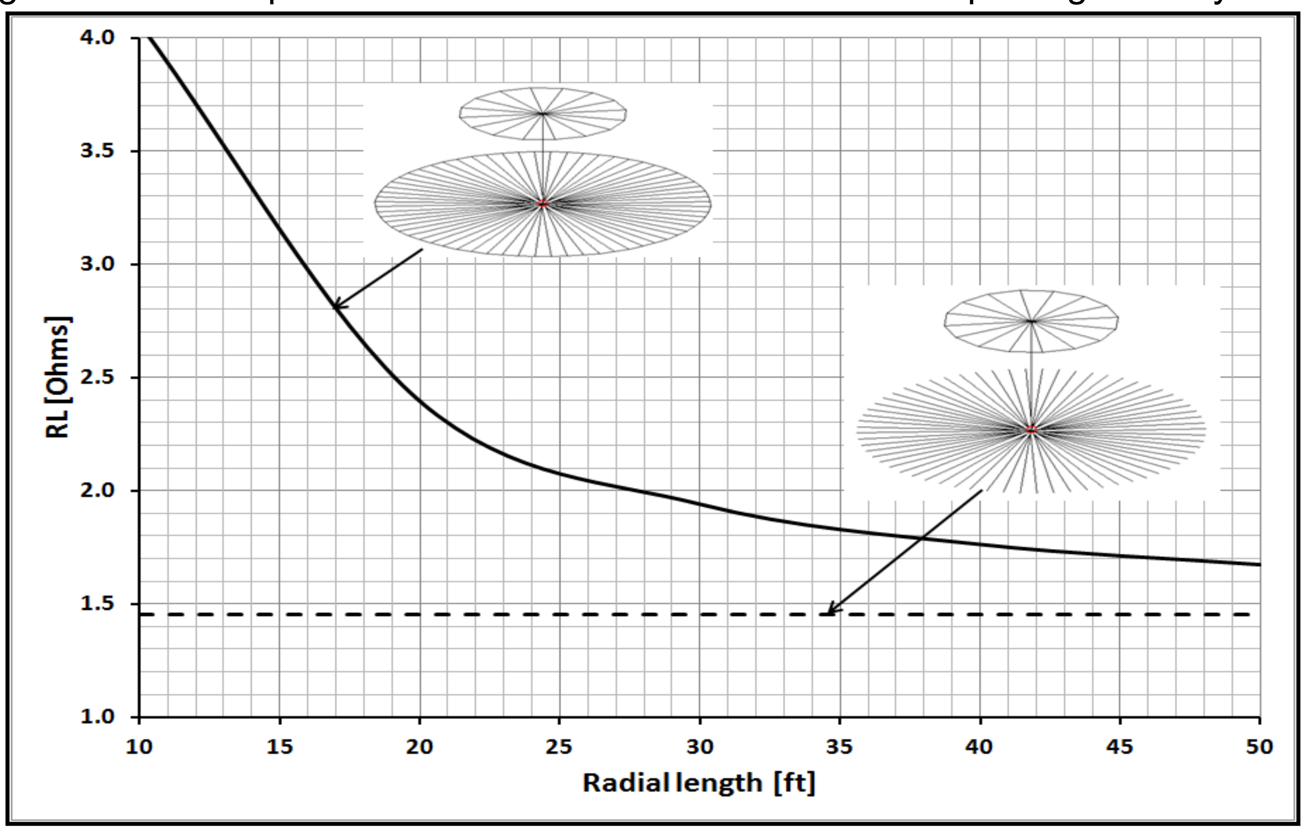

Figure 5.22 - Comparison for RL between buried and counterpoise ground systems.

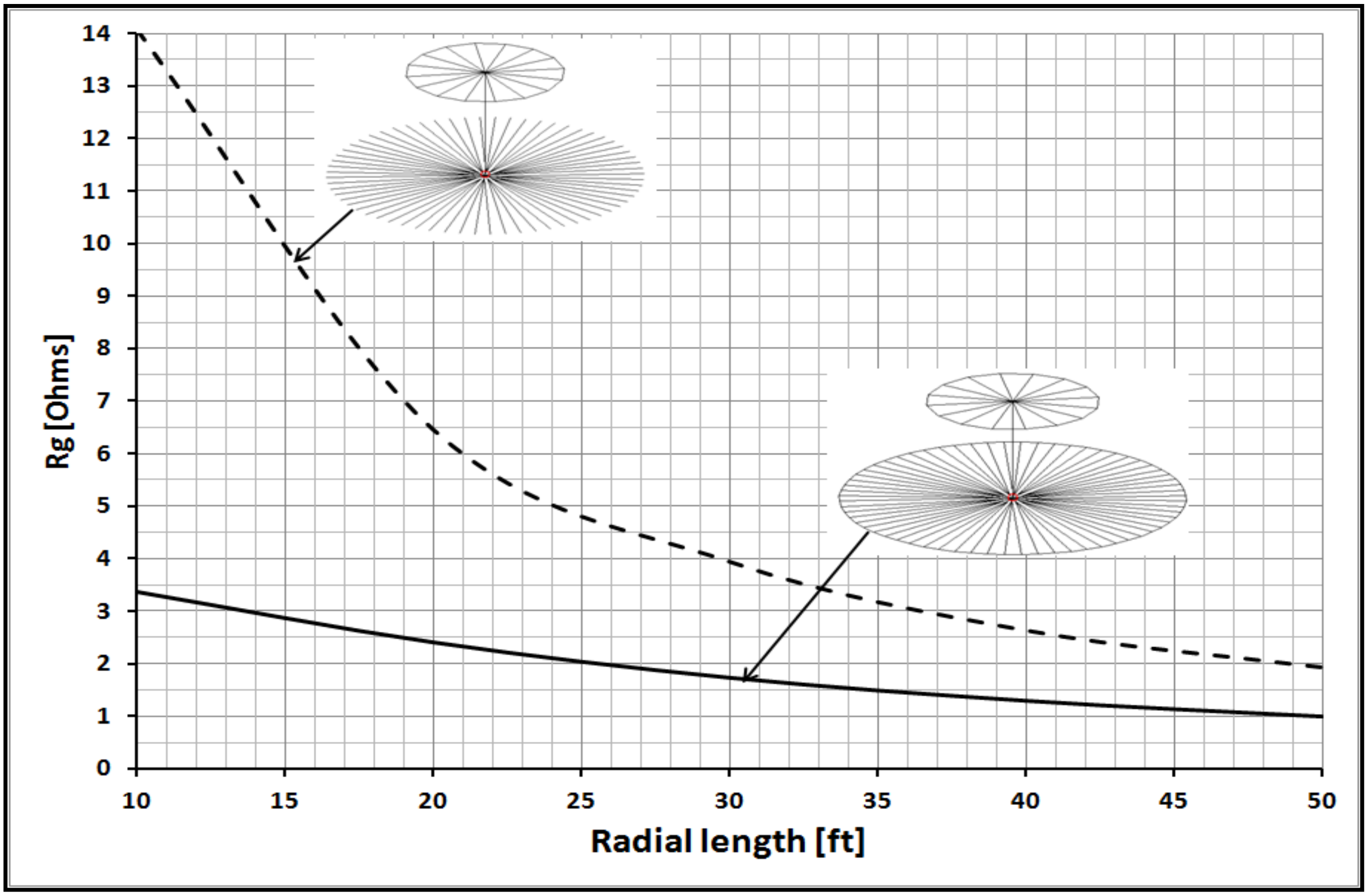

Figure 5.23 - Comparison for Rg between buried and counterpoise ground systems.

As the figures show, there are significant differences between the values and behavior of Rr, RL and Rg. As shown in figure 5.21, Rr for the vertical with a counterpoise is significantly lower than the ground based system. The reason for this is that the length of the vertical is shorter with the counterpoise (42'). For short verticals, Rr varies as the square of the length. For example comparing Rr' at 42' to Rr at 50', where Rr=0.79 Ω:

$$Rr' = \left(\frac{42}{50}\right)^2 0.79 = 0.56\,\Omega$$

Which agrees with what we see in figure 5.21. If we have a vertical taller than 50' the reduction in Rr from shortening it's length by 8' will be reduced but if the vertical is shorter than 50' the reduction in Rr will be greater.

Figure 5.22 shows how much larger RL becomes when the counterpoise is employed. This comes from two sources: first, Xc is larger due to the shorter length and second, the counterpoise itself introduces a reactance (Xcp) in series with Xc. For resonance, XL=Xc+Xcp, the loading inductor will be larger which increases RL. Increasing the radius of the counterpoise and the number of radial wires reduces Xcp.

The trends in figures 5.21 and 5.22 would seem to favor the buried ground system but figure 5.23 gives the opposite message. Figure 5.23 shows the counterpoise is indeed a very efficient ground system which has much lower ground losses as seen in the lower values for Rg. When you combine the values in figures 5.21-5.23 with equation (2) you get the result shown in figure 5.24 where the counterpoise has better efficiency for the smaller radii (<40').

In some ways however, the comparison just made between a ground base system and a counterpoise is a bit misleading because I assumed that the overall height of the vertical was limited to a fixed value (50'). If we keep H=50' and simply elevate the entire vertical 8' so that the top is now at 58' then the results are different as shown in figure 5.24 where the dashed lines are for buried wire ground systems and the solid lines are for a counterpoise with the same number of radials and radius.

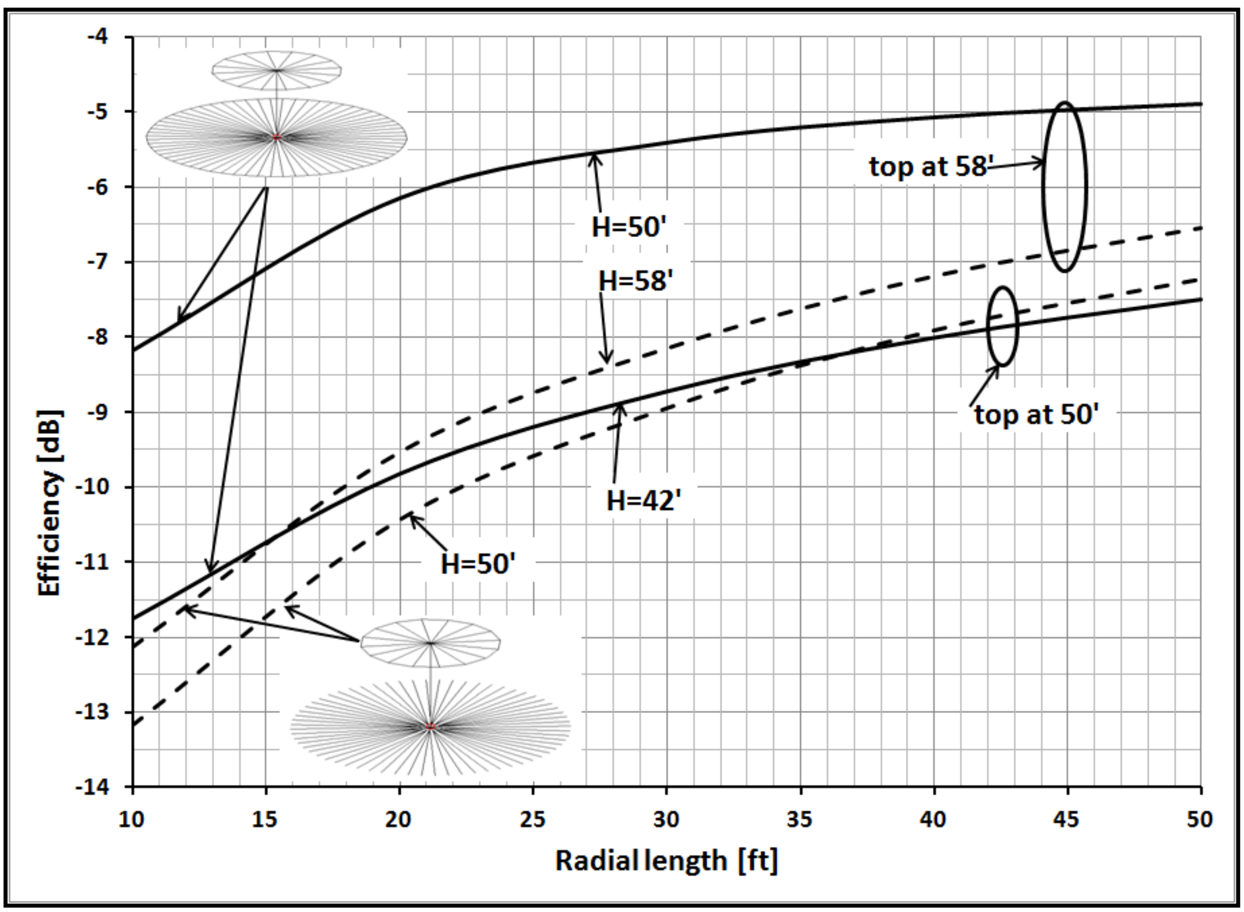

Figure 5.24 - Efficiency comparisons between counterpoise and buried wire ground systems for top heights of 50' and 58'.

What we see in the upper two traces in figure 5.24 is that simply raising the entire antenna up 8' and substituting the counterpoise for the buried wire system, there is a large improvement in efficiency and the counterpoise is superior at all radii up to 50' or more. You could argue that if we can raise the entire antenna 8' we could just as well have simply increased H to 58' and retained the buried ground system but as we can see in figure 5.24, the counterpoise is still better at least out to radii of 50'.

The point being made here is that if you have reasonable heights for the vertical and lots of capacitive top-loading but very restricted room for the ground system, then a counterpoise may be the best option. But we have to be careful in drawing general conclusions from the limited examples given here. There are many variables: ground characteristics, height of the top of the vertical, height of the bottom of the vertical, the amount of top-loading, the number of wires in the counterpoise, the radius, etc. The choice between buried wire and counterpoise ground systems is not obvious! The considerable mechanical complexity, vulnerability to ice damage and visual impact of a counterpoise may also militate against it. This choice has to be made on a case-by-case basis and will probably require modeling with NEC4 software.

5.11 Summary

This discussion has shown many examples for ground systems from which we can draw the following conclusions:

...the ground system can be elevated, on the surface or buried. Properly designed any of these will work...

...a significant number of radials must be used, 16 or more for elevated systems and 50 or more for ground based systems...

...the radial length should be somewhat greater than the height of the vertical and extend beyond the outer edge of the top-loading...

...a counterpoise may be a good choice for the ground system...

References

[1] Severns, Rudy, N6LF, "Experimental Determination of Ground System Performance for HF Verticals, Part 1, Test Setup and Instrumentation", QEX, January/February 2009, pg. 21

[2] Severns, Rudy, N6LF, "Experimental Determination of Ground System Performance for HF Verticals, Part 2, Excessive Loss in Sparse Radial Screens", QEX, January/February 2009, pg. 48

[3] Severns, Rudy, N6LF, "Experimental Determination of Ground System Performance for HF Verticals, Part 3, Comparisons Between Ground Surface and Elevated Radials", QEX, March/April 2009, pg. 29

[4] Severns, Rudy, N6LF, "Experimental Determination of Ground System Performance for HF Verticals, Part 4, How Many Radials Does My Vertical Really Need?", QEX, January/February 2009, pg. 38

[4] Severns, Rudy, N6LF, "Experimental Determination of Ground System Performance for HF Verticals, Part 5, Ground Systems For Multi-band Verticals", QEX, January/February 2009, pg. 19

[6] Severns, Rudy, N6LF, "Experimental Determination of Ground System Performance for HF Verticals, Part 6, 160m Vertical Ground System", QEX, January/February 2009, pg. 15

[7] Severns, Rudy, N6LF, "Experimental Determination of Ground System Performance for HF Verticals, Part 7, Ground Systems With Missing Sectors", QEX, January/February 2010, pg. 18

[8] Severns, Rudy, N6LF, "An Experimental Look at Ground Systems for HF Verticals", QST, Mar 2010, pg. 30

[9] R. Severns, N6LF, MEASUREMENT OF SOIL ELECTRICAL PARAMETERS AT HF, ARRL, QEX magazine, Nov/Dec 2006, pp. 3-9

[10] R. Severns, N6LF, "Verticals, Ground Systems and Some History," *QST*, Jul 2000, pp 38-44

[11] Abbott, Frank, Design of Optimum Buried-Conductor RF Ground System, IRE proceedings, July 1952, pp. 846-852

[12] Rudy Severns, N6LF, Radial System Design and Efficiency in HF Verticals, 2008, available at www.antennasbyn6lf.com

[13] Rudy Severns, N6LF, The Case of Declining Beverage-on-Ground Antenna Performance, QST June 2016, pp. 38-41

[14] Rudy Severns, N6LF, The Case of Declining Beverage-on-Ground Antenna Performance, QEX Jul/Aug 2016, pp. 7-18

[15] Rudy Severns, N6LF, Determination of Soil Electrical Characteristics Using a Low Dipole, QEX Nov/Dec 2016, pp. 5-8

[16] Brown, Lewis and Epstein, "Ground Systems as a Factor in Antenna Efficiency," *Proc. IRE* , Jun 1937, pp. 753-787

[17] George H. Brown, A Theoretical and Experimental Investigation of the Resistance of Radio Transmitting Antennas, University of Wisconsin, Madison, Ph.D. thesis, June 1933.

[18] George H. Brown and Ronold King, High-Frequency Models in Antenna Investigations, IRE proceedings, Volume 22, Number 4, April 1934, pp. 457-480

[19] George H. Brown, The Phase and Magnitude of Earth Currents Near Radio Transmitting Antennas, IRE proceedings, Volume 23, Number 2, February 1935, pp. 168-182

[20] H.E. Gihring and G.H. Brown, General Considerations of Tower Antennas for Broadcast Use, IRE proceedings, Volume 23, Number 4, April 1935, pp. 311-356

[21] G. H. Brown, A critical Study of the Characteristics of Broadcast Antennas as Affected by Antenna Current Distribution, IRE proceedings, Volume 24, Number 1, January 1936

[22] Rudy Severns, N6LF, Radiation Resistance Variation with Radial System Design, QEX Jul/Aug and Sept/Oct 2015.

[23] J. Stanley, "Optimum Ground Systems for Vertical Antennas," *QST*, Dec 1976

[24] R. Sommer, N4UU, "Optimum Radial Ground Systems", QST August 2003, pp. 39-43

[25] Al Christman, K3LC, "Maximum-Gain Radial Ground Systems For Vertical Antennas", NCJ magazine, March/April 2004, pp. 5-10

[26] Laport, Edmund, Radio Antenna Engineering, McGraw-Hill, 1952. You can find this one free on-line by Googling Edmund Laport.

[27] Burghard, George, "Station 1BCG", QST, February 1922, pp. 29-33

[28] Kelley and Hudson, "Hams Span the Atlantic on Shortwave!", QST, December 1996, pp. 28-30

[29] Admiralty Handbook of Wireless Telegraphy, His Majesty's Stationary Office, 1932, pg. 799, Figure 452

Chapter 6

Design and Fabrication of High-Q Tuning Inductors for LFMF Antennas

6.0 Introduction

The electrically small antennas typical of 630m and 2200m amateur installations can be represented by the equivalent circuit in figure 6.1.

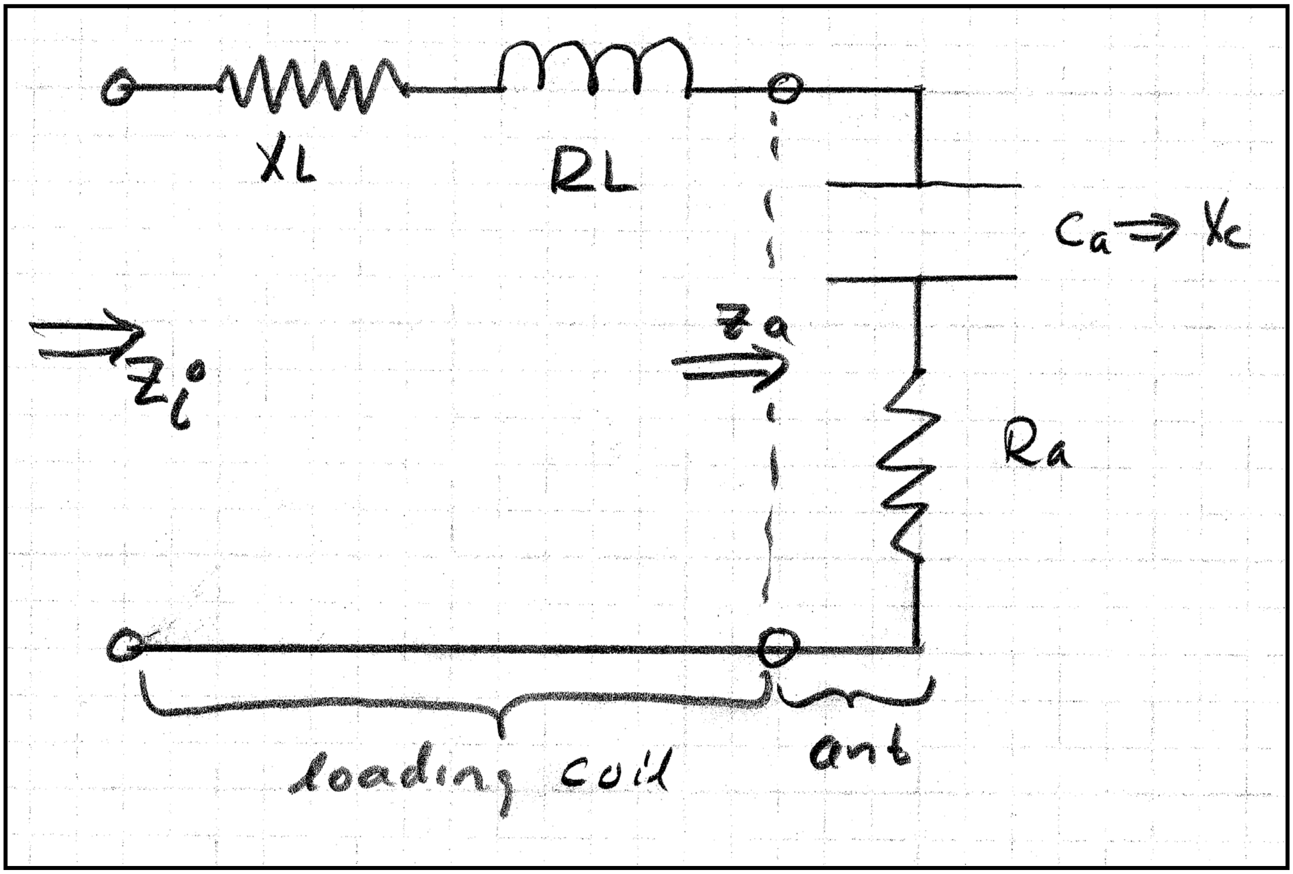

Figure 6.1 - LFMF antenna equivalent circuit.

The antenna is simply a capacitor in series with a resistor. The capacitive reactance (Xc) is very large and the series resistance (Ra) is small. Most of Ra comes from ground system and conductor losses. The radiation resistance (Rr) is typically only a very small part of Ra. To supply power to the antenna it's necessary to transform the highly reactive feedpoint impedance of the antenna to a resistive value compatible with the feedline, usually Ri=Zo=50Ω. The series inductor (XL) performs two functions: canceling the input reactance (+XL-Xc=0) and a means to transform Ra to the desired value for Ri which is often done with an adjustable tap on the inductor although there are other possibilities. Unfortunately any practical inductor will have losses (represented by RL) which can reduce efficiency substantially.

The radiation efficiency (η) can be expressed as:

$$\eta = \frac{Rr}{Ra+RL} = \frac{Rr}{Rr+RL+Rg+Rc+\cdots} \qquad (6.1)$$

Where Rr is the radiation resistance, RL is the inductor loss resistance, Rg represents ground system loss, Rc represents conductor loss and miscellaneous other losses. Usually Rr will be very small, a fraction of an ohm, but RL and Rg are typically much larger. In most cases RL and Rg dominate the efficiency so every effort must be made to minimize these losses. For an inductor:

$$XL = 2\pi fL \qquad (6.2)$$

$$Q \equiv \frac{XL}{RL} \rightarrow \quad RL = \frac{XL}{Q} \qquad (6.3)$$

Where f is the operating frequency and L is the required inductance.

Increasing loading inductor Q is a primary tool for maximizing efficiency.

This chapter addresses the design and fabrication of high Q inductors using materials commonly available. Historically many different coil constructions have been tried but the focus here is on cylindrical single layer air wound coils using round wire because these are common and practical. Examples of flat spiral or "pancake" inductors, toroidal inductors, non-circular coil forms and basket-weave windings will be shown but not discussed in detail.

The discussion covers a lot of ground at considerable length and it's fair to ask "is all this verbiage actually useful?" Many articles and even whole books on inductor design, along with free software CAD programs[1], already exist. From a practical point of view do we really need more? It turns out that the design of tuning inductors for 630m and 2200m is significantly more complicated than typical HF inductors. For example, LF, MF and HF inductor designs must take into account both skin and proximity effects:

- The resistance of a wire is ≈Rdc at low frequencies but as the frequency is increased the resistance increases very substantially, this is called "skin effect".

- When a conductor is wound into a coil the current in one turn induces loss in adjacent turns, this is called "proximity effect".

These two effects impact the design when high Q is desired but there is another effect, "self-resonance", which is rarely important at HF but frequently limits Q at 630m and 2200m:

- Coil inductors behave very much like transmission lines with a multitude of harmonically related resonances. The lowest self frequency resonance (SRF) can significantly impact both inductance and Q.

All three effects must be included in the design of LFMF inductors. Fortunately we are able to accurately calculate all these effects but some of the equations are complex and all of them are interrelated making pencil and paper calculations impractical. Spreadsheets can be used but that's practical for only the most dedicated algebraphiles. Fortunately Brian Beezley, K6STI, has created an inductor design program[1], COIL, which takes into account all the complexity while keeping it out of sight. Both COIL and spreadsheet calculations were used for this chapter.

6.1 How much inductance?

When designing a new inductor the very first question is "what value of inductance (L) is needed and at what frequency (f)?" To resonate the antenna enough XL is needed to cancel the capacitive reactance (Xa) at the feedpoint, i.e. XL=Xa (figure 6.1):

$$\mathbf{L = \frac{XL}{2\pi f}} \qquad (6.4)$$

When f is in MHz L will be in μH.

To estimate the needed inductance we can convert the values for Xa derived in chapter 3 to inductance in uH as shown in figures 6.2 and 6.3. In these figures an italic *L* identifies the total length in feet of the top-loading wire. It should be pointed out that although figures 6.2 and 6.3 assume a "T" with a single top-wire, the values for the loading inductor would be the same for any capacitive top-loading structure which provides the same amount of capacitive loading. The shape of the hat is not what's important, it's the added capacitance!

