Introduction to the Propagating Wave on a Single Conductor

Glenn Elmore

Corridor Systems Inc.

www.corridor.biz

Abstract

An overlooked solution to the Maxwell-Heaviside equations supports the existence of a propagating TM surface wave on coaxial cable as well as on a completely unshielded single conductor. This non-radiating surface wave mode exhibits attenuation much lower than coax and a relative propagation velocity of unity. It is very broadband and has practical applications from RF through microwave frequencies and beyond. This article introduces this mode, measurements and describes applications. In particular, this article describes the use of the new mode with conventional overhead power lines as a 3rd pipe and solution to the last mile problem.

Background & History

Conventional Model of Coaxial Line

Coaxial cable is perhaps the most commonly used transmission line type for RF & microwave measurements and applications. In 1894 Heaviside, Tesla and others received patents for coaxial line and related structures. A development of coax (coaxial line) theory is often provided as part of basic physics and engineering education[1], even prior to full development and use of the Maxwell-Heaviside equations, which are generally used for transmission line and macroscopic electromagnetic analysis. Accordingly, the analysis, measurement and application of coax is usually considered to be quite mature and complete.

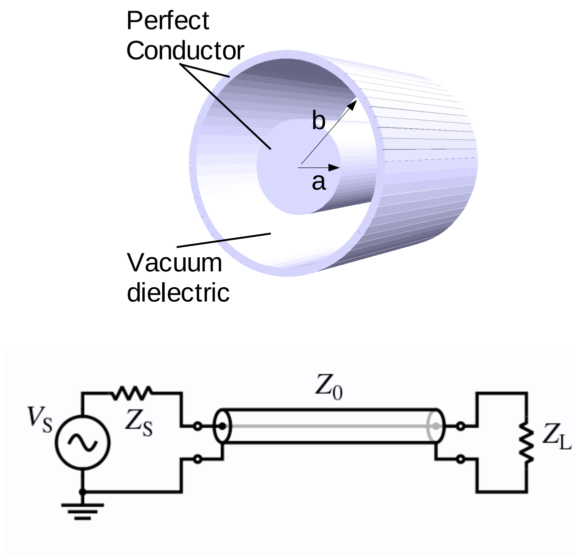

Illustration 1: Coaxial transmission line used to deliver source power to a load.

Introductory descriptions of coax often proceed along the lines of Illustration 1. Here lossless cylindrical central and outer shielding conductors are separated by a volume of empty space. This structure is examined as a means for conveying power between two points. One end is considered an input port and driven with a sinusoidal voltage source

$$V_s = A\sin(\omega t) \tag{1}$$

of magnitude A at frequency ω .

This source is applied to the line through a known impedance, Z_S . The other end of the line is terminated by a load of impedance Z_L .

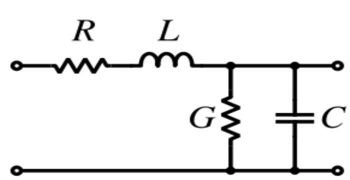

Illustration 2: Schematic model of a transmission line from the Telegraphers' Equation

Heaviside's telegraphers' equation provides a lumped circuit equivalent of an infinitesimal length of transmission line, shown in Illustration 2. For the lossless case where $R=G=0$ Ampere's law can be used to find the inductance per unit length

$$L_l=\frac{\mu}{2\pi}\ln\left(\frac{b}{a}\right) \tag{2}$$

and Gauss's law to find the capacitance per unit length

$$C_l=\frac{q}{V\,l}=\frac{2\pi\epsilon}{\ln\left(\frac{b}{a}\right)} \tag{3}$$

this describes a line with an entirely real characteristic impedance[2] of

$$Z=\sqrt{\frac{L_l}{C_l}}\ \text{ohms} \tag{4}$$

which is dependent only on the geometry of the conductors $\left(\frac{b}{a}\right)$.

Maximum transfer of power between source and load occurs when all of these impedances are equal and

$$Z=Z_S=Z_L \tag{5}$$

Current entering the line central conductor produces a real current density, $\boldsymbol{J}$. By Ampère's circuital law, this current density produces an orthogonal magnetic flux density $\boldsymbol{B}$ field (in vector form)

$$\nabla\times\boldsymbol{B}=\mu\boldsymbol{J} \tag{6}$$

in the region of empty space inside the outer conductor. An equal magnitude but opposite sense current density, $-\boldsymbol{J}$ returning from the outer shield also contributes to magnetic flux within this region. Beyond this region the magnetic effects exactly cancel and no fields due to the currents are present. This cancellation provides the shielding nature of coax.

Between the conductors the varying $\boldsymbol{B}$ field produces an electric field

$$\nabla