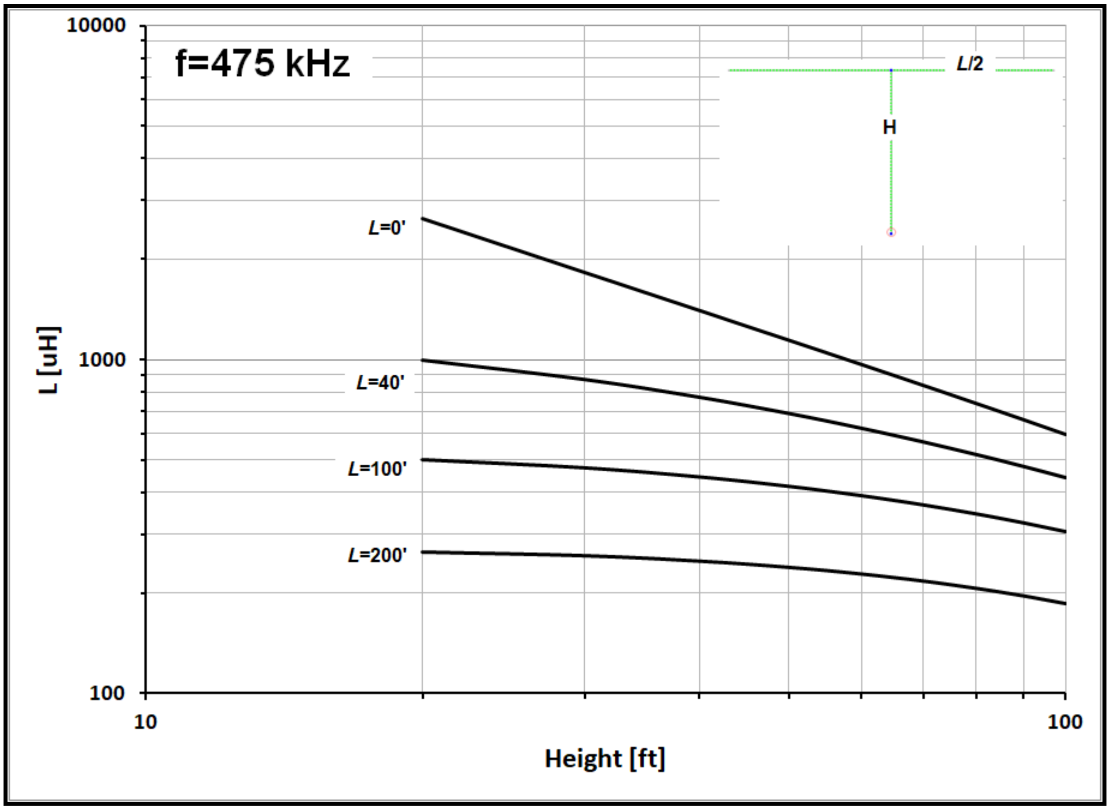

Figure 6.2 -Tuning inductor inductance for resonance at 475 kHz.

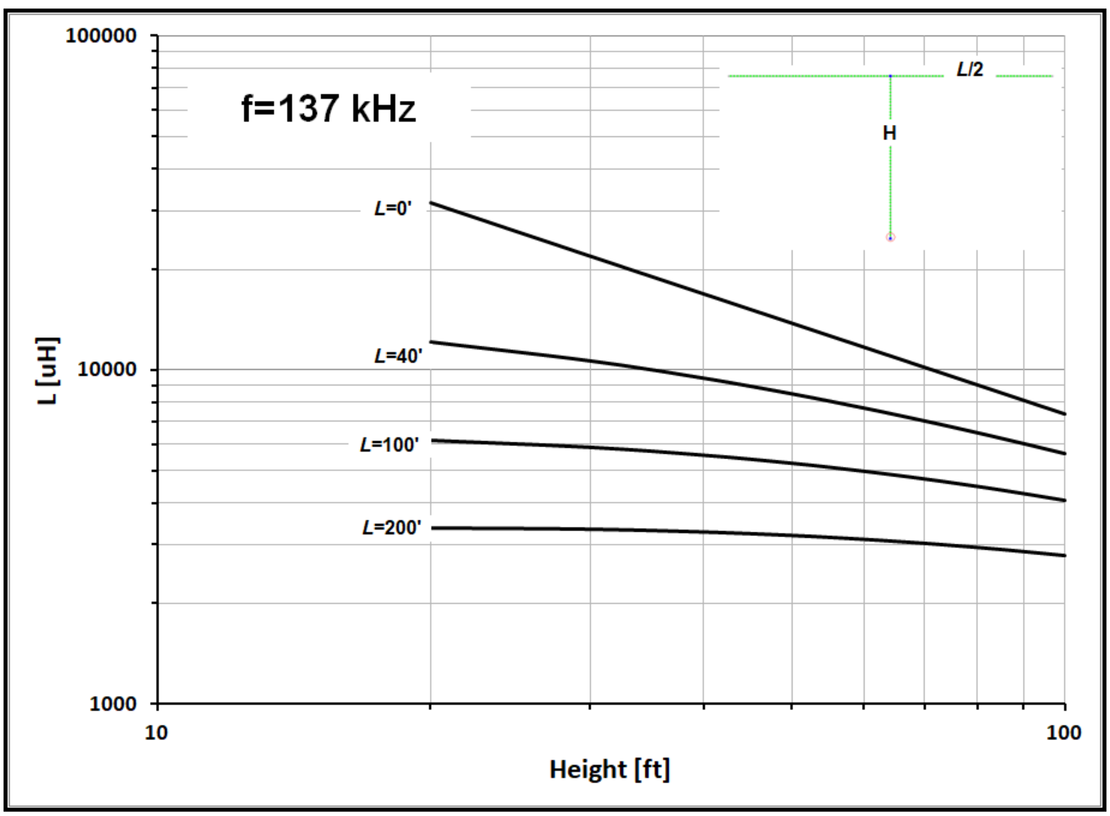

Figure 6.3 - Tuning inductor inductance for resonance at 137 kHz.

From figures 6.2 and 6.3 we see:

- At 475 kHz, for resonance, L ≈ 200µH →1000µH.
- At 137 kHz, for resonance, L ≈ 3mH→20mH.

6.2 Definitions

Some useful variables can be defined with the help of figure 6.4.

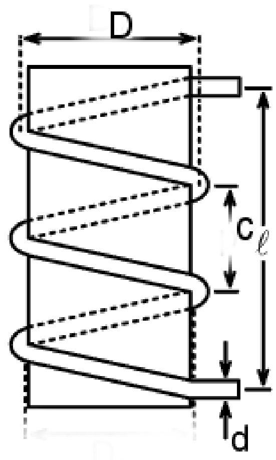

Figure 6.4 - inductor dimensions (From Knight)

c → winding pitch , center-to-center spacing between turns

d → diameter of the winding conductor

D → diameter of the winding. Wire center-to-wire center

l → coil length

L→ actual inductance at the operating frequency

lw→ length of the winding conductor

N → number of turns in the winding

Design graphs and equations can be made more general using geometric ratios as variables. For example:

- (d/c) → conductor diameter/turn center-to-center spacing ratio
- (l/D) → coil length/diameter ratio

In practice the normal ranges for these ratios are: 0.2<l/D<10 and 0.3<d/c<0.7. l/D=0.2 represents a very short winding with a large diameter. l/D=10 represents a long tubular coil with a small diameter. Figure 6.5 uses these ratios to illustrate how variable the inductance can be when a coil is wound in different ways using the same length of wire (lw). For this example lw=100' and d=0.081" (#12 wire). The dashed contours correspond to constant values of l/D. Note that for all values of d/c, maximum inductance occurs at l/D=0.45. In this example Lo varies from ≈50 to 460 uH, a range of 9:1 with the same piece of wire. How the coil is wound really matters! Note the use of Lo for the inductance, L. Lo is the very low frequency (1 kHz) inductance. Due to self resonance L may be much larger and vary over an even greater range. This will be addressed a following section.

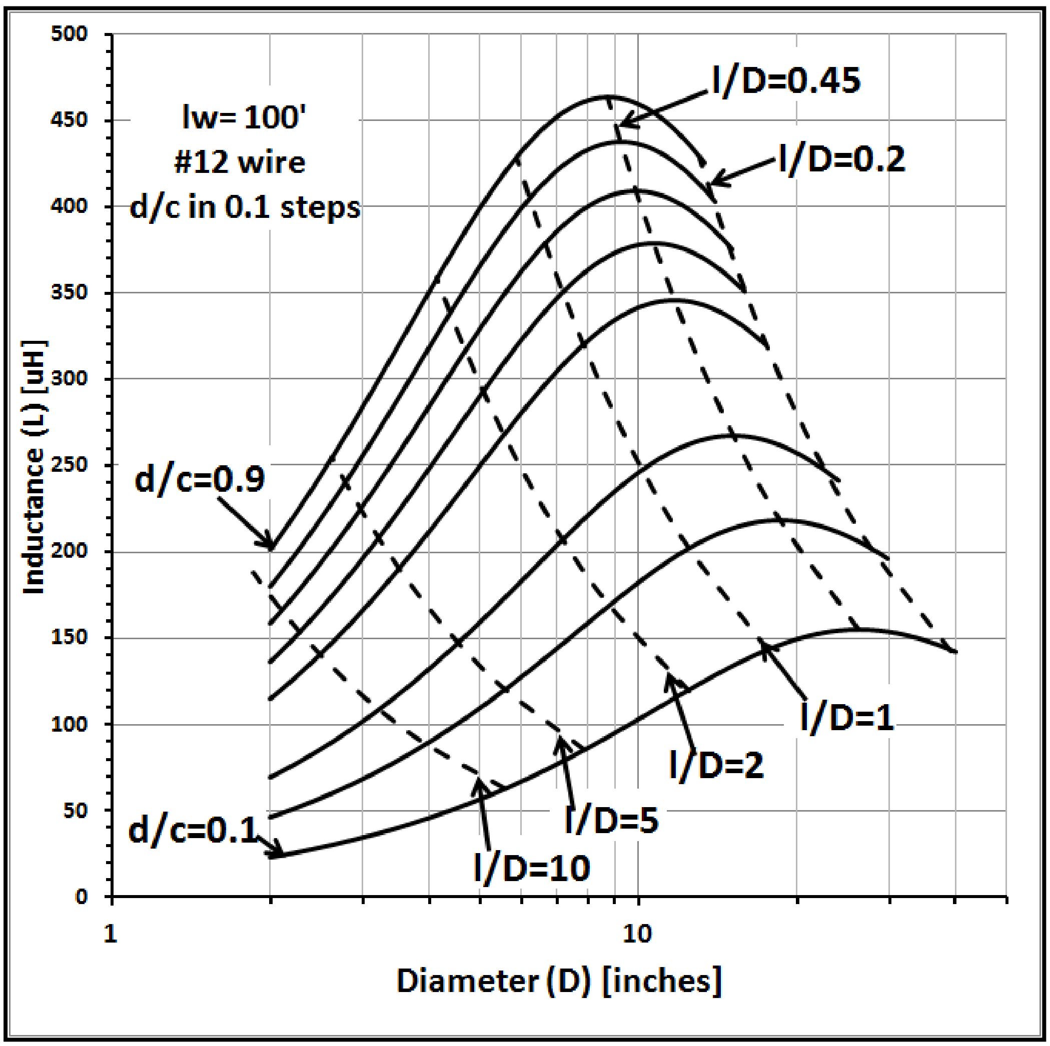

Figure 6.5 - Inductance versus diameter with d/c and l/D as parameters.

6.3 Practical inductors

A new design starts with the value for the inductance and the operating frequency (f). It's a good idea to make L 5-10% larger to allow for adjustment using taps on the coil but you don't want to go overboard as the unused portion of the coil still affects fr potentially lowering Q. Keep in mind that the actual inductance will vary with frequency so the value chosen needs to be the correct value <u>at the operating frequency</u>.

Because coil loss (RL) and value of L depend on frequency it is necessary to know f at the beginning. For amateurs, the values for f are very limited (2200m→135.7-137.8 kHz, 630m→472-479 kHz) so in the following discussion f=137 kHz is used for 2200m coils and f=475 kHz for 630m coils.

To fabricate an inductor some details are needed:

- diameter (D),
- winding length (l),
- number of turns (N),
- wire size or diameter (# or d)
- turn spacing (d/c)
- length of the wire in the winding (lw).

Some of these can be chosen initially and the rest derived either from graphs or COIL.

Insulated THHN wire is frequently used for windings. Because it's wide use in home wiring this wire is relatively inexpensive and readily available. Other types of wire can certainly be used. Three sizes are commonly used: #14 (0.064"/1.63mm), #12 (0.081"/2.05mm) and #10 (0.102"/2.59mm). For winding design some wire dimensions are needed. These are listed in table 1.

Table 1 - Typical wire dimensions

wire #	nominal diameter	diameter over insulation	assumed diameter	maximum turns/inch	d/c maximum
10	0.102"	0.165"	0.17"	5.9	0.60
12	0.081"	0.117"	0.12"	8.3	0.69
14	0.064"	0.098"	0.10"	10.2	0.64

The "nominal" diameters come straight out of a standard wire table. Using a micrometer one often finds wire diameters are a tad under the specification, perhaps a mil or more, which usually doesn't matter too much. The diameters over the insulation are measured values from a several samples but there can be considerable variation so don't be surprised if you can't get as many turns on a form as expected or the winding is not as long as predicted. The "assumed" diameter allows for some additional insulation thickness and is the value used for the graphs and calculations that follow. The maximum turns/inch and d/c maximum are based on the assumed diameters. With insulation the maximum turns/inch will be less that with bare wire. However, with bare wire some spacing will be needed between the turns and it's usually not practical to have spacings much smaller than typical insulation thickness.

6.3.1 Turn spacing warning!

Do not use tightly wound windings with no air space between turns!

Here's the reason for that warning.

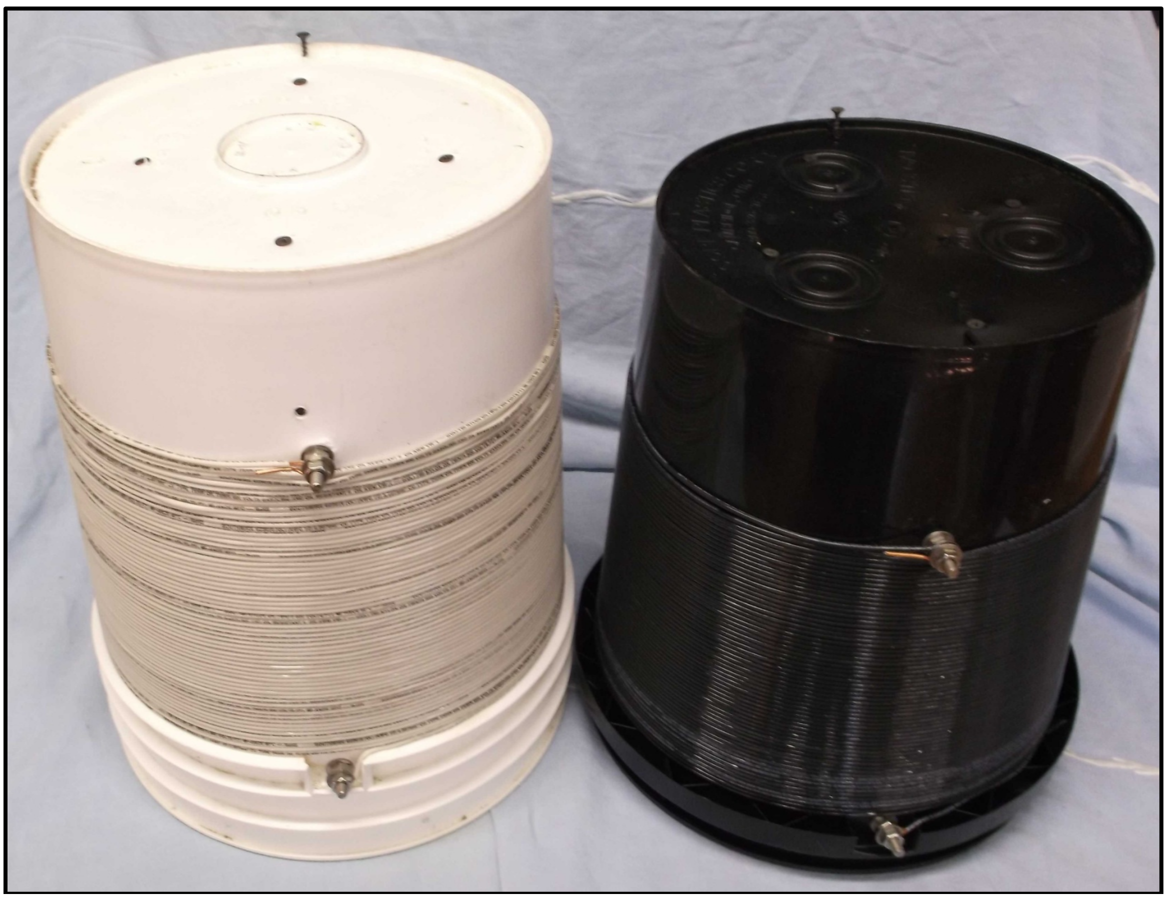

Figure 6.6 - Plastic bucket inductor examples.

Either insulated or bare wire can be used for the winding. Q measurements comparing the same coil with insulated wire versus bare wire[11,12] shows very little difference except for very tight (closely packed) windings being used. This effect was demonstrated by experiment. To illustrate tight versus loose windings two inductors (Lo≈1mH) were fabricated with insulated wire on plastic buckets. The black bucket had a very tight winding and the white bucket a somewhat looser winding. Two versions were wound on the white bucket and the Q measured. In the first example the inductor was wound with new #12 THHN wire directly off the original spool so it was smooth allowing a very tight winding like that shown on the right in figure 6.6. After completing some measurements the winding on the white bucket was unwound and as a result of handling the wire became a bit lumpy. This wire was then rewound on the same bucket but this time the winding was significantly longer, ≈+1", with small air gaps between the turns as shown on the left in figure 6.6.

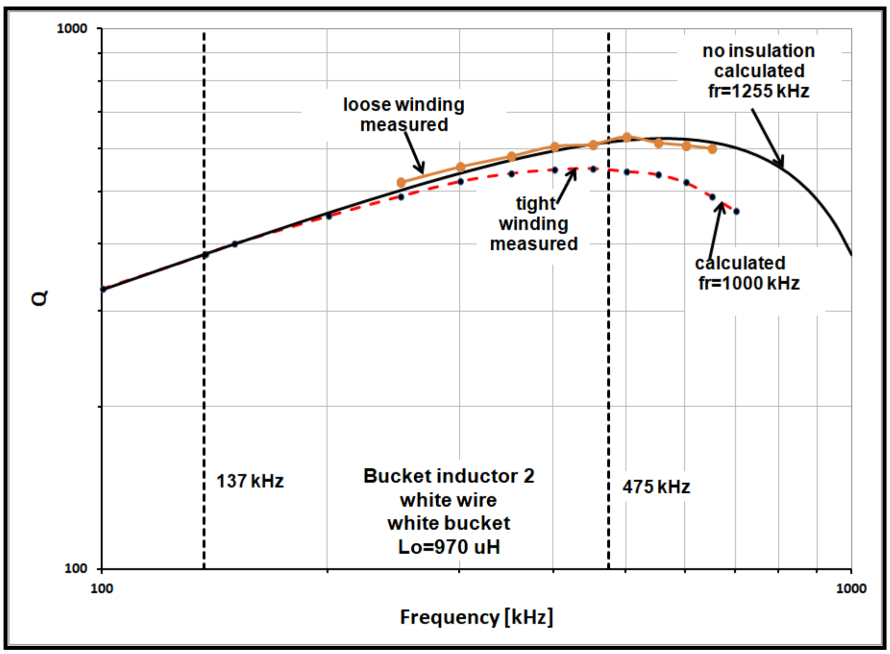

Figure 6.7 - Comparison between tight and loose windings.

Figure 6.7 compares the Q measurements with tight and loose windings. There is a substantial difference! In a tight winding the insulation on each turn is pressed closely to the turns on either side which increases winding capacitance and reduces self-resonant frequency (fr) which in turn reduces Q as f approaches fr. It should be noted that the reduction in Q with an insulated wire winding is due to a lower fr, not dielectric loss in the insulation.

Here is some more experimental work. Figure 6.8 is a trial "bucket" inductor for a variometer. Q measurements and COIL predictions are graphed in figure 6.9. Note

SFR differs by almost a factor of 3! Note the internal variometer coil was not present for this test.

Figure 6.8 - Bucket inductor.

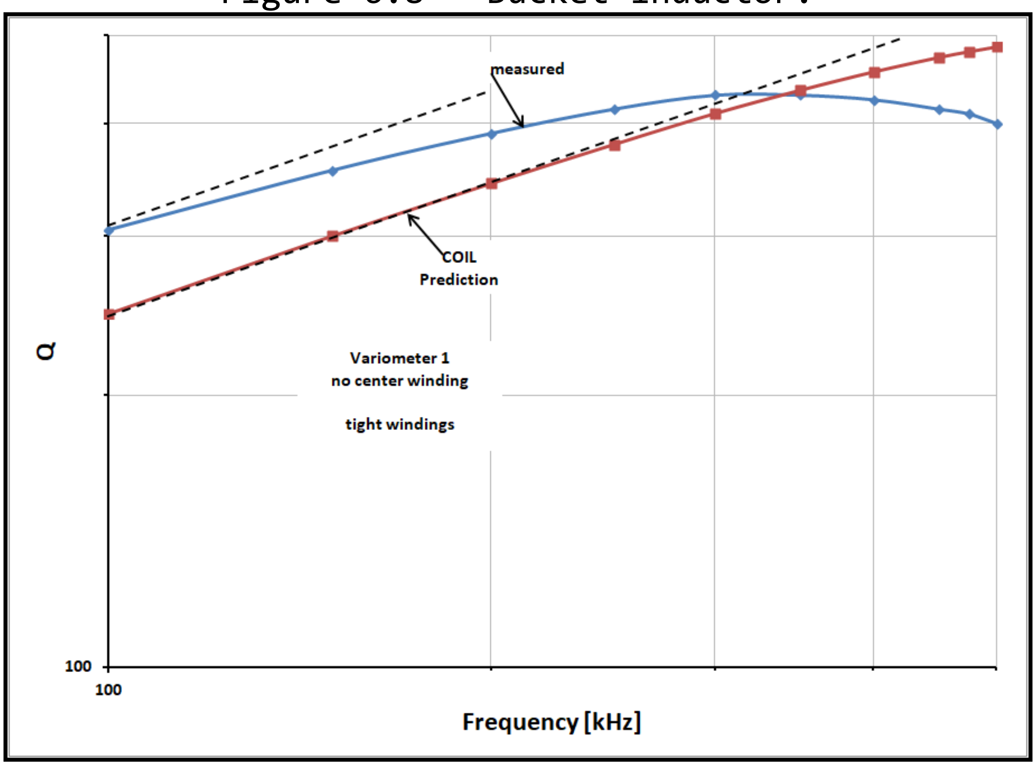

Figure 6.9 - Bucket inductor Q measurements versus prediction.

Now, same kind of bucket, same wire, about the same inductance but using a spaced winding shown in figures 6.10 and 6.11.

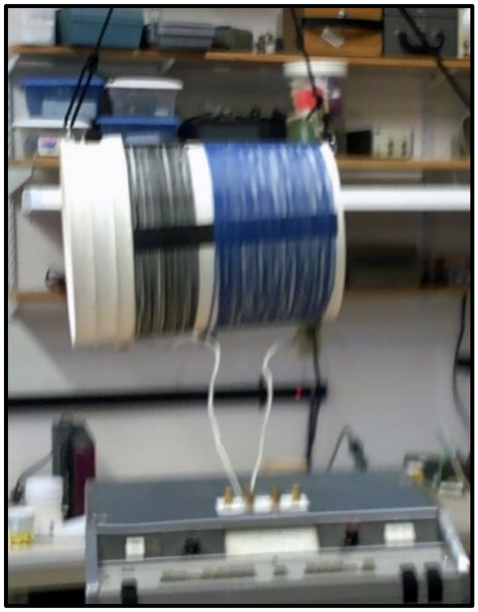

Figure 6.10 - Spaced winding

With the following comparison graph.

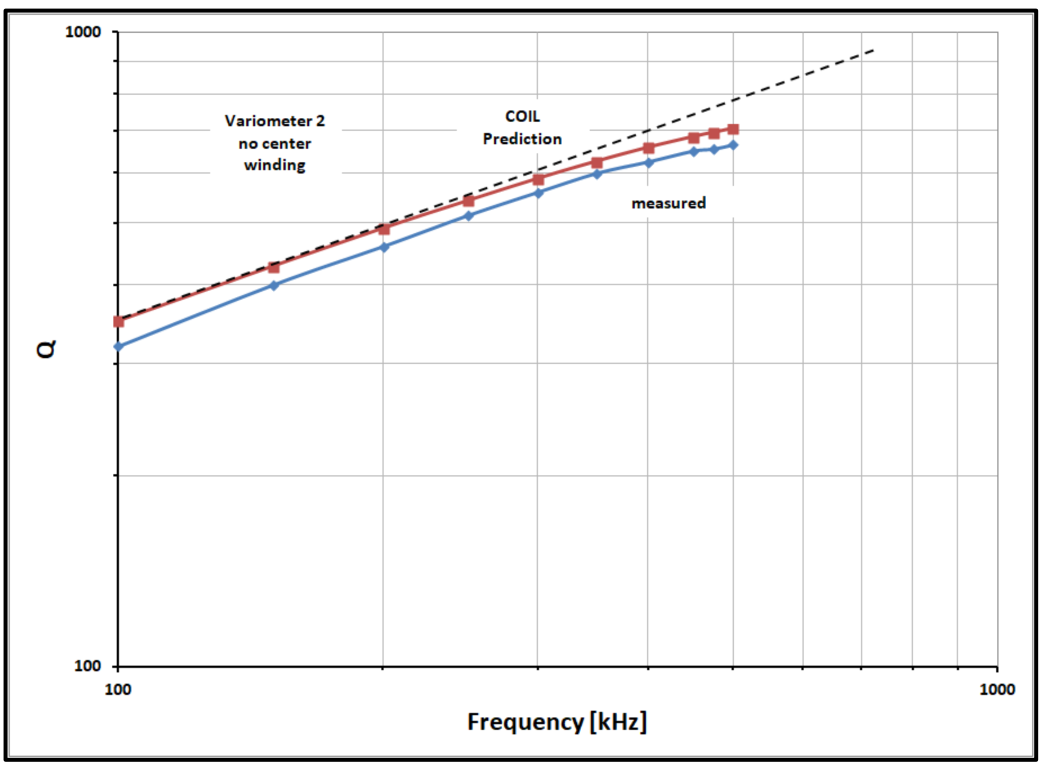

Figure 6.11 - spaced winding inductor example.

Enough said, space the turns at least a small amount.

6.3.2 Bucket inductors

Plastic buckets are inexpensive and universally available in many different sizes (1, 2, 5, 7, etc gallon) as shown in figures 6.12. Five gallon buckets are probably the most often used size for coil forms. Better quality buckets are made from High Density Polyethylene (HDPE). At LFMF PVC and HDPE have very little dielectric loss. Not all buckets are HDPE, while the HDPE is common there are many lower quality buckets on the market. Look on the bottom of a prospective bucket, there should be a small triangle with 2 in it and HDPE underneath. The wall thickness of the bucket in mils should also be there.

Figure 6.12 - Typical buckets.

A plastic bucket can be used as a coil form but some thought must be given to the winding process. Because the bucket is smooth plastic with some taper and wire insulation is also smooth plastic, the wire tends to slide around as the coil is moved. There is a simple trick which helps keep the turns in place with the desired turn-to-turn spacing: attach several (6-8) vertical strips of double sided mounting or carpet tape vertically before winding. These are the dark strips in figure 6.10. This does a good job of holding the wire in place. A simple 2"X4" frame like that shown in figure 6.14 as a "winding machine" will make the job a lot easier. Figure 6.13 shows 1/2" iron pipe fittings attached to the top and bottom of the bucket. Small square plywood blocks were used on the inner sides of the bucket bottom and lid to stiffen them and anchor the screws. The stanchion bases are attached with screws through the bucket into the blocks.

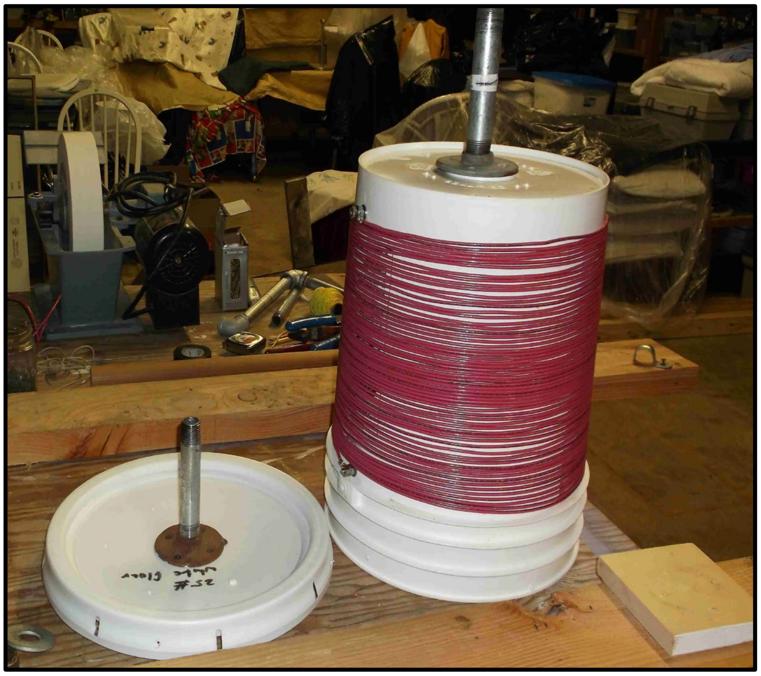

Figure 6.13 - Bucket modification for winding.

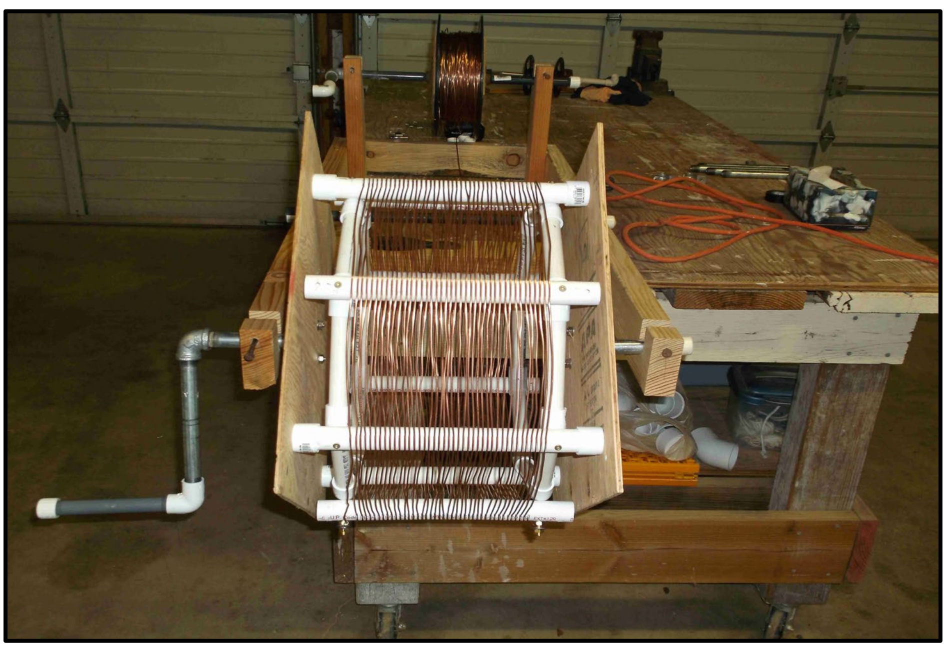

Figure 6.14 - Winding machine example.

Buckets come in a wide variety of sizes but once a choice is made both the diameter and the maximum winding length are predetermined which limits the possible inductance values. A 5 gallon bucket example can be used to illustrate these limits.

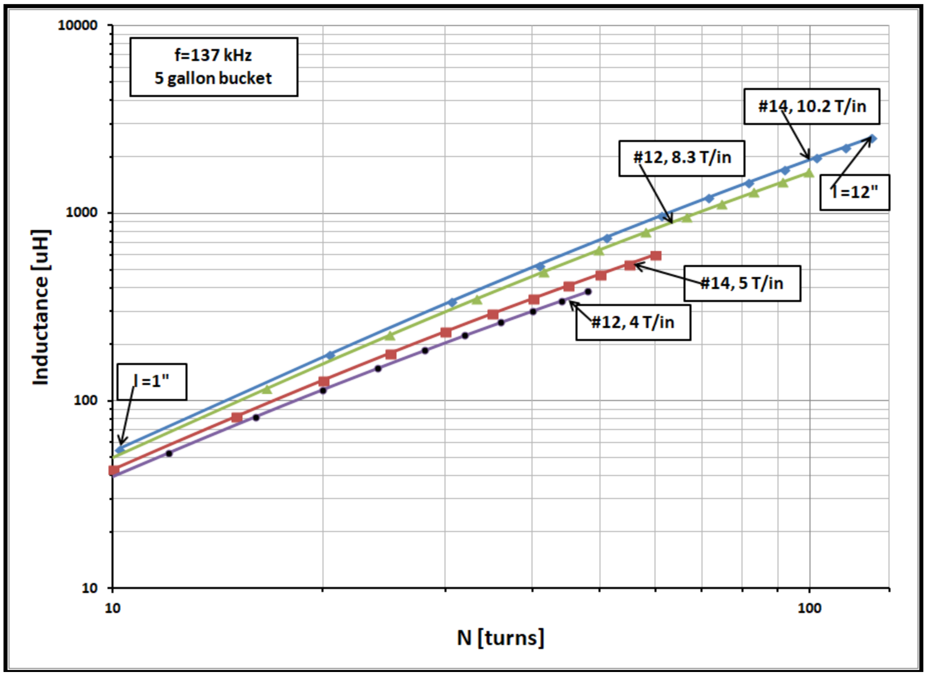

Figure 6.15A - 5 gallon bucket example L versus N at 137 kHz.

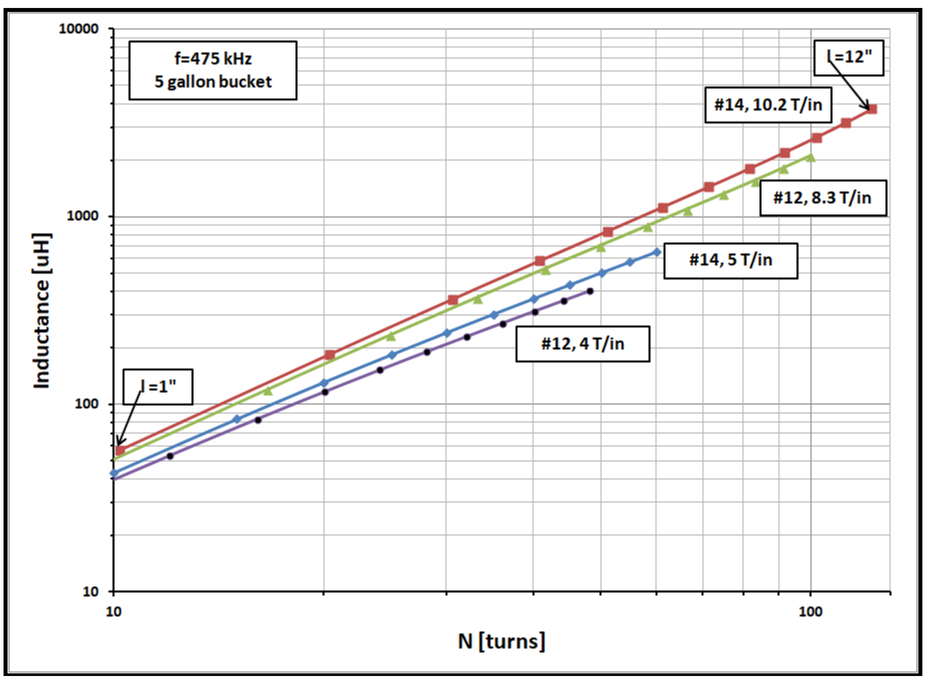

Figure 6.15B - 5 gallon bucket example L versus N at 475 kHz.

Figure 6.15 shows the relationship between N and L at 137 kHz and 475 kHz for two wire sizes (#12 & #14). For each wire size there are two turn spacings (Turns/inch).

For #14 wire 10.2 T/in represents about the tightest possible winding. 5 T/in represents wider spacing which is used to reduce proximity loss increasing Q. For #12 wire 8.3 and 4 T/in are used. The graphs illustrate that larger wire and greater spacing means fewer turns because the maximum winding length is constrained to <12". But as shown in figure 6.16, larger wire and greater turn spacing yield higher Q.

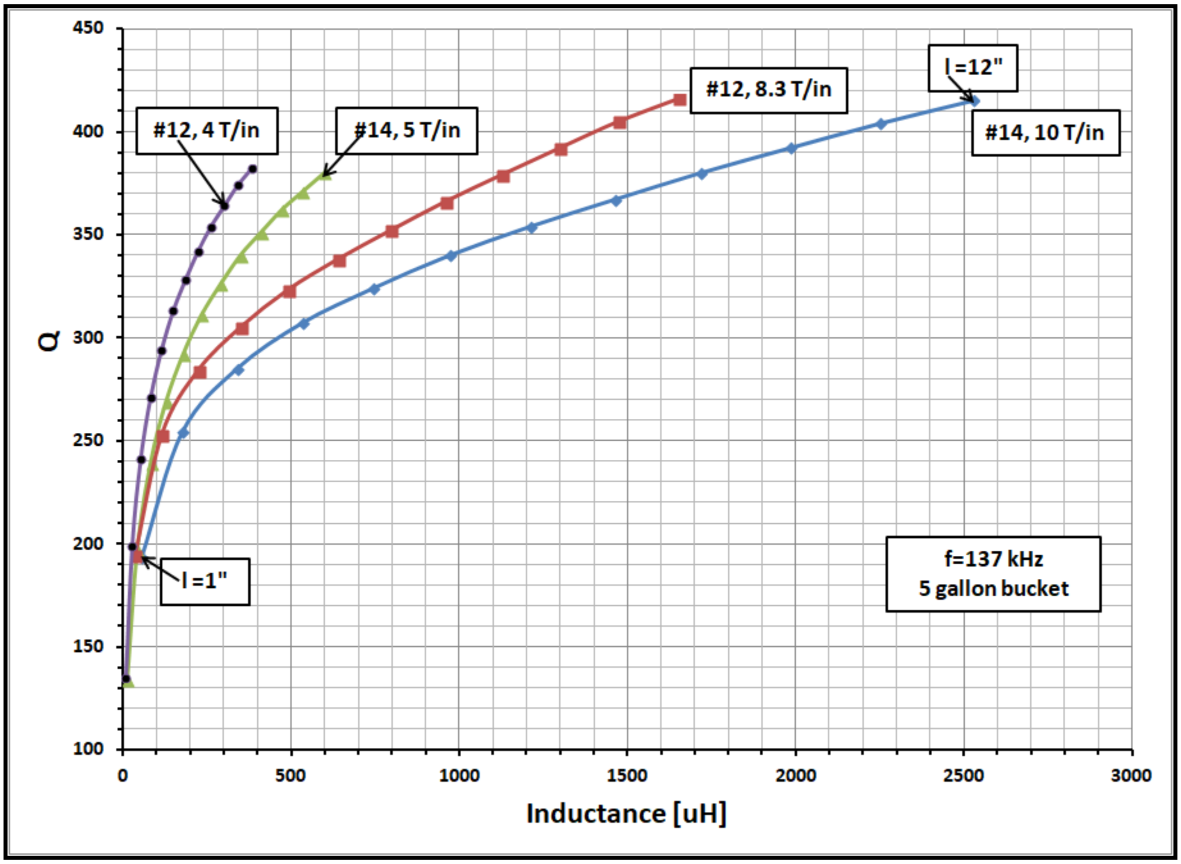

Figure 6.16A - Inductor Q at 137 kHz.

Using #14 wire tightly wound an inductance of 2.5 mH at 137 kHz and almost 4 mH at 475 kHz can be obtained but the Q for that inductor will be modest: ≈420 at 137 kHz and ≈520 at 475 kHz. As the graphs show, either increasing the wire size and/or the turn spacing increases Q but reduces the maximum inductance. For example at 475 kHz, increasing the wire size from #14 to #12 increases Q (@L=2mH) from ≈550 to ≈610 but the maximum L is now <2.1 mH. For L=400 uH, if we go from closely spaced # 14 to wide spaced #12 Q goes from 500 to 780 which is a considerable improvement, but L is now constrained to <400 uH. The user has to keep these trade-offs in mind when choosing inductor parameters.

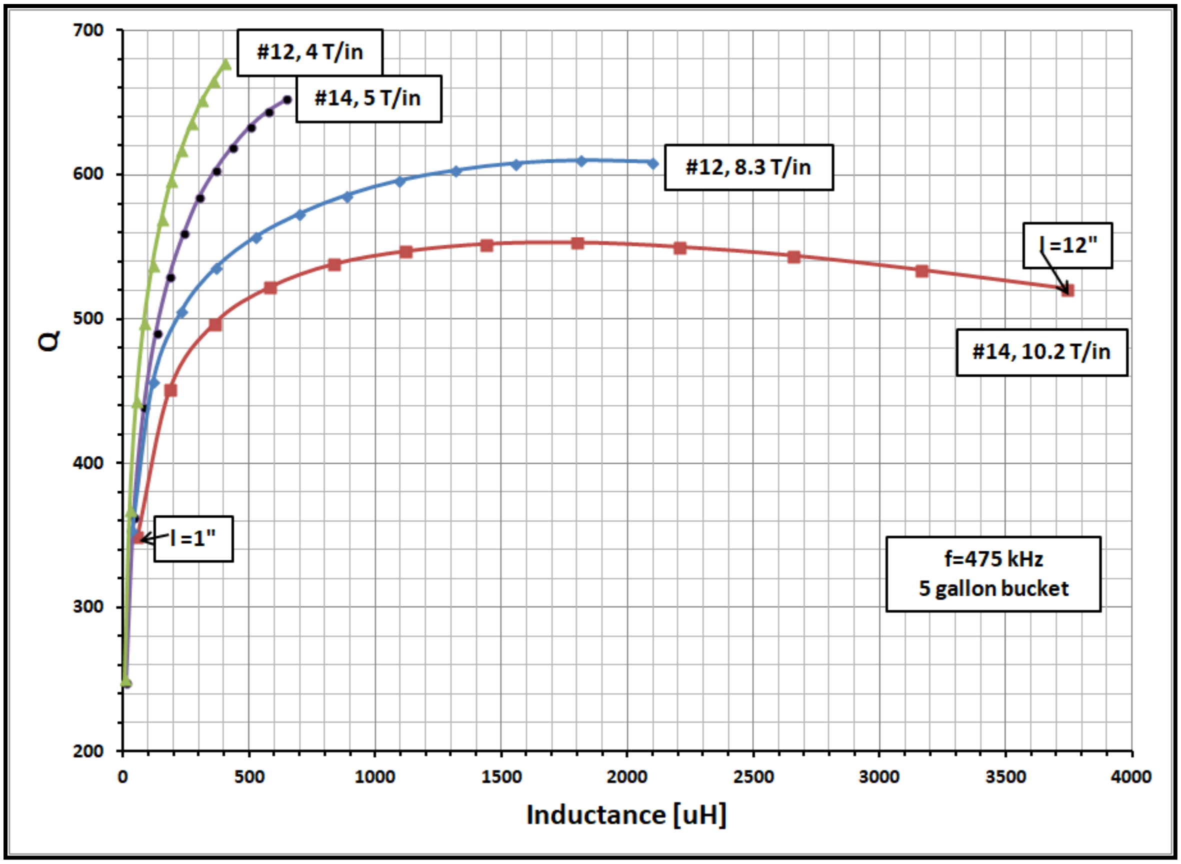

Figure 6.16B - Inductor Q at 475 kHz.

6.3.3 Maximum Q inductors

If a really high-Q inductor is wanted then most likely a special coil form will be needed. PVC pipe and fittings make it very easy to fabricate a large form with arbitrary dimensions. Figure 6.17 shows two examples of PVC cage coil forms. In figure 6.17A the octagonal rings are 1/2" pipe joined with 45° elbows. Because the winding compresses the form it is usually not necessary to glue the rings which makes fabrication much easier! The eight vertical supports used 3/4" pipe with slots cut at intervals (=c) to hold the wire. As an alternative the turns in figure 6.17B were constrained with double sided tape.

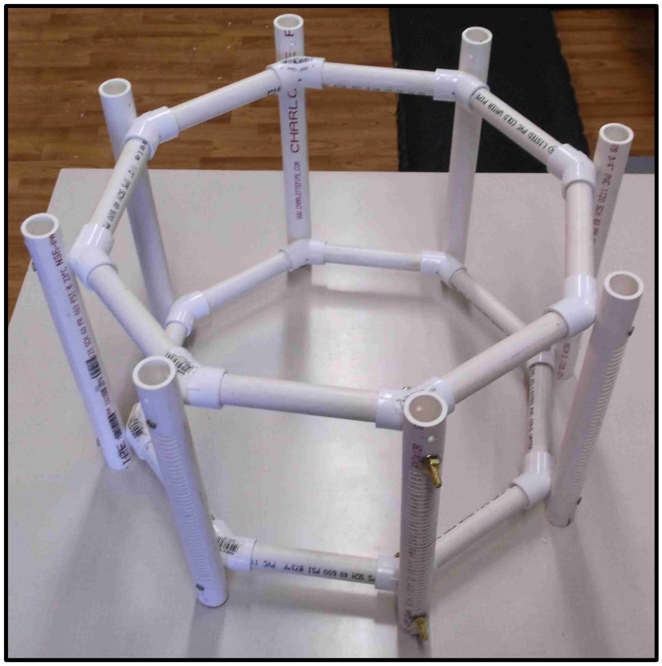

Figure 6.17A - PVC cage coil form.

Figure 6.17B - N1DAY PVC cage coil form.

Figure 6.18 - Cutting the wire slots.

Figure 6.18 shows a practical way to cut slots. Temporarily screwing the supports to a board makes it <u>much easier</u> to align all the slots and mounting holes.

A clever example of a very light weight inductor (≈18"X18") is shown in figure 6.19. Pat, W5THT, fabricated the coil form by wrapping F/G mat around a cardboard tube, then impregnating the mat with epoxy and when it had cured, soaking the assembly in water to soften the cardboard for removal, leaving a thin shell on which he wound 1/2" wide copper tape. A protective covering of paint was then applied. The light weight of the inductor allowed him to hoist it to the top of his vertical where it joined the capacitive hat. One could also purchase a sheet of thin plastic and roll it to make the coil form.

Figure 6.19 - Pat W5THT, foil wound lightweight inductor.

Optimizing Q

Using COIL a modeling experiment optimizing Q was done with very interesting results. First, Q's of 500 to over 900 were readily obtained. Second, the diameter (D) associated with each optimized test value was found to change only a small amount over the full range of inductance values for a given wire size and frequency. Third, the spacing ratio (i.e. wire diameter/turn-to-turn spacing, d/c) was found to have a very small range, d/c≈0.30-0.32, for every example over the entire range of inductance and wire size! d/c can be converted to the more useful parameter p ("turns-per-inch") for each wire size as shown in table 2.

Table 2 - Turns/inch for d/c=0.31

wire size	p [T/in]	rounded p [T/in]
14	4.84	5
12	3.83	4
10	3.07	3

Table 3 shows the averaged diameters associated with optimized inductors.

Table 3 Averaged values for D and p (turns/in) for optimum Q

	#14	#14	#12	#12	#10	#10
frequency	T/in	D	T/in	D	T/in	D
137 kHz	5	28"	4	32"	3	36"
475 kHz	5	15"	4	17"	3	19"

Table 3 suggests coil diameters of 1.5' to 3', these are not small coils! Using these values for wire size, diameter, T/in and frequency, the Q's were recalculated and graphed as shown in figure 6.20. When compared to the original "optimized" Q values, the Q values derived using the averaged dimensions were within a few percent. The difference was not worth worrying about! What this means is that right up front, for a given frequency and wire size you know the coil diameter and the turns spacing. The only missing information is the number of turns (N), the required coil form length (l) and the total length of wire needed for the winding (lw). N can be determined from COIL or figure 6.21.

The length of the winding (the minimum length of the coil form!) is simply:

$$\mathbf{l = N/p \ \ [inches]} \qquad (6.5)$$

And the total length of the wire in the winding is:

$$\mathbf{lw = \frac{N\pi D}{12} \ \ [ft]} \qquad (6.6)$$

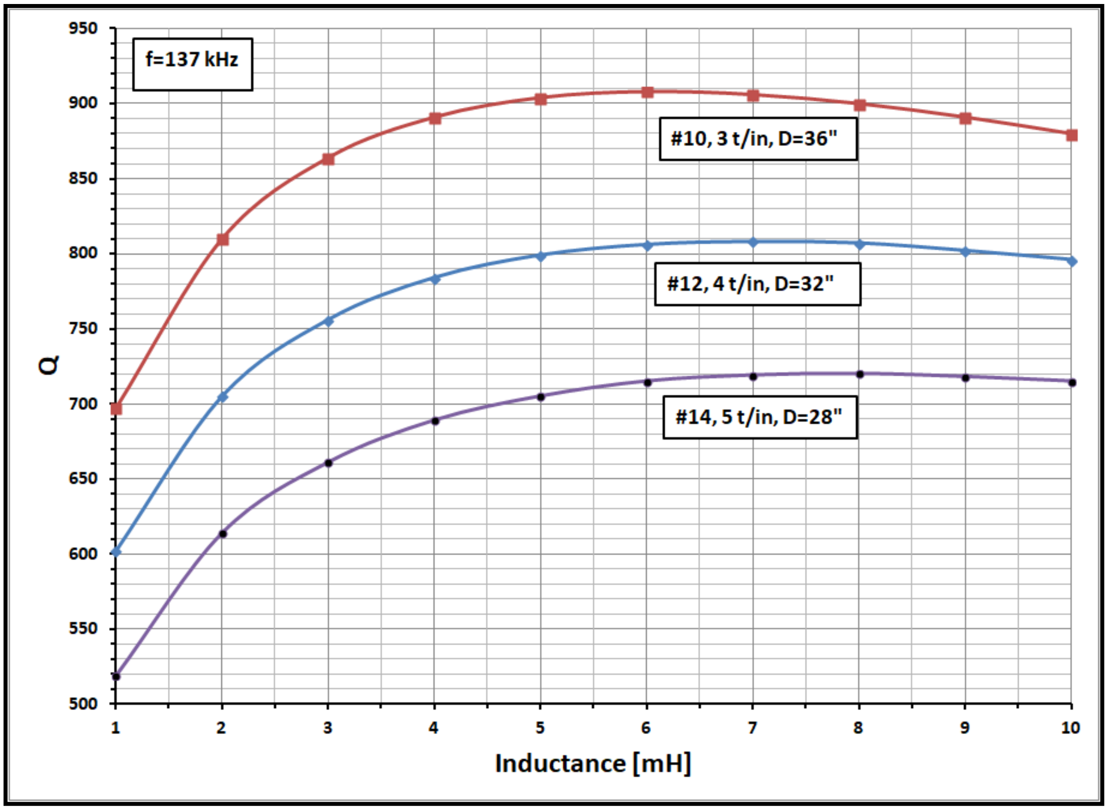

Figure 6.20A - Optimized Q at 137 kHz.

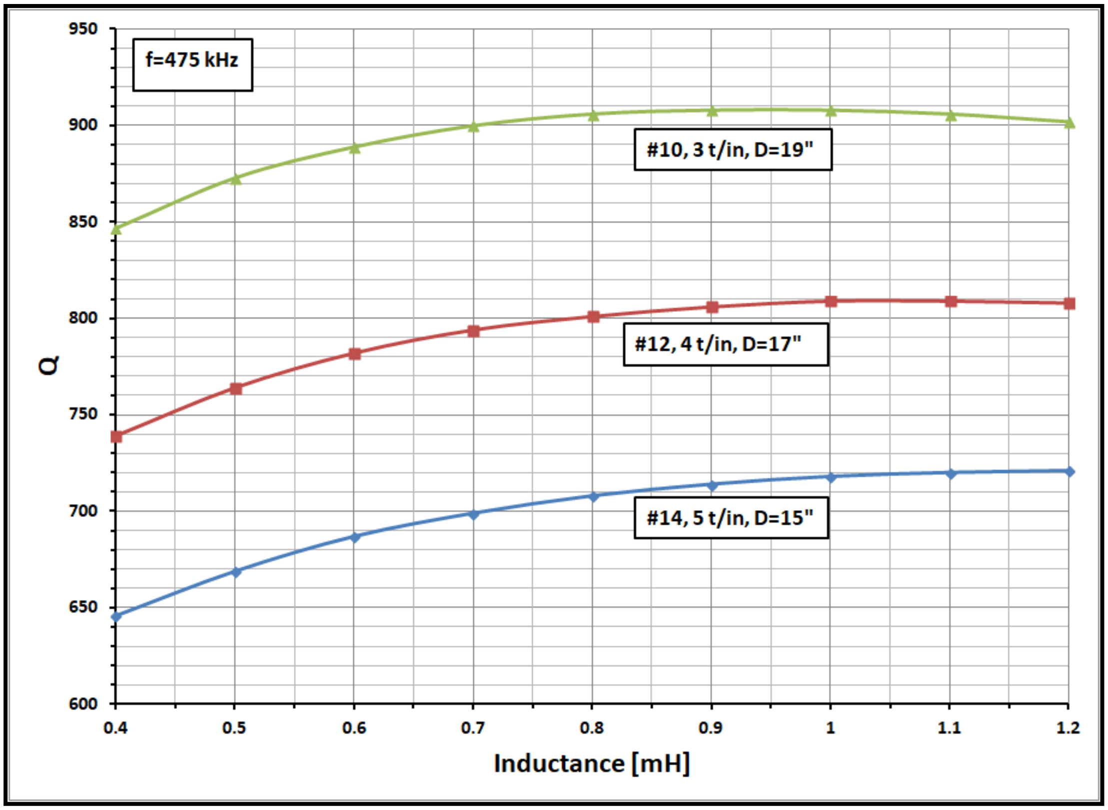

Figure 6.20B - Optimized Q at 475 kHz.

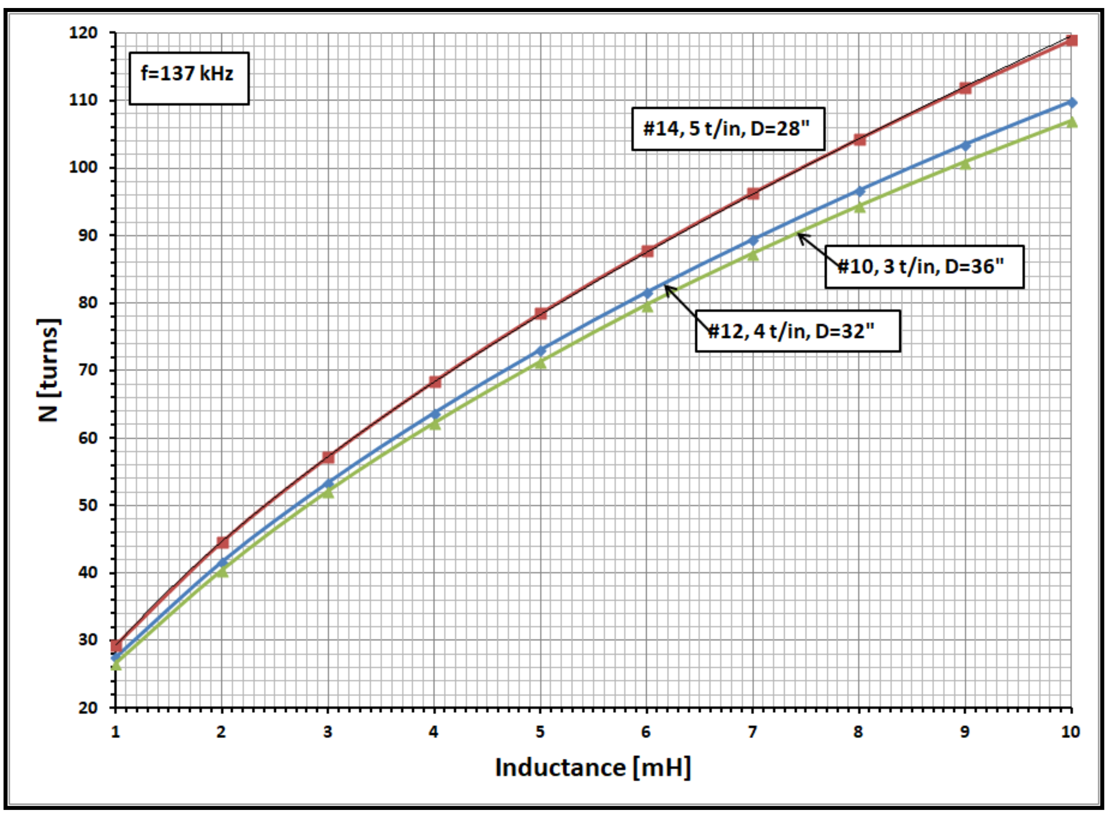

Figure 6.21A - N versus L for optimized inductors at 137 kHz.

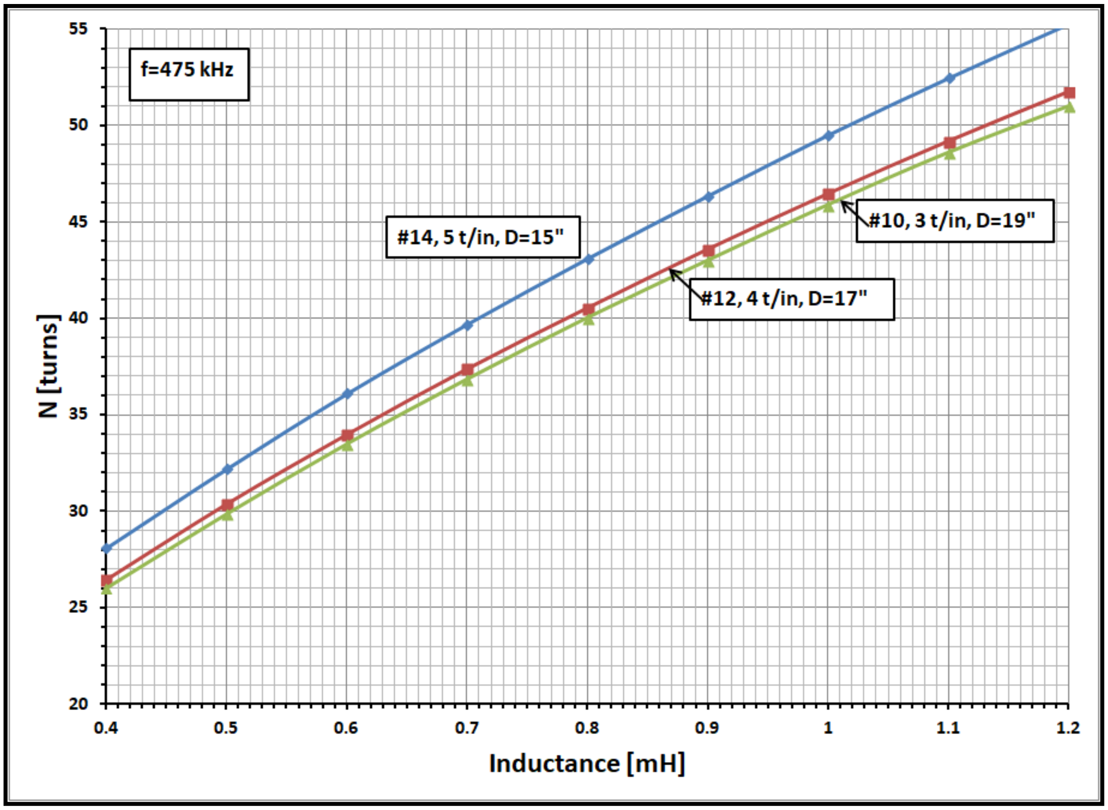

Figure 6.21B - N versus L for optimized inductors at 475 kHz.

6.3.4 Litz wire windings

To minimize skin and proximity effects might appear all we have to do is use very small wire. The New England Wire Technologies catalog suggests #40 for 137 kHz and #44 for 475 kHz. The problem with single wires this small is high Rdc. The solution is to use many small wires in parallel but simply paralleling wires in a bundle doesn't buy anything because the current still flows on the outside of the bundle. In fact ordinary stranded wire has slightly greater loss than solid wire at RF frequencies. But if we use individually insulated wires and twist the bundle during assembly in such a way that every wire is periodically transposed from the outside to the inside and then back, the current distribution can be much more uniform across the wire bundle and RL significantly lower. This type of construction is known as "litz wire". The formal name is "litzendraht", which comes from German, litzen→strands and draht→wire, "stranded wire". The strands in this wire are individually insulated and twisted to provide the required transposition. Figure 6.22 (taken from the New England catalog) shows how the strands are assembled. Initially seven strands are twisted together. To make the wire bundle larger (Rdc smaller) multiple bundles are twisted together. This process can be extended to have an Rdc equivalent to a given solid wire as shown in table 4.

Table 4 - Litz wire examples.

frequency	equiv. AWG	Cir. mil area	no. strands	strand AWG	nom. O.D.	Rdc Ω/1000'
137 kHz	12	6,727	700	40	0.118	1.76
475 kHz	12	6,600	1650	44	0.117	1.91

The advantage of litz is that it can substantially increase Q at LF and MF when used in place of solid wire. It is also very soft and pleasant to work with. But there are downsides! The cost is much higher than equivalent solid wire and there is the problem of reliably soldering 1600+ individually insulated wires to make connections at the wire ends. Soldering can be done but requires a careful choice of wire insulation and technique. Those interested in using litz should go to the wire manufactures catalogs and applications notes.

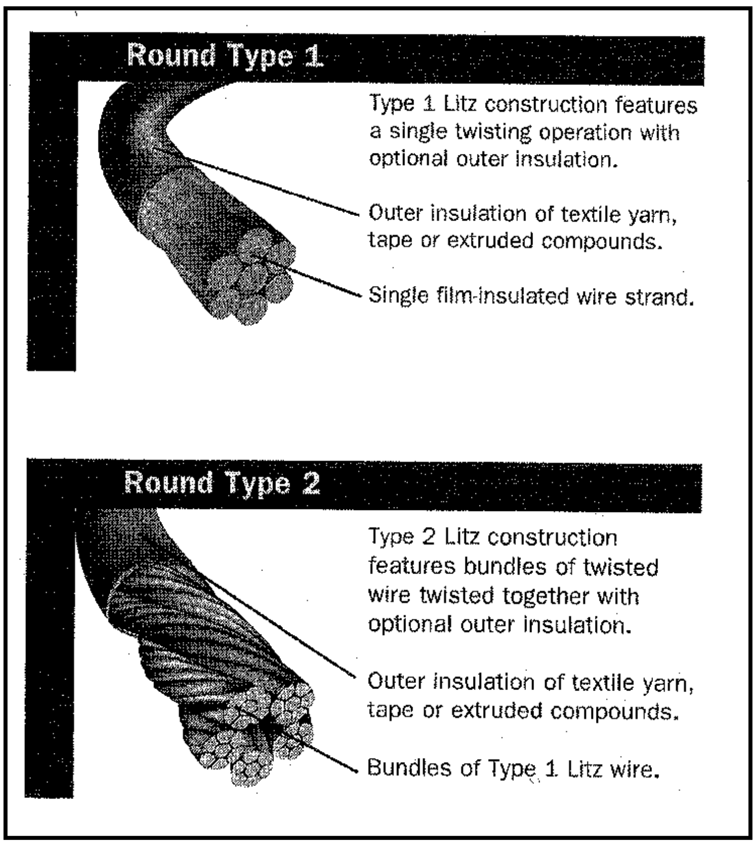

Figure 6.22 - Examples of litz construction.

Litz wire can be useful but we cannot use just any litz. Michael Perry[2] has published a formal analysis of litz wire construction which contains a cautionary tale as shown in figure 6.23!

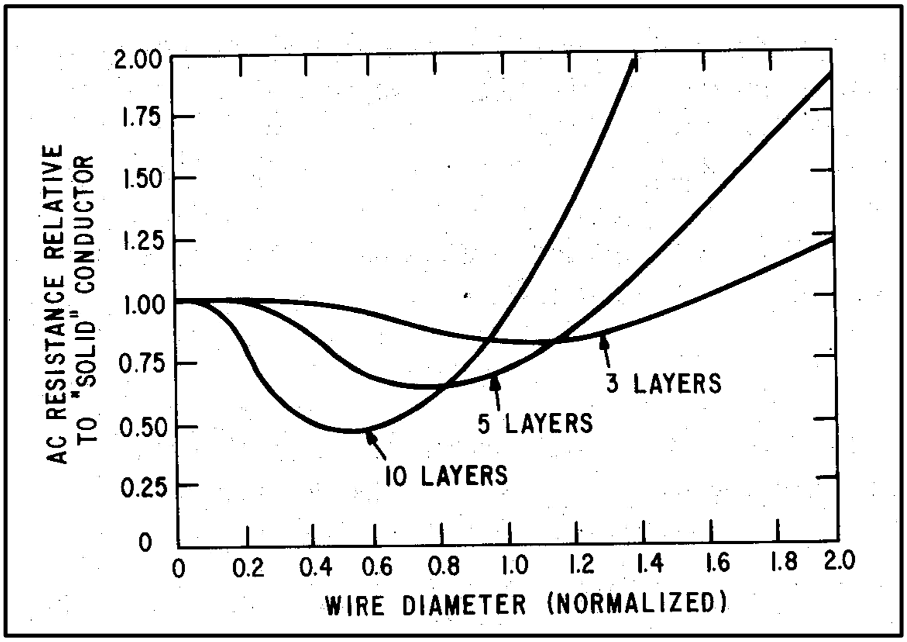

Figure 6.23 - Rac comparison between solid and litz wire. From Perry[7].

The wire diameter is in skin depths. Rac between a solid conductor and a litz conductor are compared. Litz can have a small number of wires and only a few layers or many wires forming many layers. In general the more layers and the smaller the individual strands the greater the improvement. However, there is a trap here! The reduction in Rac occurs only over a small range of wire sizes at a given frequency or, for a given wire size, over only a narrow range of frequencies. The key point shown in this graph is:

> ***If the individual wire size is too large or equivalently if the frequency is higher than the minimum Rac point, <u>Rac can be much higher using litz than in an equivalent solid wire!</u>***

The following quotation from Perry should be taken to heart if you are considering litz wire:

> ***"The foregoing analysis indicates some surprising design results which may directly contradict widely held beliefs regarding ac resistance in wires and cables. For example, suppose a solid conductor is excited at a certain frequency which results in a radius which is many times the skin-depth. Then, assume a designer switches to a cable of the same total diameter but***

with several layers of stranded wire to reduce losses such that d/δ>2. By inspection of Fig. 6.23, this process can result in far greater losses than if the "solid" conductor were employed. Stated another way, an uninformed design of Litz wire can result in a performance characteristic which is much worse than if nothing at all were done to reduce losses!

A second and important fact is that the cross-sectional area of a cable comprising stranded wires is substantially reduced from the conducting area of a "solid" cable of the same radius. This is due to the fact that each strand is usually round and insulated with a varnish or other nonconductor. The round insulated wires in the cable yield a "packing factor" which reduces the conducting area by a significant fraction, usually at least 40 percent. The transposition process further reduces the cross-sectional area available for carrying current. The final result is a substantial reduction the net savings available in ac resistance by utilizing Litz wire. Due to these limitations, the Litz wire principle for reducing ac losses must be thoroughly understood in the context of a specific application before it should be employed."

6.4 Variable inductors

Often the exact value of L needed to resonate the antenna may not be known in advance! If the antenna has already been built and an accurate measurement of the input impedance is available, L will known but the necessary instrument may not be available or the antenna may not yet have been built! With careful modeling we can get a good estimate of the value for L within ≈5-10% depending on how close the model is to the actual antenna. Even if we measure the input impedance with a VNA that measurement is only at one particular time! The short heavily loaded verticals used at LFMF have high Q's, i.e. very narrow bandwidths and are very prone to detuning, particularly as the seasons change from dry to wet. The shunt capacitance of the antenna will change with soil conductivity which changes with moisture content. Frost or water droplets on the wire will also detune the antenna. To accommodate this change in shunt capacitance some adjustment of L is almost always needed. How much adjustment is needed? Referring to figure 6.1, the antenna and loading coil form a simple series resonant circuit where the resonant frequency (fo) can be expressed by:

$$\mathbf{fo} = \frac{\mathbf{1}}{\mathbf{2\pi\sqrt{L \cdot Ca}}} = \frac{\mathbf{1}}{\mathbf{\sqrt{XL \cdot Xa}}} \quad [\mu H]\ (6.7)$$

At the least you will want to be able to tune across a band. On 630m the band is 7 kHz wide or ≈1.5%. On 2200m the band is 2.1 kHz wide, also ≈1.5%. fr varies as the square root of XL so a range of ≈3% minimum is needed. Again, as a practical matter, you should be able to vary the value of L over a range of at least 5% to 10%, with a resolution or steps smaller than 1%.

6.4.1 Tapped inductors

One of the simplest ways to vary L is with taps as illustrated in figure 6.24. The idea is to have a few widely spaced taps for coarse adjustment and then a group of closely spaced taps for finer resolution. The initial coarse adjustment is made by tapping down from the top of the coil and the fine adjustment by tapping up from the bottom (referring to the picture). However, installing a lot of taps can be a nuisance. An alternative is to put only a few taps on the main loading inductor and add a small roller inductor, like that shown in figure 6.25, in series for fine adjustment. This is a particularly convenient arrangement when doing final adjustments or making adjustments to compensate for seasonal changes. While roller inductors are often seen at ham flea markets usually the inductance is <100 uH which is usually too small for all the variation needed. The best option is usually a roller inductor for trimming in series with a larger tapped high-Q inductor providing the bulk of L.

Figure 6.24 - Large commercial inductor.

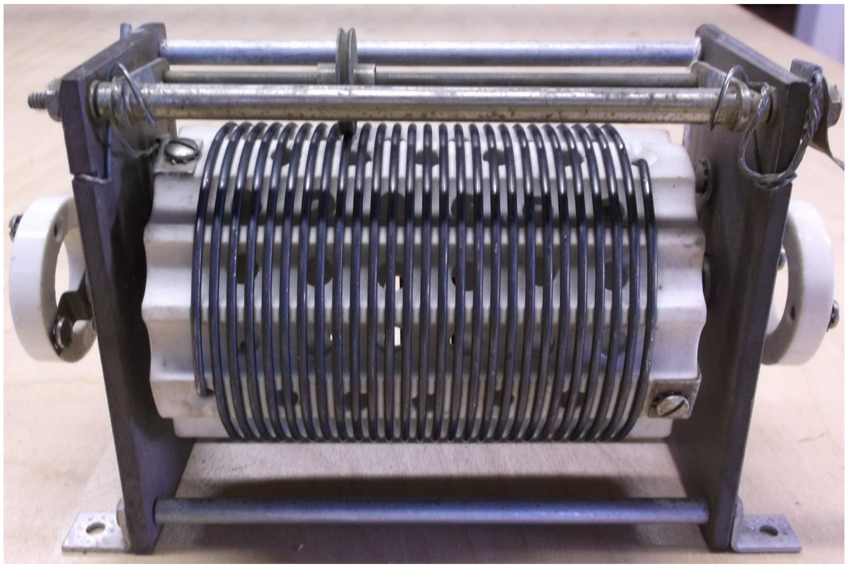

Figure 6.25 - Roller inductor example.

Tap placement

Locating taps requires some thought. When l and D are constant, L is proportional to N^2. However, when you are adding/removing turns or moving between taps the rate of change of L will vary between because you're changing both N and l. For small N the rate of change of L is close to $\propto N^2$ but as more turns are added the rate of change decreases approaching $\propto N$. Keep this in mind when selecting tap locations.

One additional point when using taps, SRF does not change greatly when moving to a lower tap. An analog for that situation is tapping down on a transmission line as shown in figure 6.26. SRF is still determined by the total length of the transmission line. In practice moving to a tap will shift the location of the parasitic capacitance (Cp) and this will shift SRF but usually not a lot.

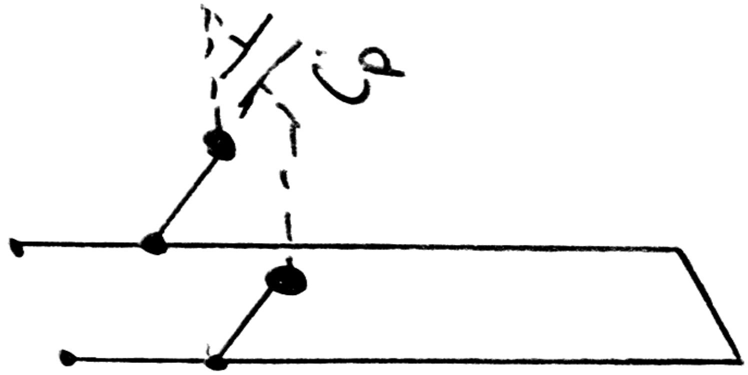

Figure 6.26 - Tapped transmission line model.

6.4.2 The variometer

Another option is to use a "variometer", which mechanically varies the coupling between two windings in series. Early radio books show an astonishing range of mechanical arrangements well worth reviewing for useful ideas. One of the most common arrangements is shown in figure 6.27 where a small secondary coil is inserted inside a primary coil, connected in series with the primary and rotated to change the coupling. The outer or primary coil has length = l_1, radius = r_1 and N_1 turns. The inner or secondary coil has length = l_2, radius = r_2 and N_2 turns. The angle between the coil axis is θ°.

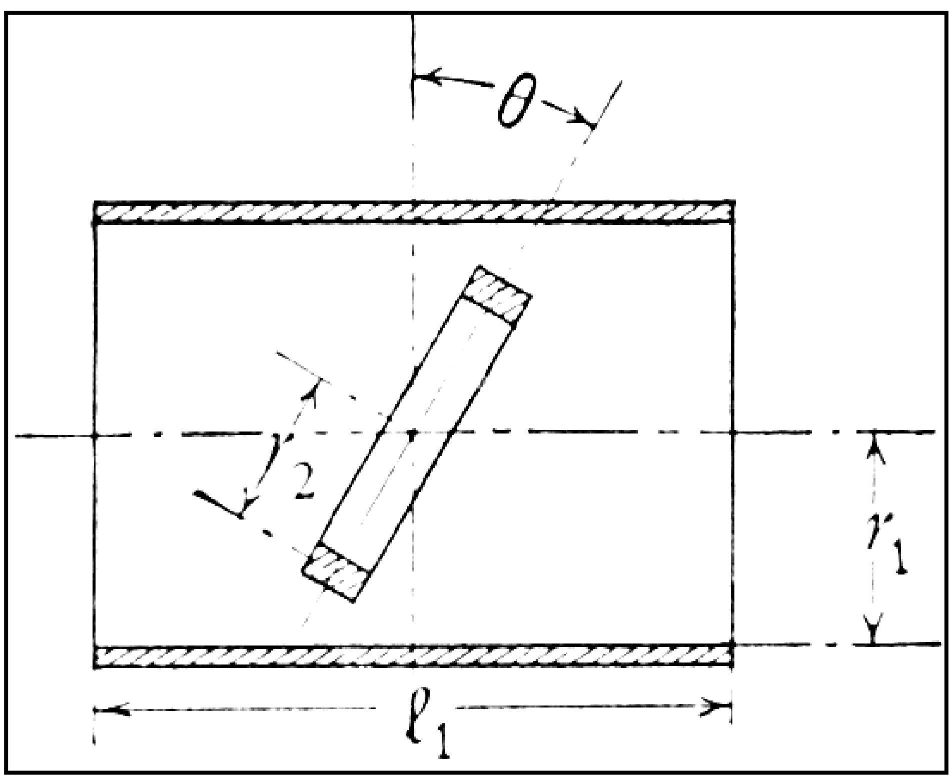

Figure 6.27 - Variometer principle.

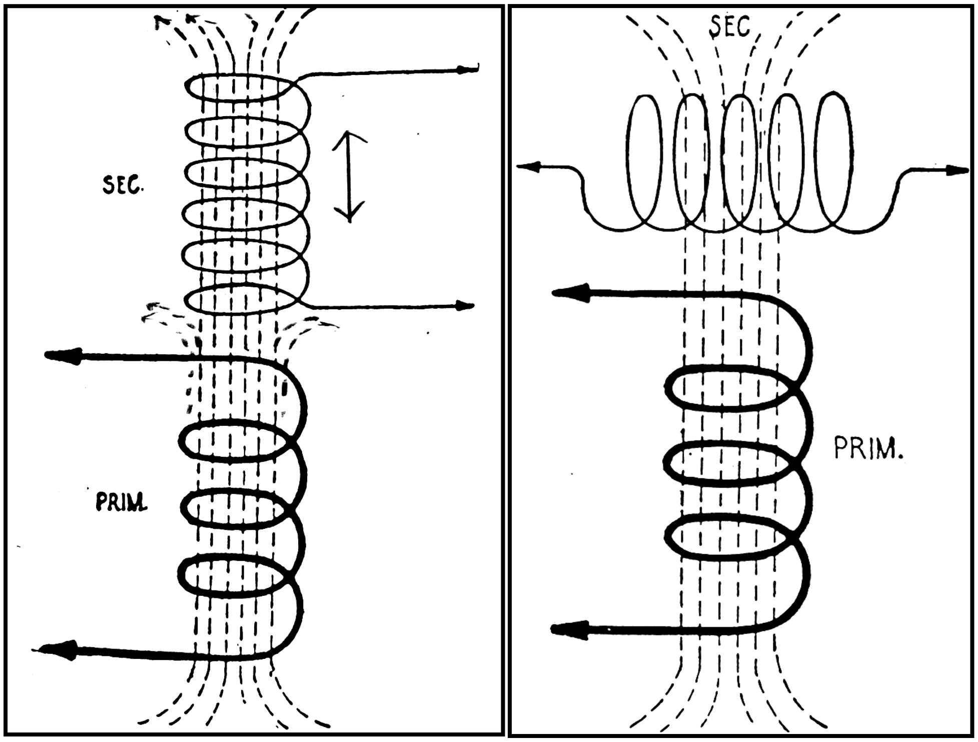

Figure 6.28 - Approximate coil flux.

Figure 6.28 is a sketch of the magnetic field associated with two coils. On the left the axis of both coils is collinear. On the right the axis' are at 90º. When the axis is parallel most, <u>but not all</u>, of the magnetic flux is the primary coil also passes through the secondary coil.

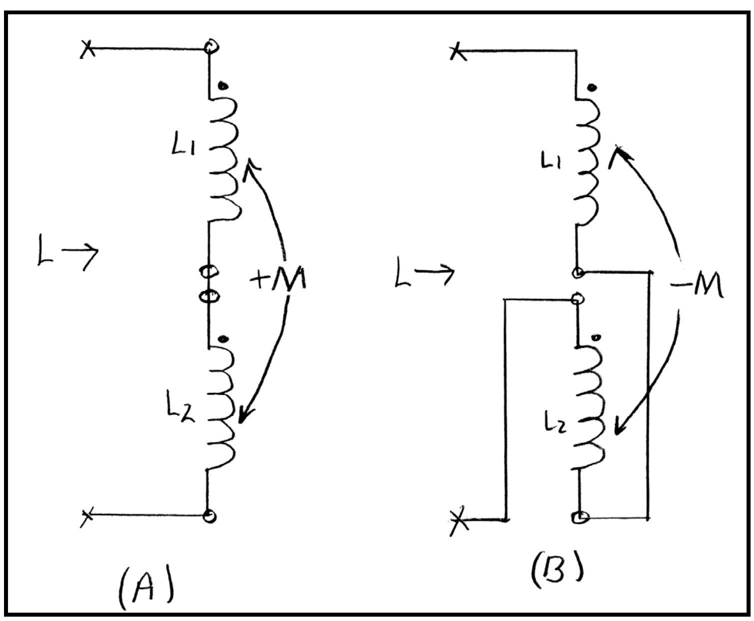

Figure 6.29 - Series aiding versus series opposing.

How do we get a variable inductance by varying the coupling? Figure 6.29 shows two coils (L1, L2) connected in series. In (A) L1 and L2 are connected series aiding and in (B) series apposing. The value for L can be expressed as:

$$\mathbf{L = L1 + L2 \pm 2M} \quad (6.8)$$

Where M is the "mutual" inductance:

Note that M can be either + or -. +M corresponds to series aiding connection and -M to series apposing which can be adjusted by rotating the secondary coil. M will vary approximately as the cos(θ).

We can calculate M from:

$$\mathbf{M = \frac{0.4 N_1 N_2 \pi^2 r_2^2}{l_1 + r_1}} \quad \mathbf{uH} \quad (6.9)$$

<u>Where r and l are in meters</u>.

6.4.2.1 Bucket variometers

We can gain good understanding of variometers by designing and testing an example shown in this section. A frequent form of variometer among amateurs is built on a plastic bucket. An example is shown in figure 6.30

Figure 6.30 - bucket variometer example.

Figure 6.31 shows three possibilities for the inner inductor, L2. the largest form is the top of a 2 gallon bucket, r_2≈4.6"/0.117m. The smallest form is a section of 4" PVC pipe, r_2≈2.09"/0.053m The middle form is the top of a 1 gallon bucket, r_2≈3.5"/0.089m. For this discussion we'll compare two of them: PVC pipe and the 2 gallon bucket top.

How much variation in inductance (±ΔL) is wanted? L1 for this example is ≈639 uH. A typical range of variation would be ±10% or ±60 uH in this example. For this inductor N1=56 turns, l_1=11"/0.279m, r1=5.4"/0.137m.

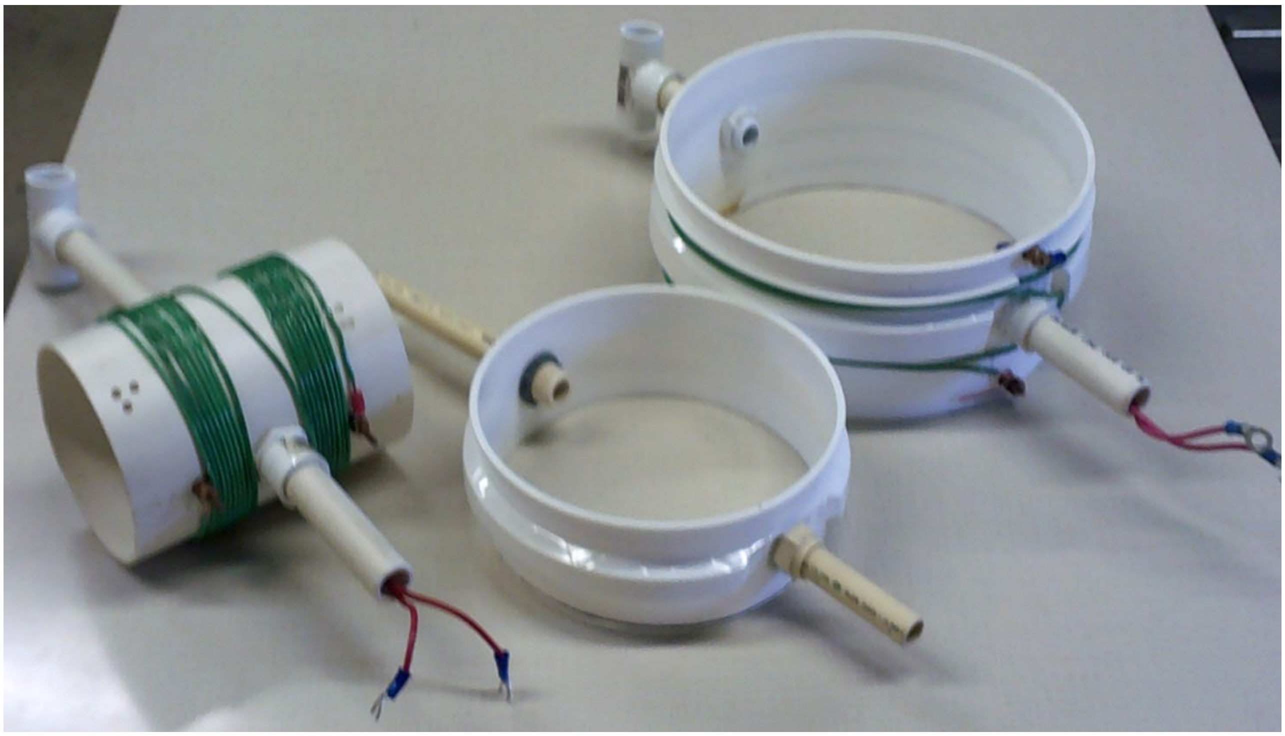

Figure 6.31 - Alternative L2 coil forms.

From equation 6.8, ΔL=2M

$$\Delta L = 2M = \frac{0.8 N_1 N_2 \pi^2 r_2^2}{l_1 + r_1} \quad \text{uH} \qquad (6.10)$$

We know the following variables:

ΔL=60 uH, N1=56, l_1=0.279m, r_1=0.137m

For the 2 gallon ring, r_2=0.117m and for the PVC pipe, r_2=0.053m.

What we don't know is N_2! Rearranging equation 6.10 solving for N_2:

$$\mathbf{N_2 = \frac{\Delta L(l_1 + r_1)}{0.8 N_1 \pi^2 r_2^2}} \qquad (6.11)$$

We can now calculate N_2 for the two examples:

- 2 gallon bucket top, N_2=4.1 turns, use 4 turns.
- PVC pipe, N_2=19.5 turns, use 19 turns.

Winding the calculated number of turns on each form and inserting them into the bucket and measuring, ΔL=61 uH for the large ring and ΔL=62 uH for the PVC pipe showing that equation 6.11 is approximate but certainly close enough for practical purposes.

In this example L≈639 uH with 56 turns. Tapping down one turn L≈614 uh which is a shift of ≈25 uH. Choosing ΔL=60 uH is comparable to moving the tap roughly two turns. You could use less ΔL which should improve Q as shown in the next section but then you would need to insert a sufficient number of taps.

6.4.2.2 Variometer Q

There is one very important consideration: how is the Q of L1 affected by inserting L2 inside it? In figure 6.27 we see that L2 is inside L1 immersed in the internal field of L1. This means that L2 will have it's normal self loss but also additional loss from the field of L1. In addition, the external field of L2 will interact with the winding for L1, more loss. There is also another reason for decreased Q.

$$\mathbf{Q = \frac{XL}{RL}} \quad (6.12)$$

When L1 and L2 are connected series aiding XL will be a maximum but when they are series apposing XL will be lower. RL however, will probably not be lower so Q must decrease as L decreases. Here are some Q measurements at 475 kHz for the variometer shown in figure 6.30:

No center coil, Q=700

For the 9" diameter 4T center coil, Q=500 for Lmax and Q=420 for Lmin.

For the 4" diameter 19t center coil, Q=550 for Lmax and Q=530 for Lmin.

The message here is that adding a variable center coil to an inductor to create a variable inductor will significantly reduce Q. I suggest the following guidelines:

Use as few turns as possible on L2, i.e. design for the minimum needed inductance variation. Use L2 for fine adjustments. Use taps on the coil for course adjustments.

From these admittedly limited experiments it appears that very large and small L2 diameters give lower Q. I suggest having the diameter of L2 equal to roughly half that of L1.

6.4.2.3 Classic examples

Figure 6.27 is just one of many arrangements as shown in figures 6.32 through 6.34. Some of these examples show flat strip windings but round wire will also work.

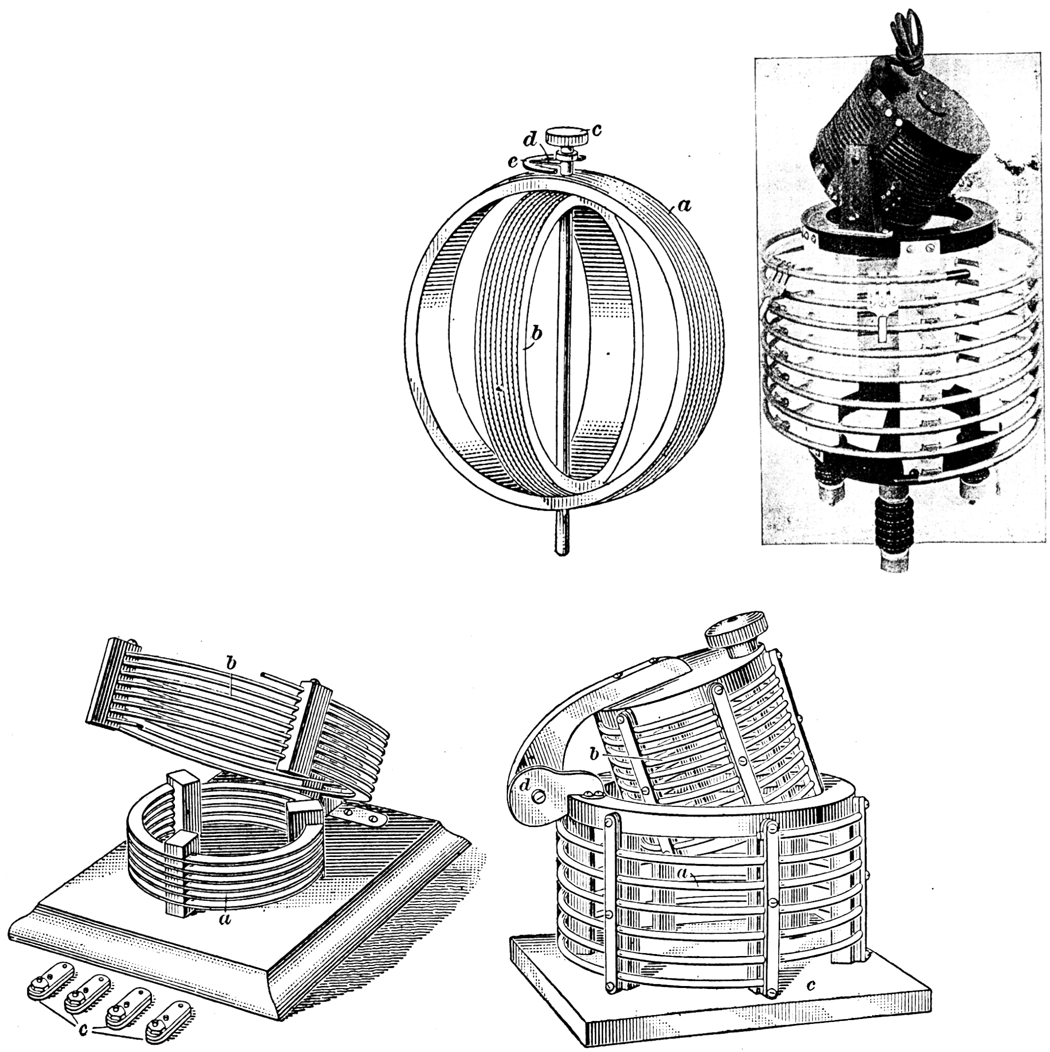

Figure 6.32 - Variometer examples

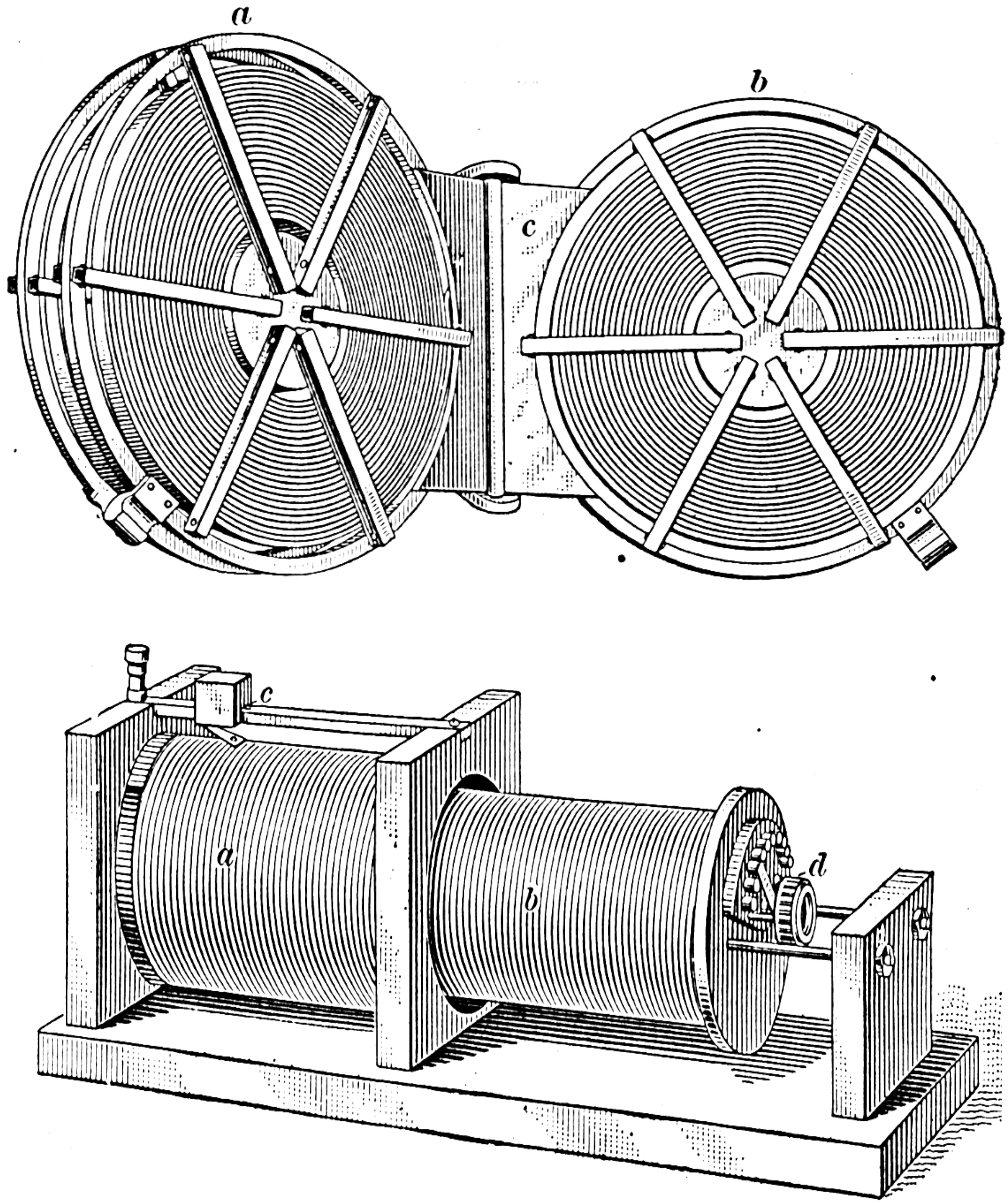

Figure 6.33- More variometer examples.

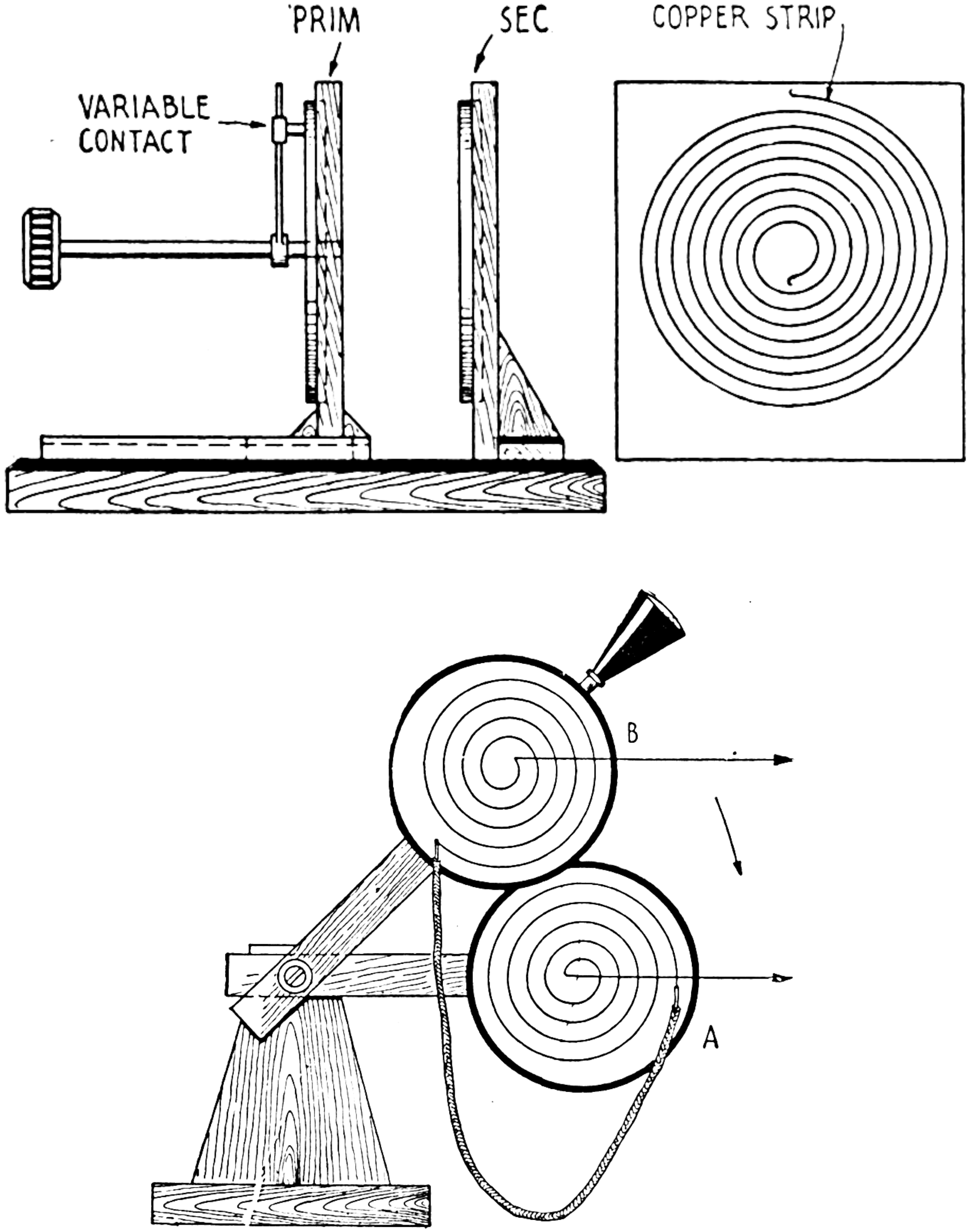

Figure 6.34 - Even more examples.

6.4.2.4 Home brew examples

Figure 6.35 - Ralph, W5JGV variometer example.

As shown in figure 6.35, Ralph, W5JGV, has built lovely variometers using PVC pipe and copper wire. The inductance of these variometers is not large but adequate for fine adjustment in combination with a tapped main inductor. The other option is to incorporate the variable inductance into the main inductor as shown in figure 6.36.

LFMF hams have shown great creativity in the design of practical variometers as the following examples show.

Figure 6.36 - Laurence, KL7L. Bucket variometer example.

Figure 6.37 was fabricated by John, KB5NJD. The base inductor is wound on the outside of a plastic bucket. Inside the bucket is a smaller rotatable inductor. The two inductors are connected in series. By rotating the inner inductor the total inductance goes from the sum of both to the difference. Given the need for adjustment at inconvenient times (pitch dark and snowing) many variometers have some form of remote tuning. KB5NJD used an inexpensive TV antenna rotor for remote tuning!

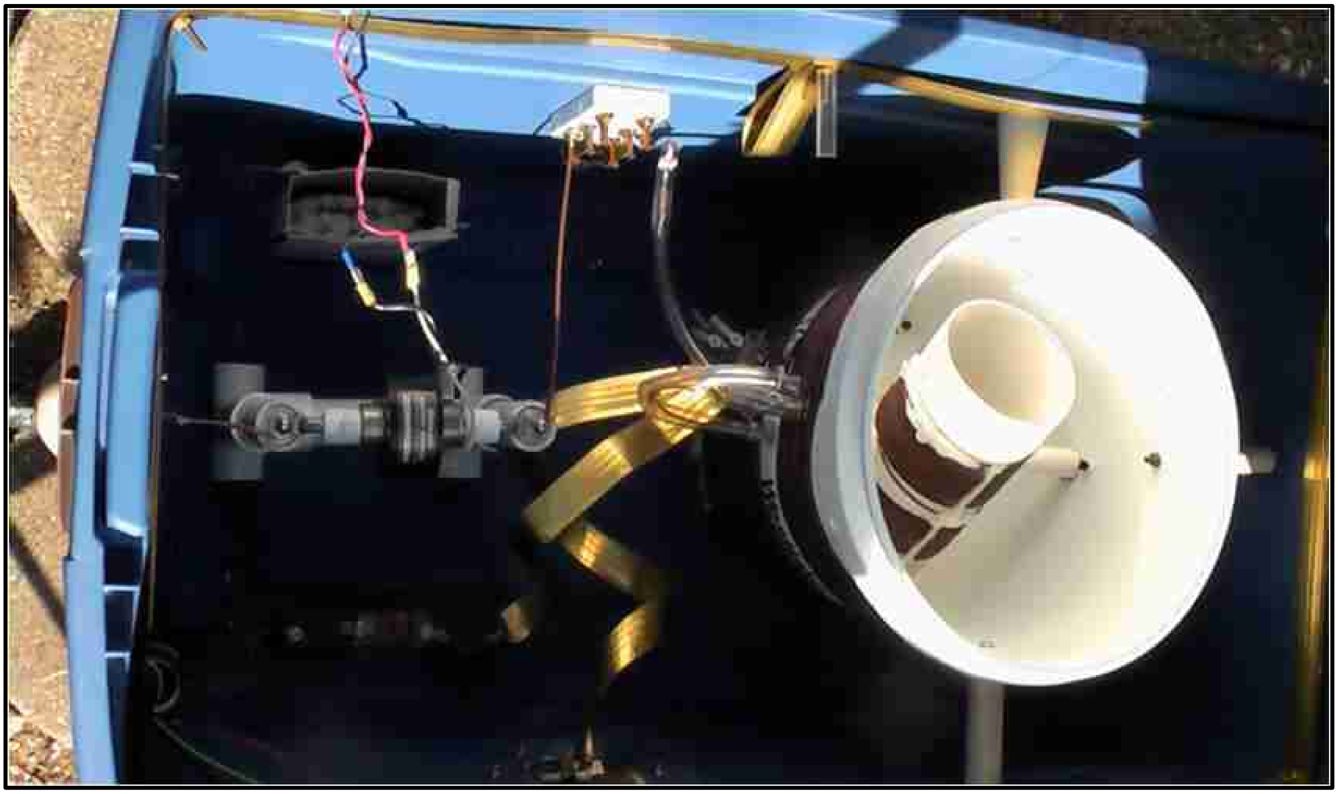

Figure 6.37A- KB5NJD variometer.

Figure 6.37.B - John, KB5NJD variometer adjustment with a TV rotor.

Figure 6.38A - Jay, W1VD, WD2XNS variometer.

Figure 6.38B - Jay, W1VD, WD2XNS variometer.

Figure 6.38C - Jay, W1VD, WD2XNS variometer enclosure.

Figure 6.39 - Steve, KK7UV, variometer.

Figure 6.40A -Neil, W0YSE, variometer.

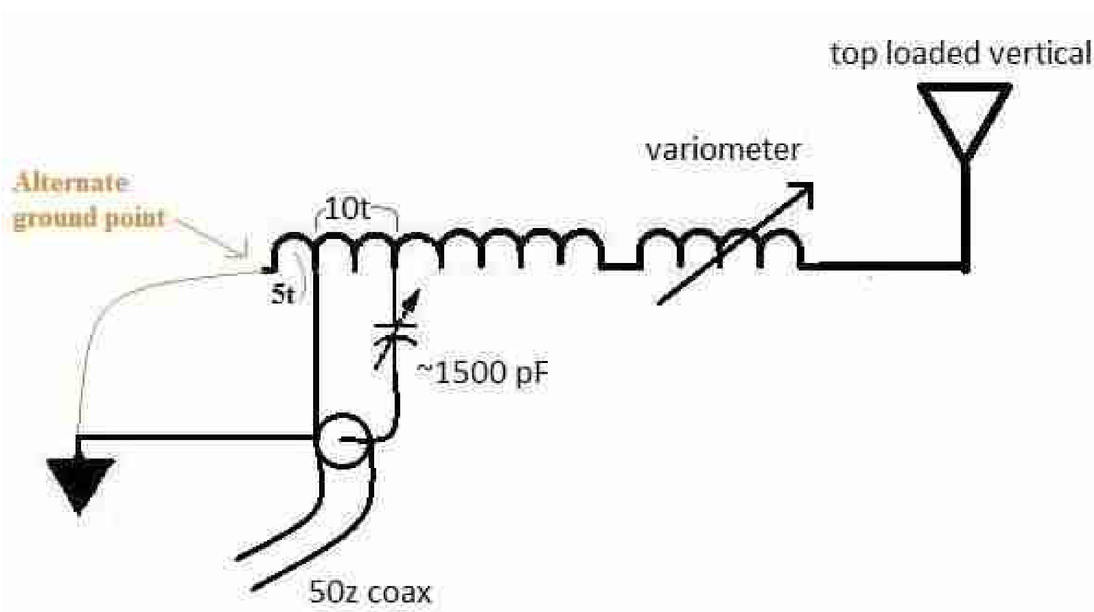

Figure 6.40B - W0YSE tuning unit circuit diagram.

Figure 6.40C - W0YSE variometer location.

Neil, W0YSE, located his variometer just outside a window of the shack. Adjustment is manual: open window, twist variometer knob, close window.

6.5 Winding voltages, currents and power dissipation

The current at the base of the antenna (Io) is also the current in the inductor. The voltage across the inductor is the same as the voltage at the base of the antenna (Vo).

Io is determined by the radiated power Pr and the radiation resistance Rr:

$$\mathbf{Io} = \sqrt{\frac{\mathbf{Pr}}{\mathbf{Rr}}} \quad (6.13)$$

As explained in chapter 1, the maximum radiated power (Pr) is limited to 1.67W on 630m and 0.33W on 2200m. Combining the band specific values for Pr with Rr values we can use equation (6.13) to create the graphs for Io shown in figures 6.41 and 6.42. Note that L is the overall length of the top-wire in feet in all of the graphs in this section.

Vo is the voltage at the feedpoint:

$$\mathbf{Vo = XiIo} \quad (6.14)$$

We can use typical Xi values from chapter 3 to generate values for Vo as shown in figures 6.43 and 6.44. Despite the low radiated powers (Pr) the voltages at the base will often be >1kV and can be much higher, particularly when H is small. This must be kept in mind when selecting a base insulator. A high Vo also means there will be significant voltage turn-to-turn in the loading inductor and across matching network components.

PL is the power dissipated in the loading inductor, PL=Io^2RL. As shown in figures 6.45 and 6.46 this power can easily be >100W (assuming that level of transmitter power is available). The loading inductor must dissipate PL without damage! In general the larger the physical size of the inductor the better heat can be dissipated. QL=300 is assumed for both 137 kHz and 475 kHz. This is a bit pessimistic given the earlier discussion since increasing QL reduces PL proportionately:

$$\mathbf{\frac{PL_2}{PL_1} = \frac{QL_1}{QL_2}} \quad (6.15)$$

In short verticals with limited top-loading Io, Vo and PL can be very high. The key to reducing these values is to use sufficient top-loading.

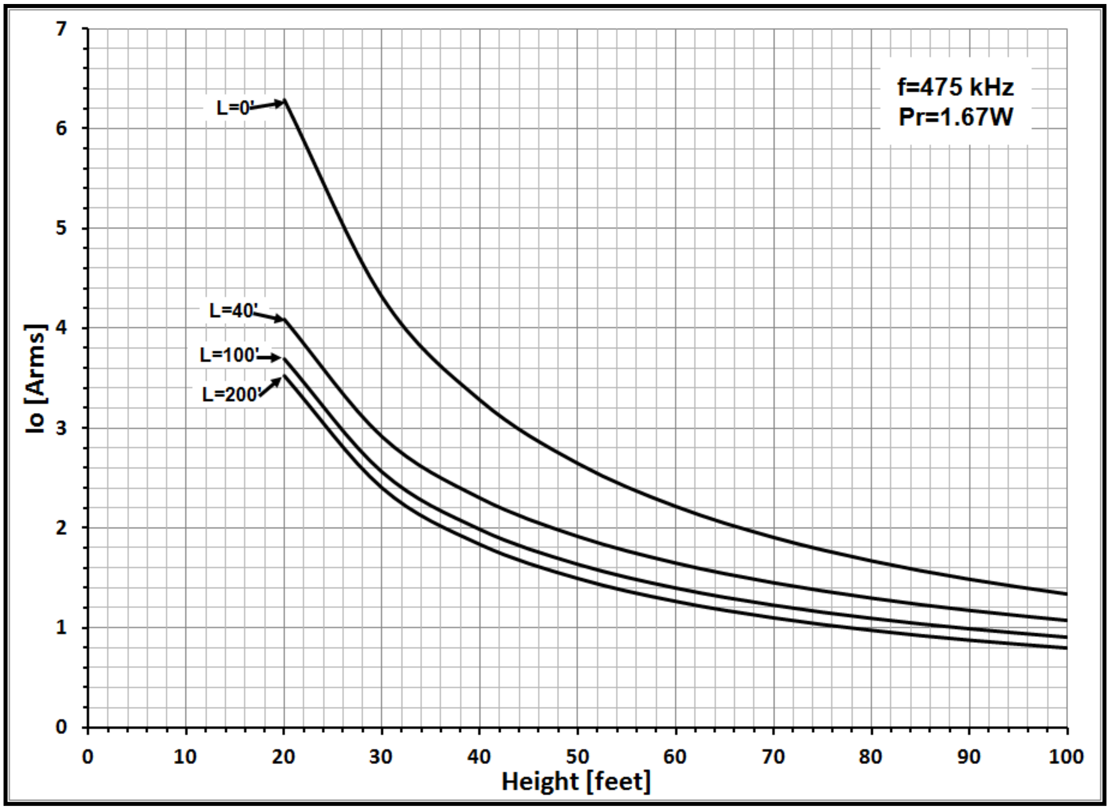

Figure 6.41 - Io for Pr=1.67W at 475 kHz.

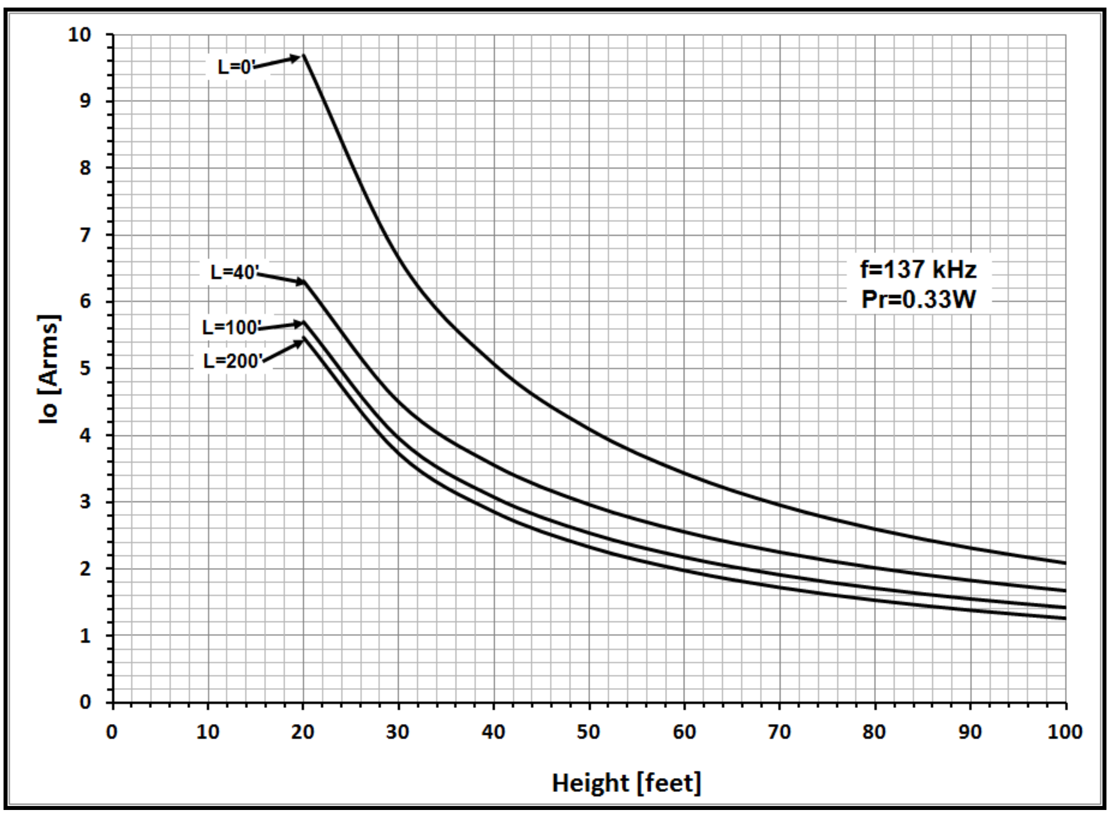

Figure 6.42 - Io for Pr=0.33W at 137 kHz.

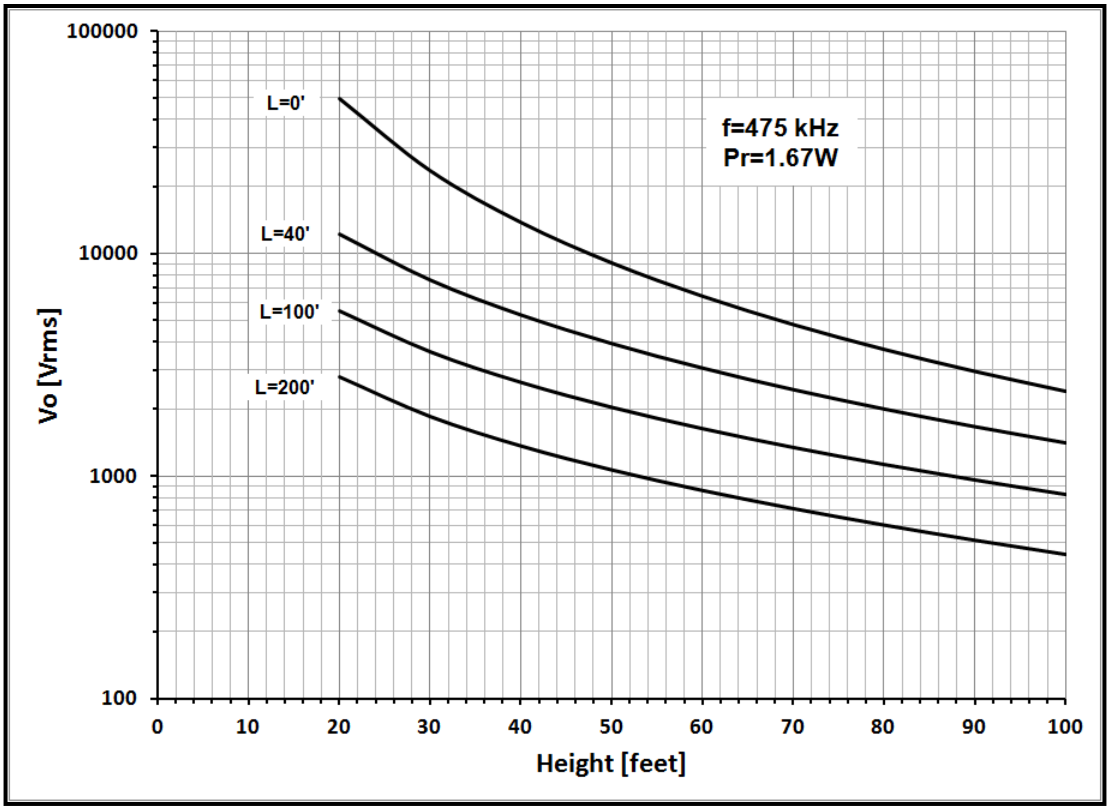

Figure 6.43 - Vo for Pr=1.67W at 475 kHz.

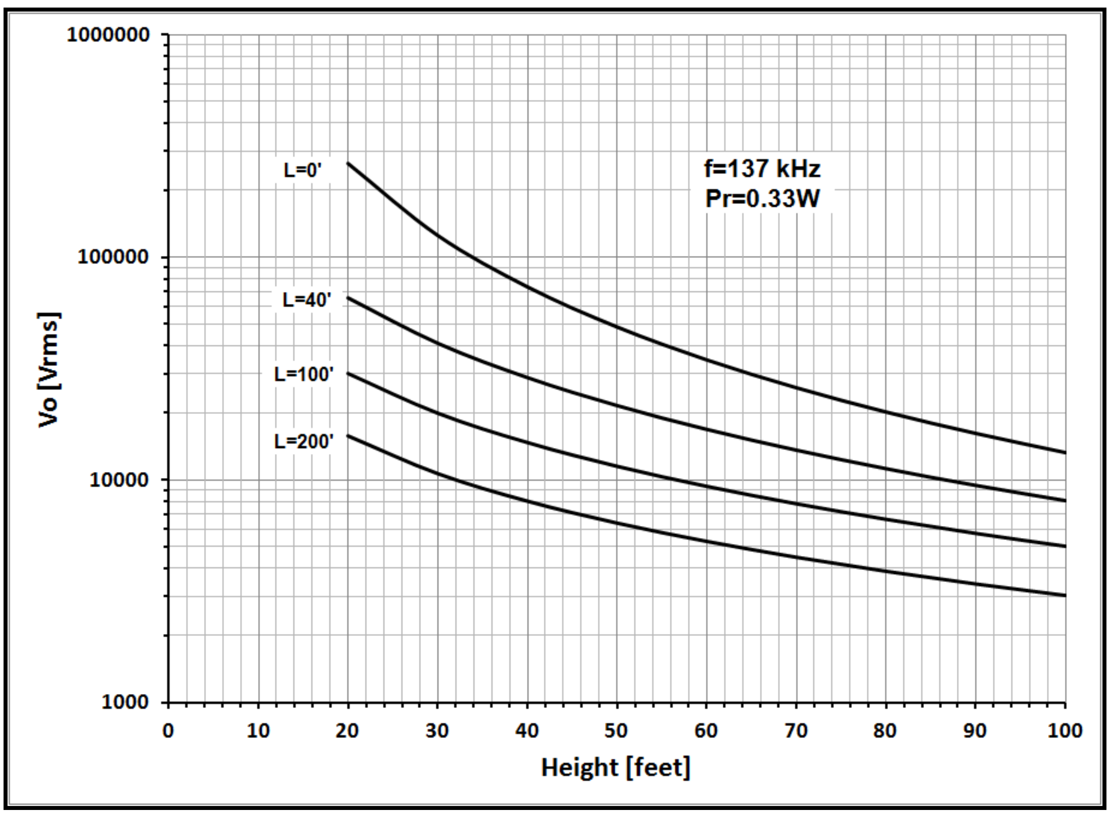

Figure 6.44 - Vo for Pr=0.33W at 137 kHz.

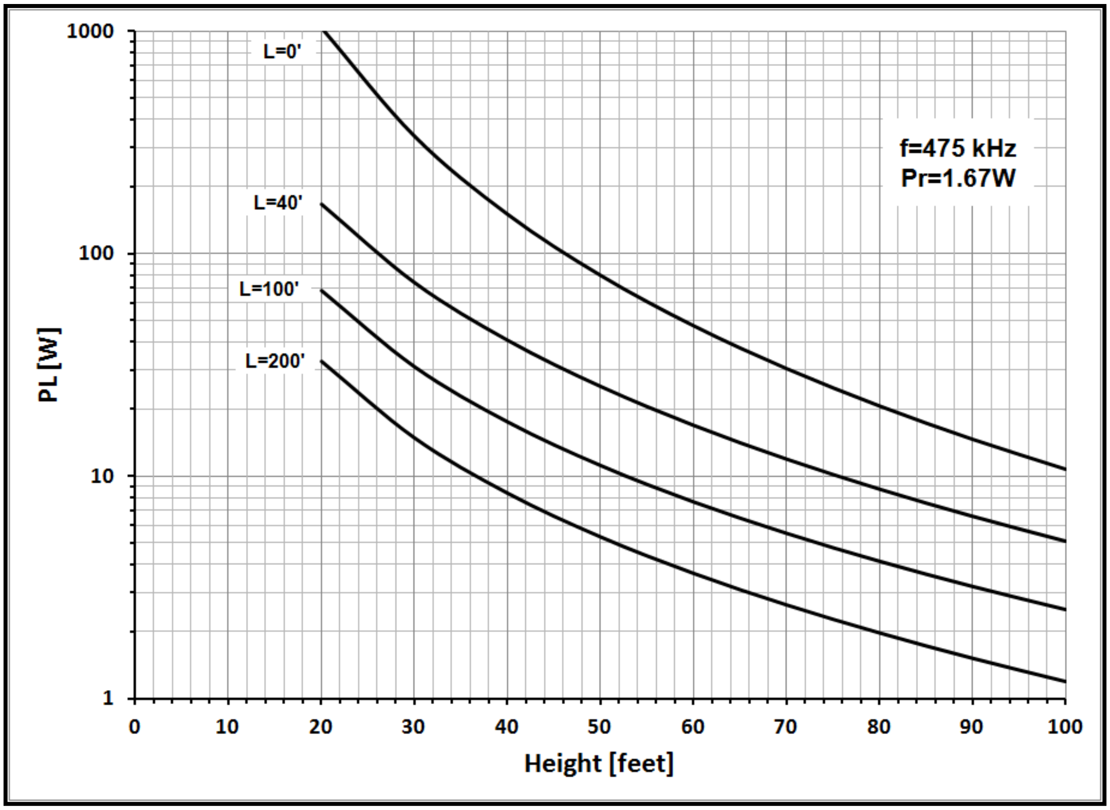

Figure 6.45 - PL for Pr=1.67W at 475 kHz.

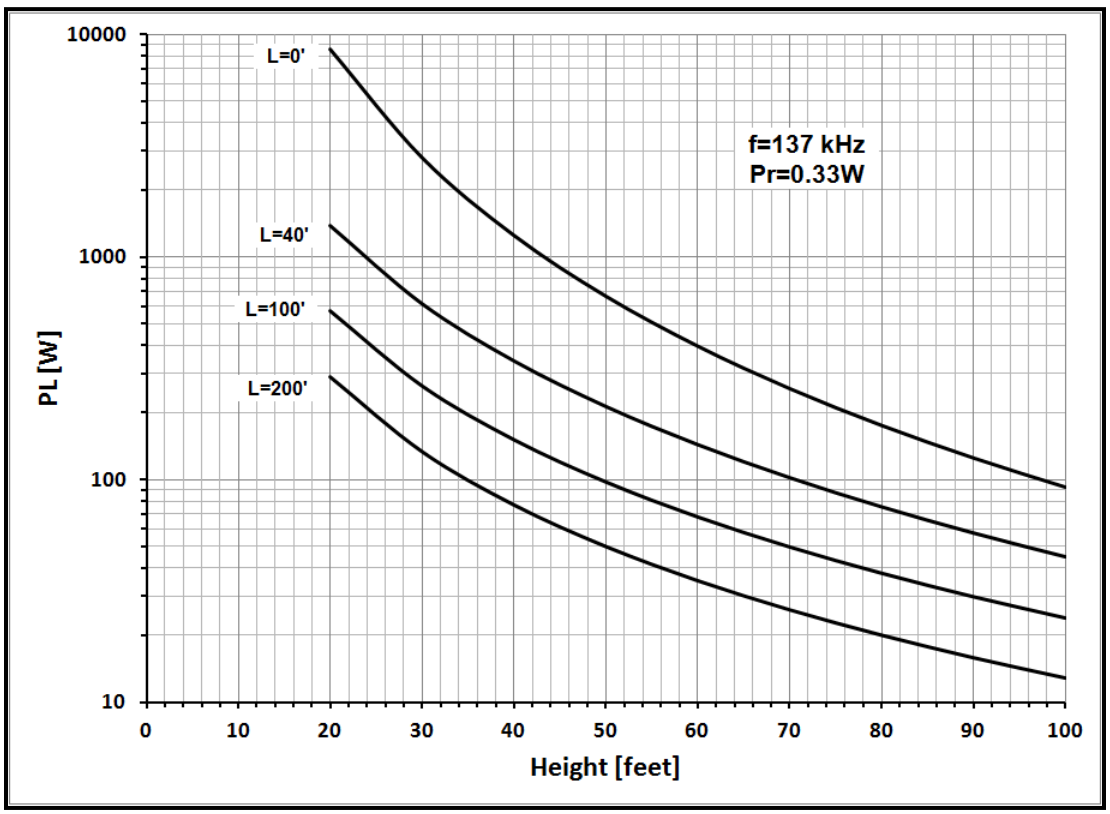

Figure 6.46 - PL for Pr=0.33W at 137 kHz.

6.6 Enclosures

Most amateurs will use some form of large plastic box for the tuning inductor enclosure. These are inexpensive, readily available in a very wide range of sizes and have little or no effect on QL. One shortcoming of typical plastic containers is their susceptibility to degradation from the UV in sunlight. A coat of white house paint is usually enough to allow them to last several years. Metal enclosures can also be used although large enclosures will usually be custom fabricated and are expensive. In general a metal enclosure needs to be substantially larger than the inductor. In particular the spacing from the ends of the coil to the enclosure wall should be at least equal to the coil diameter. A conducting enclosure will tend to reduce both L and QL if it isn't large enough.

The primary purpose of the enclosure is to protect the coil from the weather, in particular keep it dry. As the following experimental work shows, moisture can severely degrade inductor Q. Some form of enclosure usually covers the coils but they can still be quite damp and perhaps even have some icing. K6STI sent me a link (http://www.n3ox.net/tech/coilQ/) to a note reporting some experiments on wet inductors. The news was not good, moisture does not benefit Q! After some discussion I ran a few simple experiments on two large coils using an HP4342A Q-meter to judge the effects of water on Q. All the Q measurements were made at 475 kHz. In each test the Q-meter Cr was adjusted to re-resonate by adjusting for peak Q on the meter. Most of the experiments were repeated to check consistency.

6.6.1 Experiment 1

I had on hand the bucket inductor shown in figure 6.47. L≈650uH. I began with the coil dry: Q=460. Using a spray bottle, I sprayed water over the outside of the winding to simulate rain: Q=200! The coil was not happy being wet!

Figure 6.47 - Bucket inductor.

6.6.2 Experiment 2

The next test used the bare #12 wire coil on a PVC cage shown in figure 6.48. L≈1 mH @475 kHz.

Figure 6.48 - PVC cage inductor.

Note: in the photo there is a plastic bag of ice lying in the bottom of the coil. The ice was placed there later in the experiment. The initial test was a dry coil, Q≈700 which was then sprayed: Q≈400. Wiping the coil with a towel, the Q increased to 500 and slowly rose as the coil dried returning to ≈700 after some hours. This coil was very sensitive to even a small film of moisture.

The water used for the initial test was from my well which has a very small amount of salt in it. As a check I bought a gallon of distilled water, rinsed the spray bottle carefully, and when the coil was again dry, re-sprayed it with distilled water: Q≈530. Then I switched back to well water and re-sprayed: again Q≈400. One might argue that rain water is closer to distilled water than my well water but any inductor outside will have deposits from the air and rain water also brings down local pollution so I don't think the improvement using distilled water is cause for joy. Up to this point the shift in Q-meter Cr was very small, a pF or so, essentially all the variation in Q was due to additional loss not a shift in SRF.

6.6.3 Experiment 3

Brian had suggested that ice might have higher losses than water so I ran a test. As shown in figure 6.48 I placed a bag of ice inside the coil: Q≈250, not good! To see if ice had more loss than water I allowed the ice to thaw completely and then put the now water-bag back in the coil: Q<200. That looks like the ice is less lossy than water but I think that's deceiving. The ice was pretty lumpy and the contact with the winding intermittent but when thawed the bag of water lay much closer to the winding. In practice I think the ice and water have the same effect. Outside in icing conditions the ice might very well build up a much thicker layer on the winding than a thin film of water so the effect might be much greater.

6.6.4 Experiment 4

I wanted to see if water on the outside of an enclosure would have any effect so I placed a plastic bag over the cage inductor as shown in figure 6.49. Note, the bag is quite close to the coil. I sprayed the outside of the bag with the coil connected to the Q-meter: no observable effect. Now if the bag had heavy icing then maybe it might make some difference but I was not able to run that experiment.

6.6.5 Experiment 5

During all the earlier experiments the shift in Cr was quite small, a pF or so, even when Q was severely reduced. Just for the heck of it I placed the one gallon jug of distilled inside the coil: Q dropped from 700 to 360 and Cr was reduced by 5 pF. This was the only time I saw a significant shift in Cr, which would be a dielectric effect.

Figure 6.49 - Coil with plastic cover.

6.6.6 Conclusions

This series of experiments was hardly rigorous, but I think they convincingly give the message: keep your coils dry and out of the weather and condensation. To fight condensation the enclosure needs to be, drained and ventilated but be careful, don't make the vent holes large enough for the bees and wasps to get in. A hornet nest in the coil does not improve Q! The rule of thumb suggesting the walls of any enclosure, soil or other objects be at least one coil diameter away from the coil seems like reasonable advice.

References

[1] Brian Beezley, K6STI, COIL.exe, http://ham-radio.com/k6sti/

[2] Michael Perry, Low Frequency Electromagnetic Design, Marcel Dekker, Inc. 1985, pp. 107-116

[3] Rudy Severns, N6LF, Conductors for HF Antennas, Nov/Dec 2000 QEX, pp.20-29

[4] Rudy Severns, N6LF, Insulated Wire and Antennas, QEX March/April 2018, pp. 22-28

Appendix A4

Rg and Ground Systems

A4.0 A closer look at ground systems

Chapter 5 provided a number of practical examples of ground systems. For the most part the performance of these systems was derived from NEC modeling with very little math. For many readers that's more than sufficient but some will want more information. This appendix gives some additional information that should provide some insight.

A4.1 Feedpoint equivalent circuit

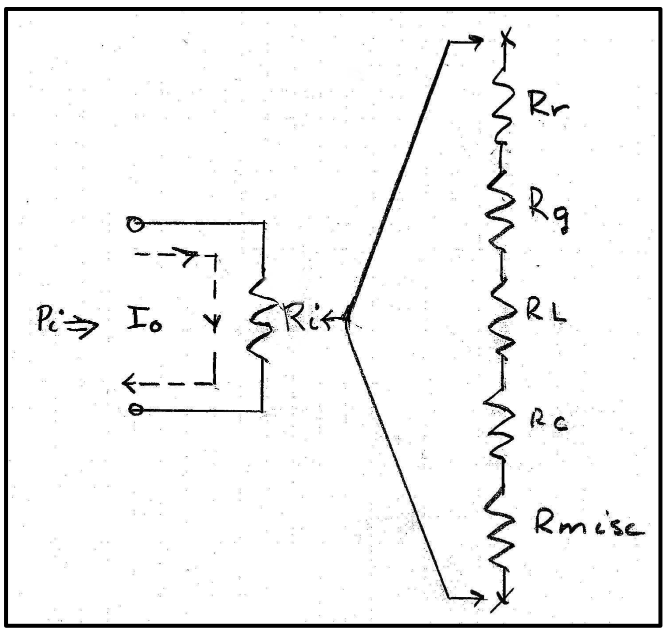

Figure A4.1 - Equivalent circuit for the resistive part of the feedpoint impedance.

Figure A4.1 shows an equivalent circuit used to represent the resistive part of an antenna's feedpoint impedance (Ri). Io is the current at the feedpoint and the input power $Pi=Io^2Ri$.

- Rr is the radiation resistance representing the radiated power (Pr).
- Rg accounts for the power dissipated in the soil (Pg).
- RL represents the tuning inductor loss (PL).
- Rc represents the loss in the conductors (Pc).
- Rmisc represents other losses such as insulator leakage, etc.

This discussion is focused on Rg and it's relation to Rr. The other losses are important but have been addressed elsewhere.

Rr is very dependent on the specific details of the antenna: i.e. dimensions and loading. Rr can also be a function of soil electrical characteristics and ground system design. Although this effect is prominent at HF it's significantly less at LFMF.

Rg depends on soil electrical characteristics which vary with frequency, details of the ground system and <u>the antenna associated with the ground system</u>. If we modify the antenna, even without changing the ground system or soil, Rg will change. We have to remember that neither Rr nor Rg is a physical resistor, they are "accounting tools" we use to keep track of where the input power (Pi) is going. Because Pg depends on the electric and magnetic field intensities at the ground surface which change when the antenna is changed, Rg is dependent on the details of the antenna as well as the ground system itself.

A4.2 Definitions for Pr, Pg, Rr and Rg

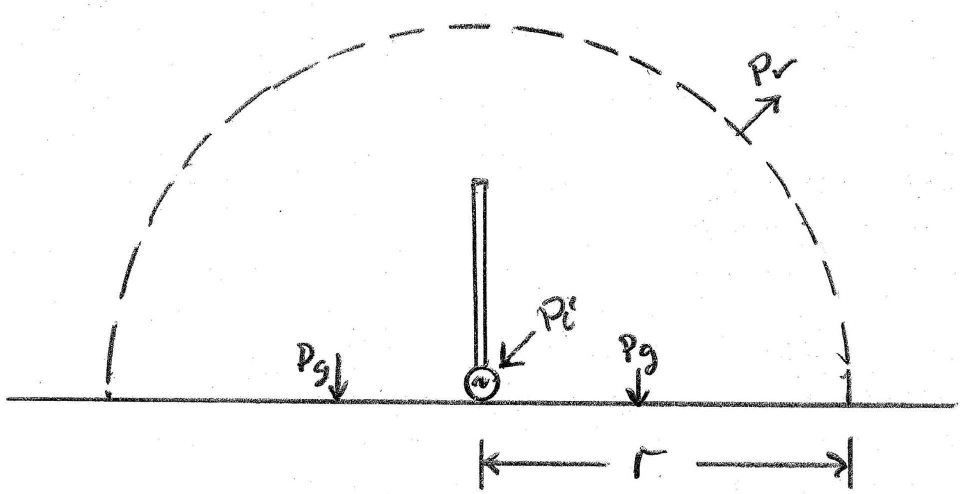

Figure A4.2 - Pr and Pg.

Figure A4.2 illustrates "Pr" and "Pg". The dashed line represents a hypothetical hemispherical surface enclosing a vertical antenna. The hemisphere radius r is usually set r=λ/2 because that is approximately the outer boundary of the reactive near-field for verticals with a height of λ/8-λ/4. Pr is defined as the total power radiated through the hemisphere. Pg is defined as the power flowing into the ground surface and dissipated in the soil.

Rr and Rg are defined in terms of Pr and Pg:

$$\boldsymbol{Rr} \equiv \frac{\boldsymbol{P_r}}{\boldsymbol{I_o^2}}\ \Omega \quad \text{(A4.1)} \qquad \boldsymbol{Rg} \equiv \frac{\boldsymbol{P_g}}{\boldsymbol{I_o^2}}\ \Omega \quad \text{(A4.2).}$$

A4.3 Efficiency

We can state efficiency **η** as:

$$\boldsymbol{\eta} = \frac{\mathbf{1}}{\mathbf{1}+\frac{\mathbf{R_g}}{\mathbf{R_r}}+\frac{\mathbf{RL}}{\mathbf{R_r}}+\frac{\mathbf{R_c}}{\mathbf{R_r}}+\frac{\mathbf{R_{misc}}}{\mathbf{R_r}}} \qquad \text{(A4.3)}$$

The purpose of the a ground system is to minimize the Rg/Rr term. It should be kept in mind that it's not necessary for Rg/Rr to be zero. When Rg/Rr becomes small compared to the sum of the other terms then that ground system is in a the region of diminishing returns. Before designing the ground system we need to maximize Rr as described in chapters 3 and 4, minimize RL as described in chapter 6 and also minimize conductor loss (Rc). When this has been done we can judge how extensive the ground system should be.

A4.4 Soil characteristics

For this discussion some soil electrical definitions will be helpful.

$\boldsymbol{\sigma}$ = soil conductivity in Siemens/meter [S/m], Siemen=Mho

$\boldsymbol{\varepsilon_o}$ = permittivity of a vacuum = 8.854 X 10^{-12} [Farads/m]

$\boldsymbol{\varepsilon_r}$ = relative permittivity or relative dielectric constant

$\boldsymbol{\varepsilon}$=$\boldsymbol{\varepsilon_o \varepsilon_r}$= effective permittivity or dielectric constant [Farads/m]

$\boldsymbol{\mu_o}$ = permeability of free space = 4π 10^{-7} H/m

$\boldsymbol{\omega}$=2πf

f= frequency

We can combine $\boldsymbol{\omega}$, σ, $\boldsymbol{\varepsilon_o}$ and $\boldsymbol{\varepsilon_r}$ into the loss tangent (**D**).

$$\mathbf{D} \equiv \tan \boldsymbol{\delta} \equiv \frac{\sigma}{2\pi f \varepsilon_o \varepsilon_r} = \frac{\sigma}{\omega\varepsilon} \quad \text{(A4.4)}$$

A4.5 Ground system geometries

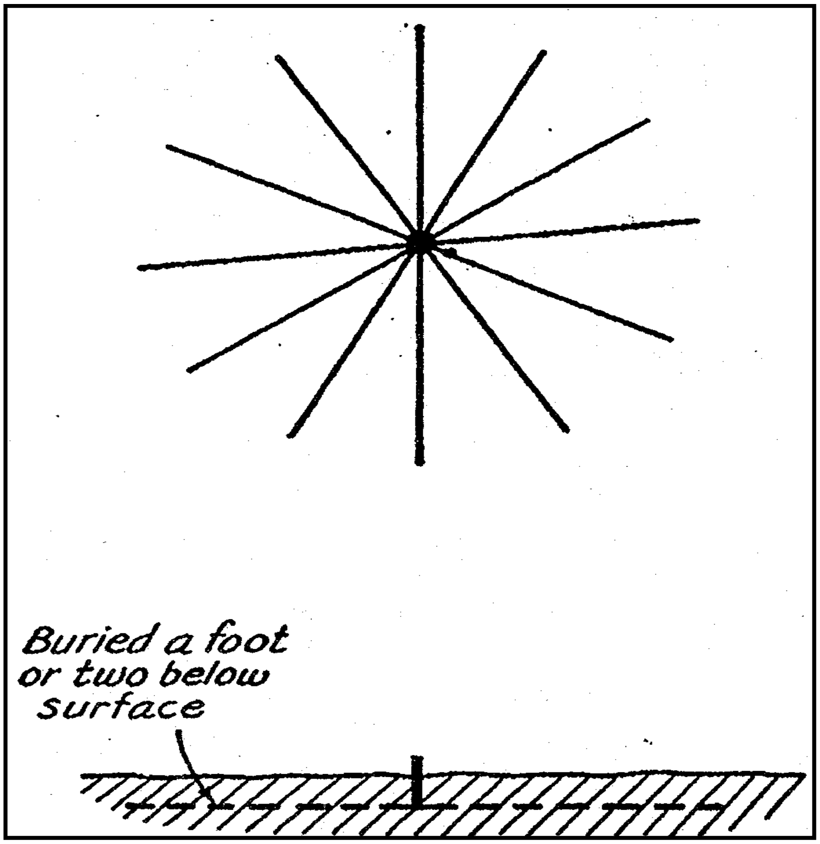

Figure A4.3 -Typical radial wire ground system.

We have many choices for ground systems, from a simple ground stake to a complex web of wires which may be elevated above ground or lying on the ground surface or buried in the soil a short distance (usually 3"-12"). Figure A4.3 shows a typical radial wire ground system for a simple vertical. In the case of an elevated system the radials may be long enough to be resonant. While this is often practical on 160m, on the lower LF-MF bands it's usually not so non-resonant or "capacitive" systems are used.

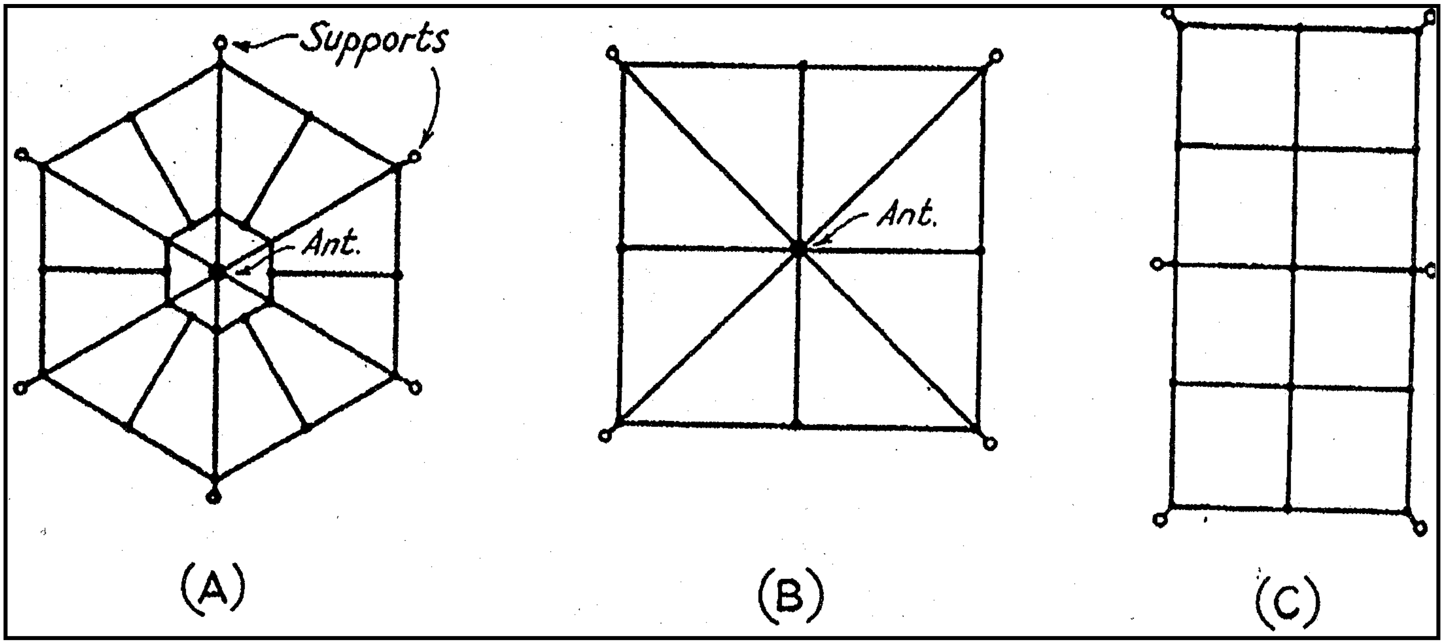

Figure A4.4 - Alternative wire ground systems.

Figure A4.4 shows some alternative wire ground systems. Very often these are elevated and, when non-resonant, referred to as "counterpoises". However, these systems can also be placed on the ground surface. In particular the rectangular wire grid shown in (C) is often placed under vertical loop transmitting antennas. It's widely assumed that a vertical loop transmitting antenna does not require a ground system and this is true, the ground system is not "required". What is often not appreciated however, is the substantial ground loss associated with loops placed close to ground. Adding a ground system like that shown in (C) can substantially improve the efficiency.

A4.6 Models for ground systems

Several different models are used to explain ground systems. They vary from simple to mathematically complex and their explanatory power varies from limited to very detailed.

Figure A4.5 is the classic model seen in amateur literature. The figure shows the radials elevated but the basic argument is the same when the radials are buried. The idea is that the vertical is capacitively coupled to the both the radial system and the soil via the displacement currents (D) flowing in the capacitances. Some D flows directly to the radials and then back along the radials to the vertical as conduction currents (I). However, some of D flows into the soil and then back into the radial system. Conduction currents flowing in the soil result in dissipation (i.e. Rg). This simple model has considerable explanatory power! We can see that more and/or longer radials increase the coupling to the radials and partially shield the soil, reducing the soil current and Rg. We can also see that the capacitance from the soil to the radials will be significantly reduced when the radials are elevated. In practice even a small elevation results in significantly less ground current.

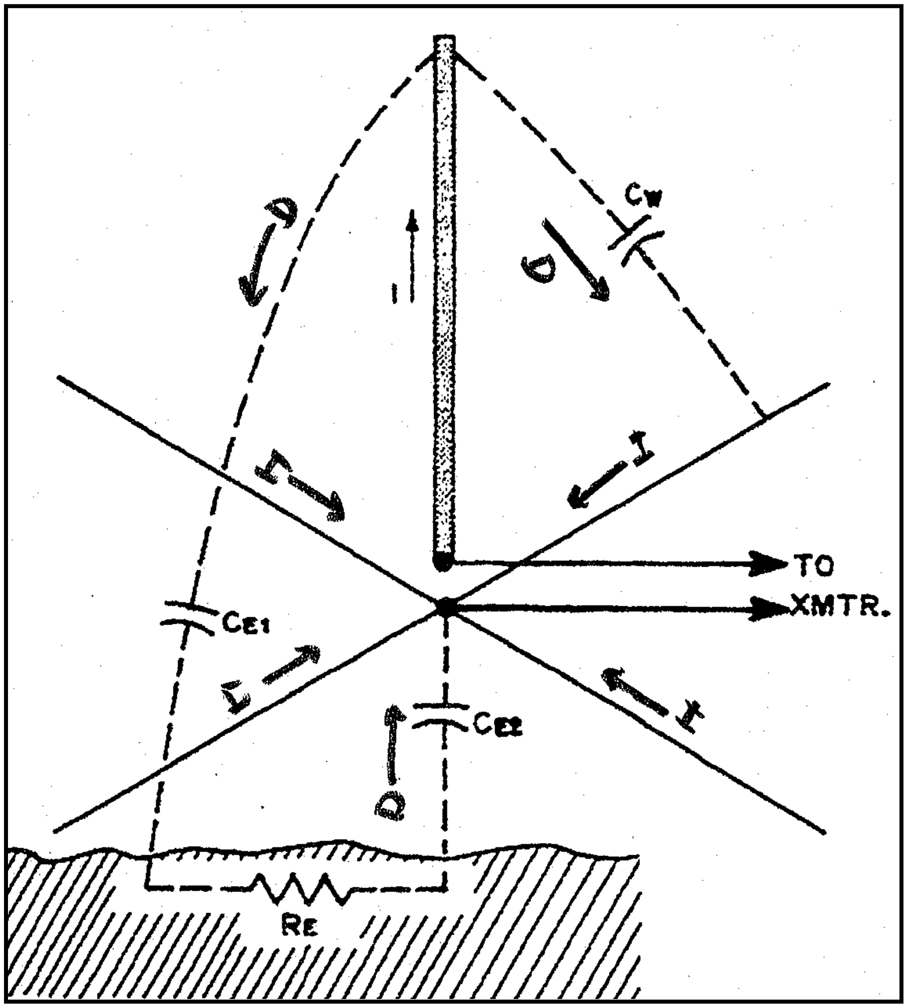

Figure A4.5 - Simple capacitive equivalent circuit model.

However, this model is not very useful if we want to determine the specific details regarding the currents and associated ground loss. For that kind of information we can change to the model shown in figure A4.6. This model recognizes that the current in the vertical creates an electromagnetic field around the antenna. As indicated there will be both electric (E-field) and magnetic (H-field) components. The fields interact with the radial system and ground introducing loss. While this model can determine Rr and Rc accurately it's very complex mathematically.

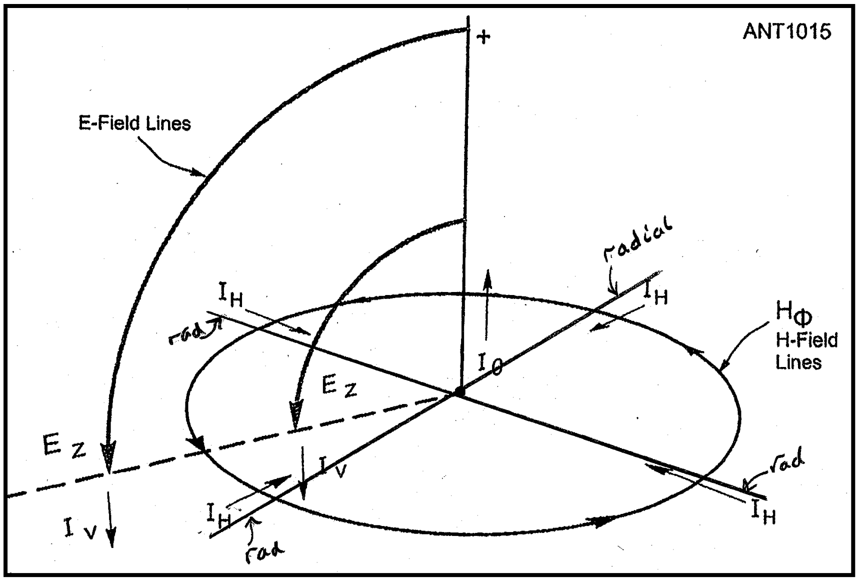

Figure A4.6 -Field model for a vertical and ground system.

Fortunately there is an intermediate model derived from optics. When an antenna is placed over soil some of the input power (Pi) is radiated and some is absorbed in the soil. One of the earliest quantitative analysis regarding ground loss and propagation of radio waves over lossy soil was done by Arnold Sommerfeld. About 1896[37] he solved the general problem of the diffraction of electromagnetic waves (EM) in lossy media, i.e. reflection and refraction at an interface between two media, air and soil. Some years later Sommerfeld used this insight to solve the general problem of waves interacting with and propagating over real lossy ground[44, 45]. The "Sommerfeld Ground Model", is still widely used. His analysis was based on the diffraction theory which represents the physical processes.

We can understand his view with an analogy to a lamp placed over the surface of a pond as shown in figure A4.7(A). We know that if a light source (the lamp) is placed over the surface of a pond some of the light will be reflected from the surface but the rest will be refracted into the water and absorbed. Light is electromagnetic radiation. Radio waves are also electromagnetic radiation only much lower in frequency. As shown in figure A4.7(B), instead of a lamp over a pond, we could substitute a short vertical conductor carrying an RF current. This short conductor is an antenna. A portion of the radiation is reflected from the soil and the rest is refracted into the soil and dissipated. The lost radiation is the ground loss, Pg. In most recent editions of Antenna Book and other texts the reflection part of figure A4.7 is discussed at length in the context of the formation of the far-field pattern.

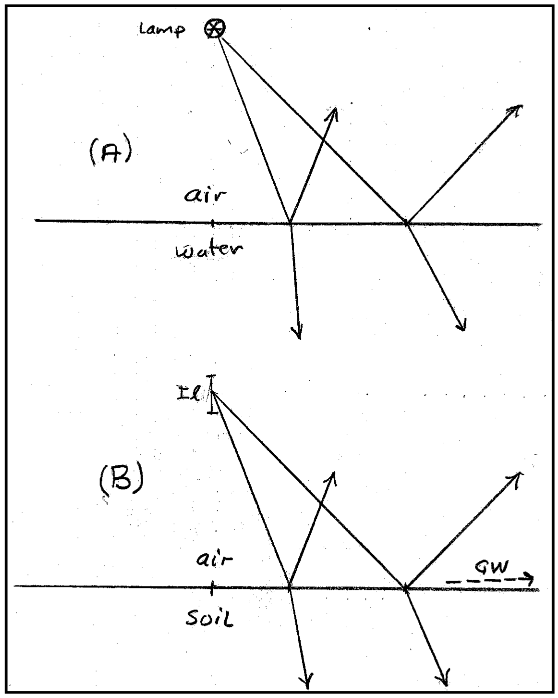

Figure A4.7 - A lamp over a pond (A) and a short vertical antenna over soil (B).

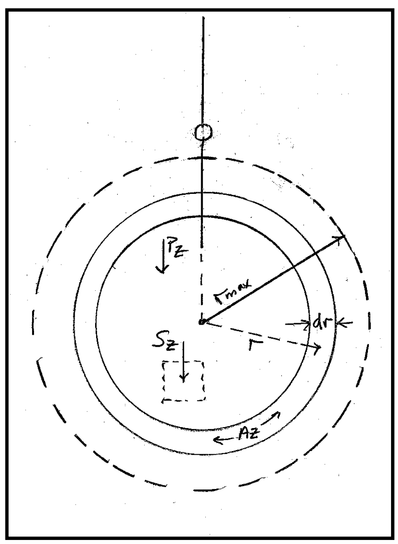

Figure A4.8 - A short vertical over ground.

By way of an example we can shift the view point as shown in figure A4.8. Sz represents the intensity of the power flow in W/m^2 into the ground surface at a given point on the surface. For this part of the discussion we will let Pg=Pz, the total power flowing into the soil at the surface within a radius r. Az=2πrdr is the area of ring at radius r with width dr. rmax is the maximum radius which for this example will be 1000' (≈λ/2 at 475 kHz). H= 40' with the bottom end 10' above ground. The frequency is 475 kHz.

With Pi=1W we can ask "what is the total power dissipated in the soil near the antenna, for radii (r) out to 1000'. Using techniques shown in Appendix A.tbd we can calculate and graph the intensity of the power flow (Sz) across the air-soil interface into the soil as shown in figure A4.9 for three different soils. We can take a further step and sum the power flow through the ring Az as shown in figure A4.10. Finally we can sum Sz within a radius rmax to get the total power flowing into the soil as shown in figure A4.11.

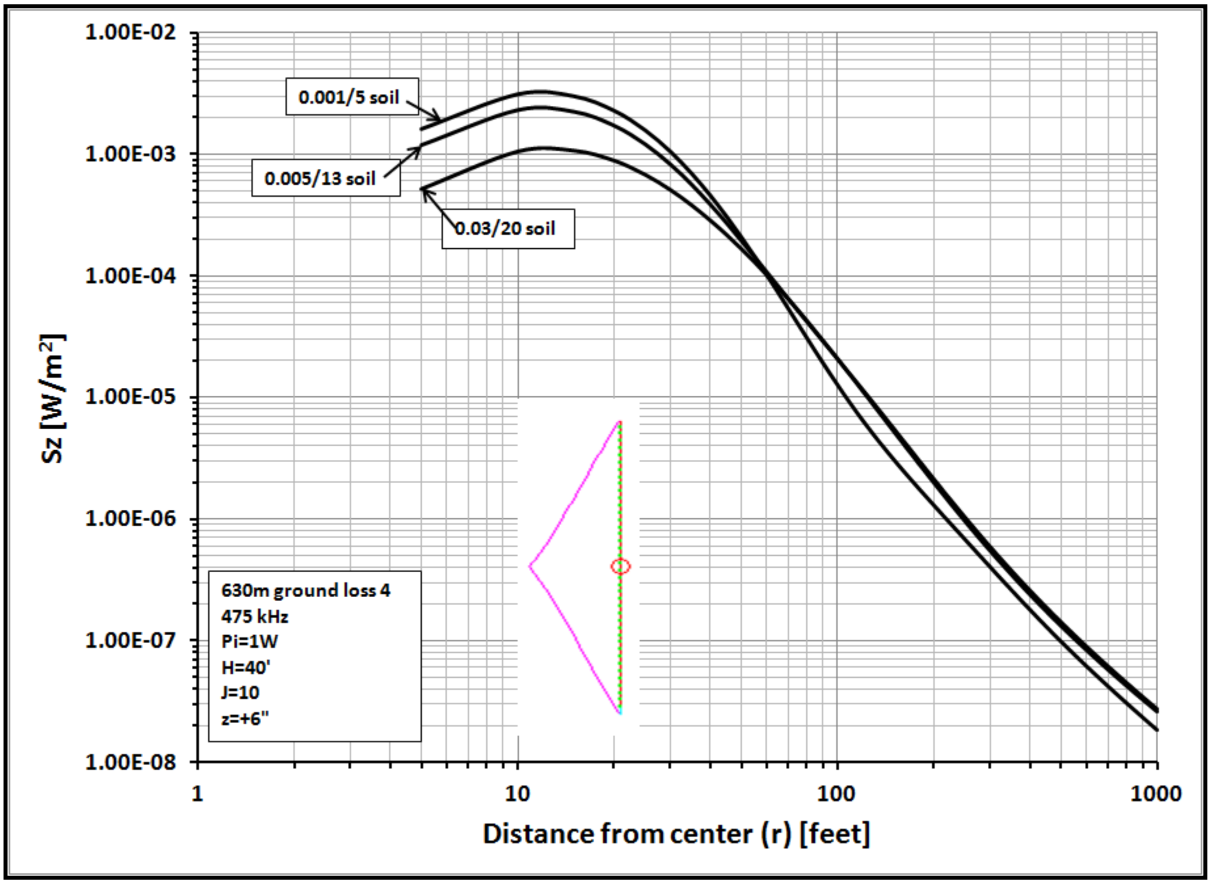

Figure A4.9 - Power density (Sz [W/m^2]) near soil surface versus radius r.

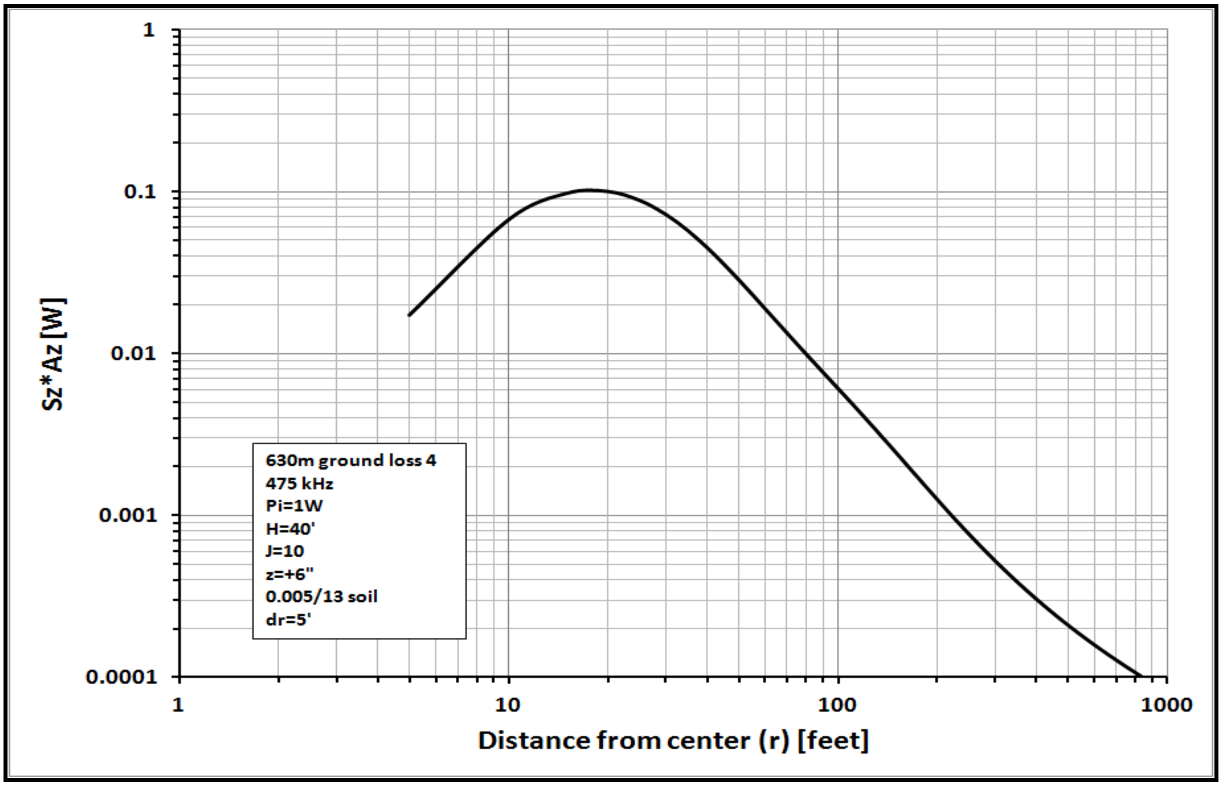

Figure A4.10 - Power dissipation in a ring (dr=5') at a distance r.

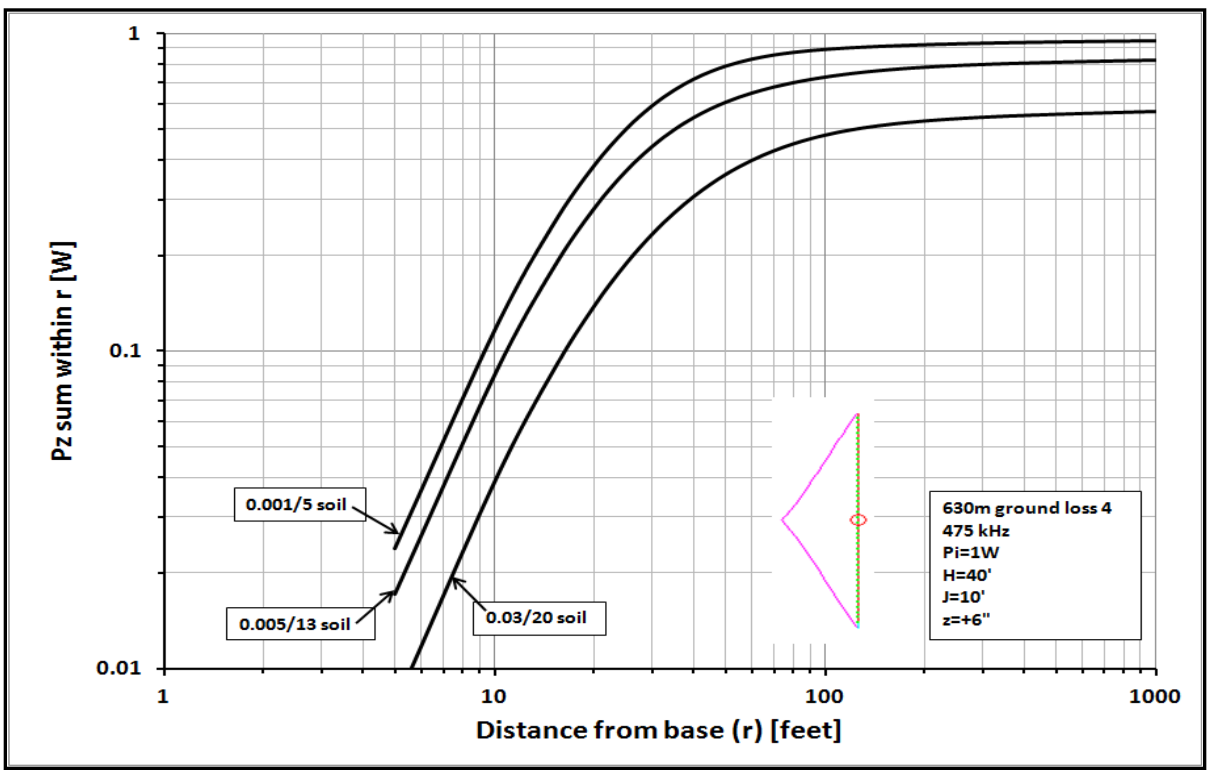

Figure A4.11 - Power flow (sum of Sz) into the soil within a given radius, 6" above the soil surface.

These graphs show several interesting things. First, most of the power is lost close to the base, within 50'≈0.025λ in this example. Concentration of power loss close to the base is typical of verticals and tells us that when installing a ground system we need to pay particular attention to the ground system near the base.

Second, the loss is strong function of the soil characteristics. The dominant influence at MF is the conductivity but a HF ε_r becomes important. For this example, the effect of different soils, including perfect ground, on Rr and Rg is summarized in table A4.1.

Table A4.1 - summary

soil=	ideal	0.001/5	0.005/13	0.03/20
Ri=	0.14Ω	2.736 Ω	0.7483Ω	0.3258Ω
Rr=	0.14Ω	0.134Ω	0.1412Ω	0.1403Ω
Rg=	0Ω	2.601Ω	0.6433Ω	0.1855
Xc=	10160Ω	10160Ω	10160Ω	10160Ω
Io=	2.673Arms	0.6046Arms	1.129Arms	1.752Arms
Pi=	1W	1W	1W	1W
Pr=	1W	0.049W	0.18W	0.43W
Pg=	0W	0.95W	0.82W	0.57W

Third, in this example Pi=1W so we see that that for the poor soil almost 95% of the radiated power is absorbed in the soil! This indicates the need for a ground system.

In the case of the lamp over the surface of a pond if we wanted to reduce the light lost into the water we could simply install a mirror on the surface under the lamp. The greater the mirror diameter the less light lost. We can do exactly the same thing with the antenna in figure A4.9 by placing a buried radial system under the antenna as shown in figure A4.12. In this example there are 60 radials buried 1' in average soil (0.005/13). The radials are connected at the center but are not connected to the vertical conductor. The radial lengths are varied from 50' to 150' for the Pz calculations shown in figure 4.13.

The radial system is not a perfect "mirror". It's effectiveness will depend on the number of radials and their length. More numerous and/or longer radials make for a better "mirror" and lower soil loss.

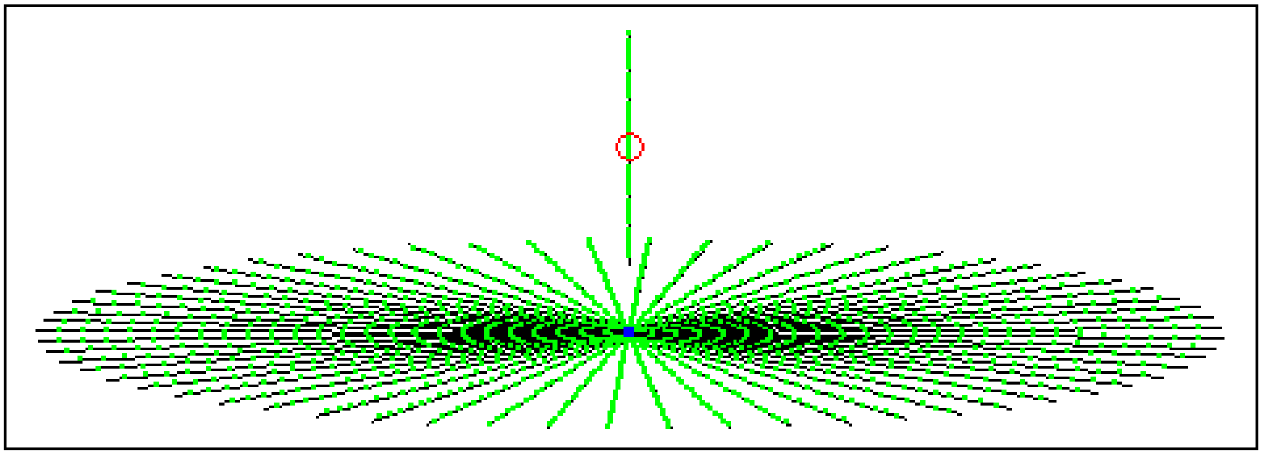

Figure A4.12 - Antenna with a buried radial ground system.

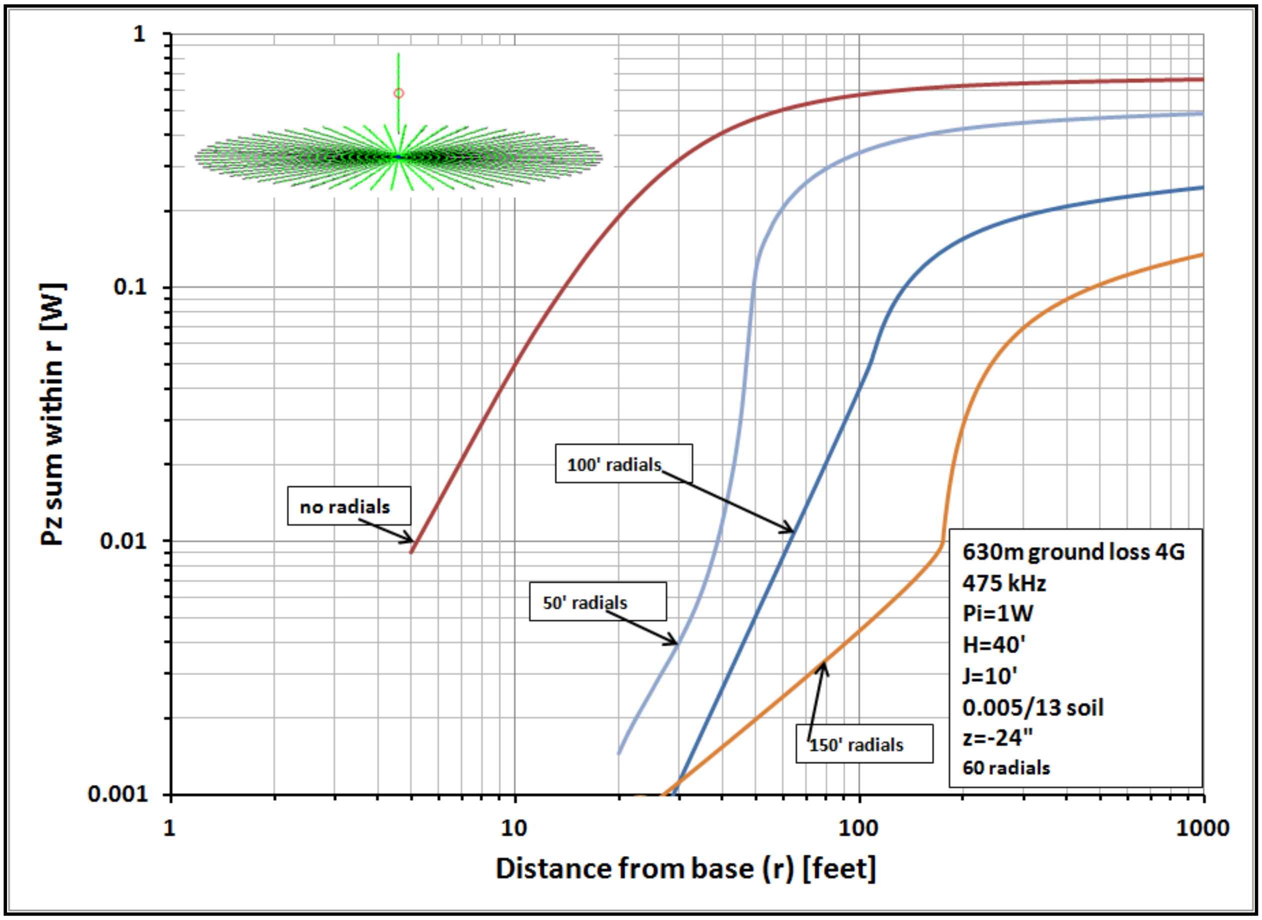

Figure A4.13 - Pz sum: 60 radials, 50', 100' and 150', z=-24".

From figure A4.13 we can see how effective a radial ground system can be in reducing Pg in the soil under a vertical. As the radials are made longer the power absorbed under the ground system is drastically reduced, extending the radial lengths extends the radius of the "mirror". The number of radials is also important because the reflectivity of the ground system improves as the wires are brought closer together which is what happens when more numerous radials are used as illustrated in figure 4.14.

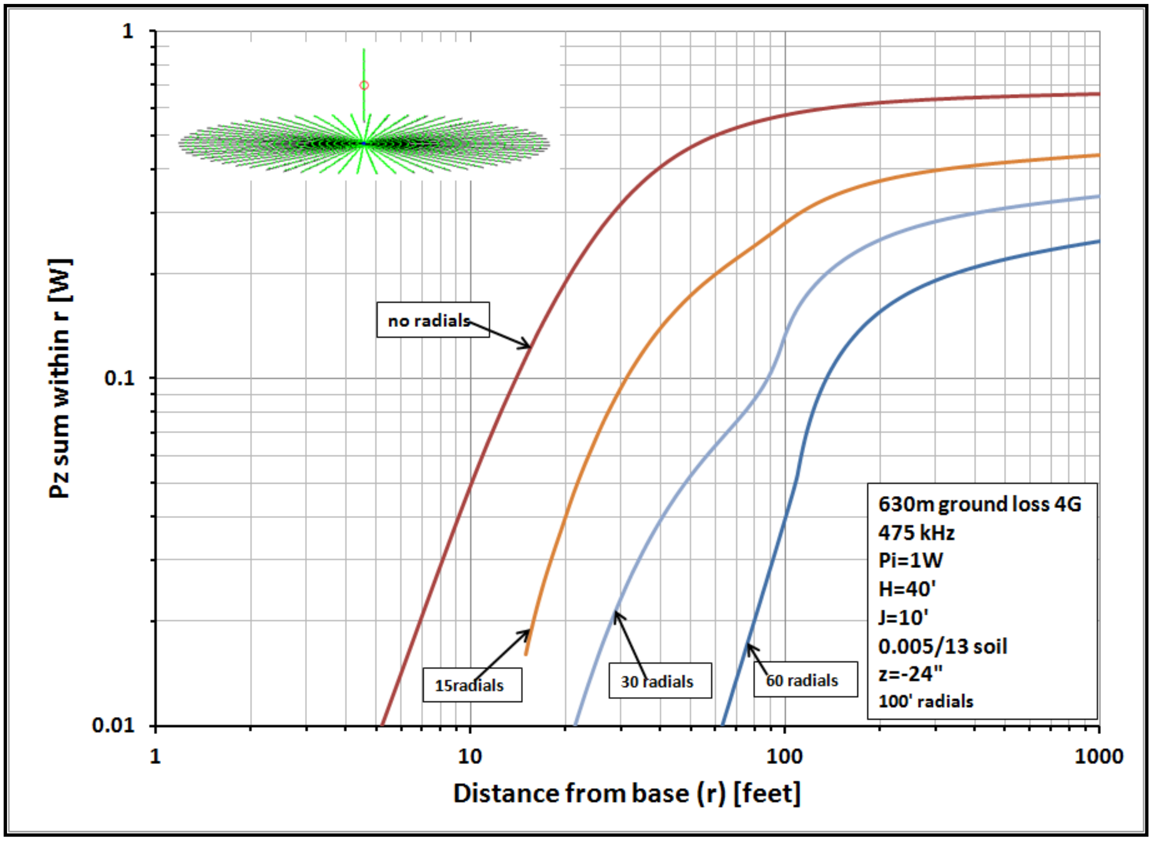

Figure 4.14 - The effect of radial number on Pz.

A4.7 Pg in a grounded vertical

The antenna example in figure A4.9 was useful for explaining Pg but the antenna itself is not very practical, at least at LF-MF. Let's now look at a much more typical 630m amateur antenna, the top-loaded vertical shown in figure 4.15 with two different ground system options: a single long ground stake and an extensive buried radial wire system. H=50' and the 8-spoke hat has a radius of 15' and a skirt wire. The ground stake is 1" in diameter and extends into the soil 50'. The radial system has thirty 50' radials buried 1'.   This exercise illustrates the shortcomings of a simple ground stake system.

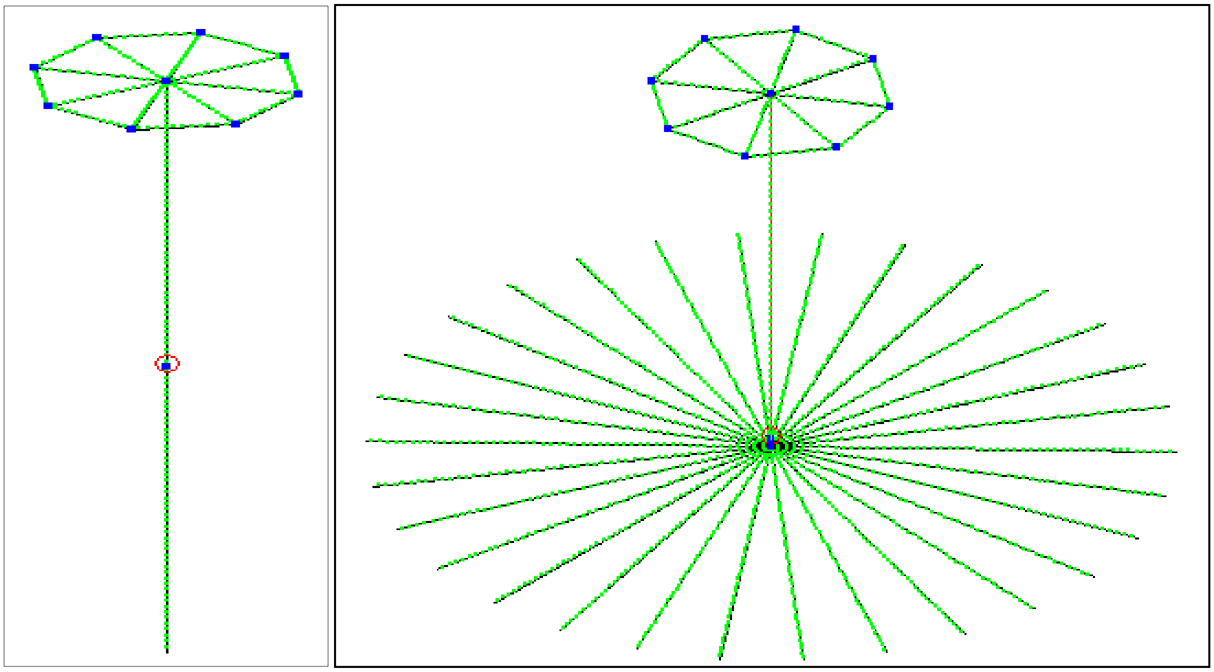

Figure 4.15 - 630m top-loaded vertical with two ground systems.

As indicated in figure 4.16, Pz represents the power radiated downward into the soil and Px represents the power radiated from the ground rod.

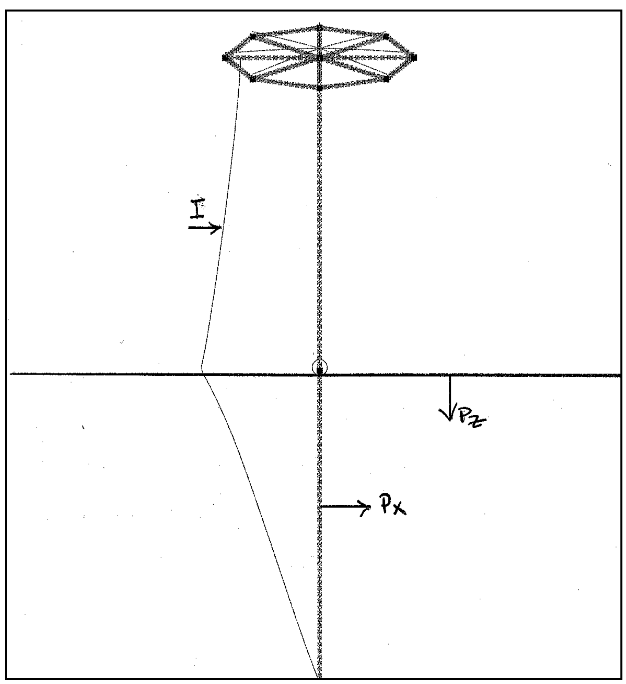

Figure 4.16 - Ground loss model using refraction.

When we do the calculation assuming RL=0 and Pi=1W, we get the following result: Pr=0.03W, Pz=0.27W and Px=0.70W. The efficiency is about 3% with 27% the power being refracted into the soil from the upper part of the antenna and 70% is radiated from the lower half of the antenna directly into the soil! this picture also emphasizes that the ground rod is not just a auxiliary element, it is a radiating part of the antenna!

If we replace the ground rod with the radial system we find for Pi=1W, Pr=0.24W and Pg=0.76W. The radial ground system increases the radiation efficiency from 3% to 24%! A larger ground system would further improve the efficiency.

At this point we've seen a general argument why we might want to use a ground system with a vertical. The practical details are more complicated and there are different kinds of ground systems, not just buried radial systems. More details are given in chapters 5 & 6.

4.8 Radial systems as reflectors

The concept that a wire ground system could be viewed as a mirror or reflector for EM waves has been carefully investigated from the very earliest days of Radio by many researchers[37-43]. There are several variables: polarization of the EM wave, arrangement of the wires, the distance between the wires and the ground and the characteristics of the soil. Solving this problem for the general case requires some pretty advanced mathematics but fortunately the nature of the fields around a verticals usually allows us to use a simpler approximation.

We can understand the interaction of the field on this combination of soil and grating by using a transmission line analogy shown in figure A4.17.

A wave traveling in free space is equivalent to a wave traveling along a ideal transmission line with a characteristic impedance Zo. The space above ground is represented by a parallel wire transmission line with an impedance equal to free space, i.e.:

$$Zo = \sqrt{\frac{\mu_o}{\varepsilon_o}} = \frac{E}{H} = 376.7\Omega \quad \text{(A4.5)}$$

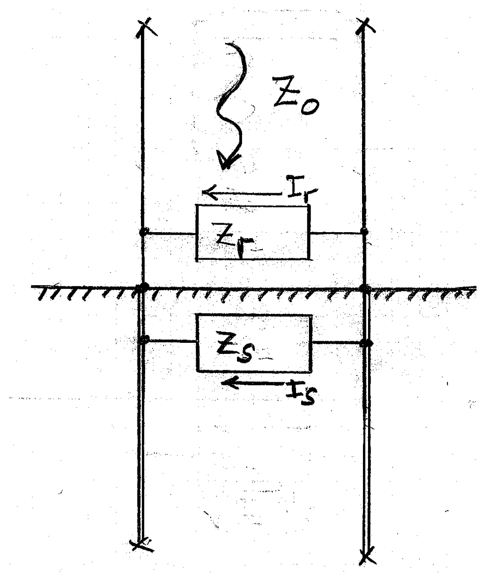

Figure 4.17 - Transmission line analogy.

The grating can be viewed as an array of parallel two-wire transmission lines with a characteristic impedance Zr:

$$\boldsymbol{Zr = jXr = j\left[f\mu_o d' \ln\left(\frac{d'}{\pi d}\right)\right]} \quad \text{(A4.6)}$$

Where: ***d'***= wire spacing and ***d***=wire diameter in meters. Note that Zr is an inductance with no dissipation. This is not strictly correct, ground system wires will have some loss but it's usually small. We also have to realize the equation (A4.6) is valid only for d'<<λ[tbd-Abbott]. Typically you have to keep d'<2.5 m at 475 kHz.

When the ground system is arranged in a radial fan like that shown in figure A4.8 the spacing (d') between the radial wires will vary with the distance from the base (L') and the number of radials (N):

$$\boldsymbol{d' = 2L' tan\left(\frac{\pi}{N}\right)} \quad \text{(A4.7)}$$

The soil can also be represented by a transmission line but we have to take into account the conductivity (**σ**) and relative permeability (**εr**) to determine the characteristic impedance Zs:

$$Zs = \frac{Zo}{\sqrt{\varepsilon_r}}\left(\frac{1}{\sqrt{1-jD}}\right) \quad \text{(A4.8)}$$

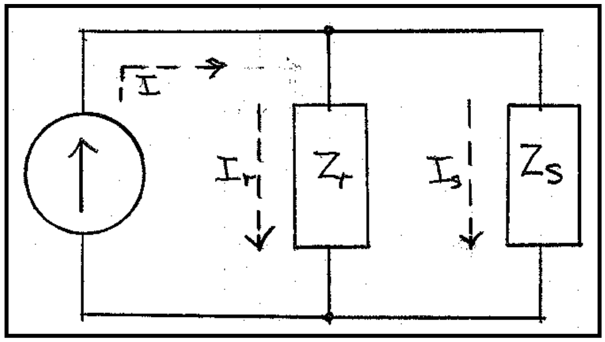

Figure A4.15 - Zr-Zs equivalent circuit.

As indicated in figure A4.14 We can join the three transmission lines (air, wire grid and soil) which gives us the equivalent circuit shown in figure A4.15. Keep in mind this is an analogy, I represents the incoming wave, Ir represents the portion of the wave diverted into the ground system and Is represents the portion of the wave absorbed in the soil.

If Zr is small compared to Zs then very little energy will be delivered to Zs and lost. What we want is Zr <<Zs so the majority of the current flows in the ground system and not the soil. The wave energy goes into inductive storage in Zr and is reradiated, i.e. reflected. We can use this model along with some spreadsheet calculations to obtain guidance on wire spacing, wire size and radial numbers in ground systems. Most soils are capacitive so Xs is usually negative.

For calculations it helps to restate Zs as:

$$Z_s = R_s + jX_s \quad \text{(A4.9)}$$

Where:

$$R_s = |Z_s|cos\theta, \quad X_s = -|Z_s|sin\theta, \ \theta = \frac{tan^{-1}(D)}{2}$$

$$D \equiv \frac{\sigma}{\omega\varepsilon}, \qquad |Z_s| = \frac{Z_o}{\sqrt{\varepsilon_r}}\left[\frac{1}{(1+D^2)^{1/4}}\right]$$

The power loss in the soil is proportional to:

$$Pg \propto \left|\frac{Is}{I}\right|^2 \quad \text{(A4.10)}$$

The smaller this ratio the better!

From figure A4.15:

$$\left|\frac{Is}{I}\right| = \frac{|Zr|}{|Zr+Zs|} = \frac{Xr}{\sqrt{Rs^2+(Xr+Xs)^2}} \quad \text{(A4.11)}$$

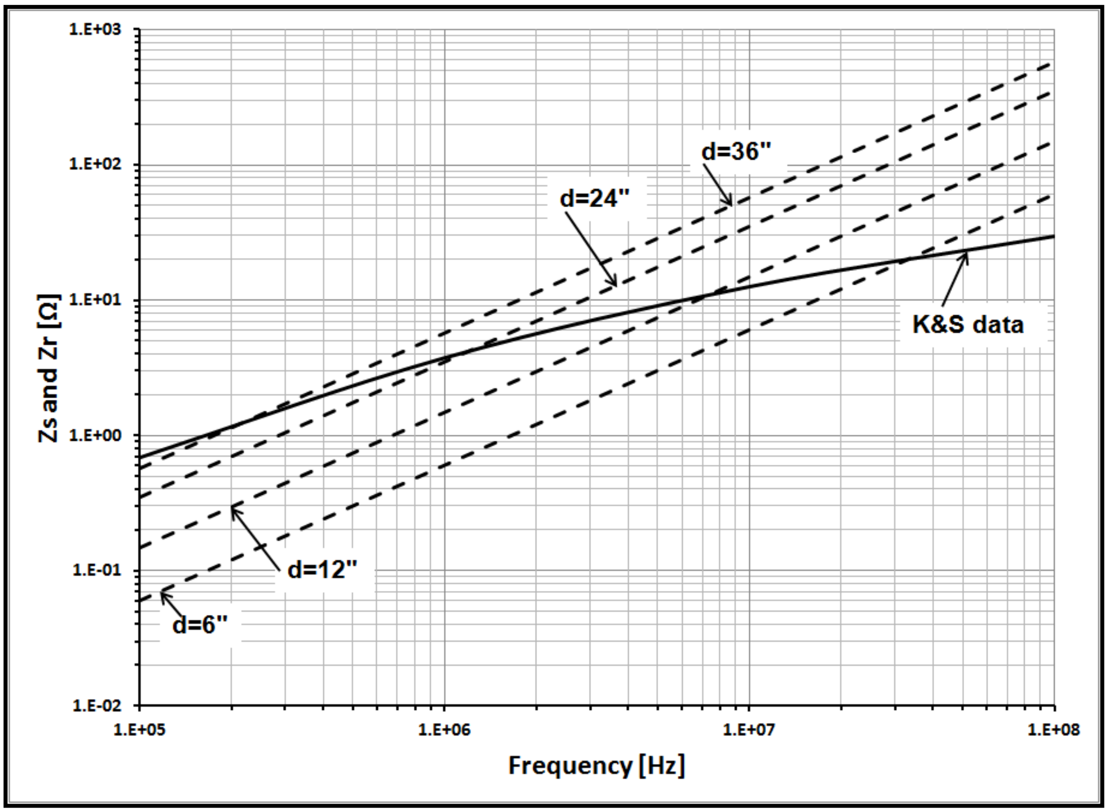

Figure A4.16 - Soil and transmission line impedances.

The solid line in figure A4.16 represents the value of Zs for a typical soil as a function of frequency (100 kHz- 100 MHz). The data from which Zs was calculated was taken from King and Smith[tbd]. This particular data set is for moderate conductivity soil (σ≈0.01 S/m and Er≈60 @ 475 kHz). The dashed lines in figure A4.16 represent Zr for different spacing's (d' =6"-36").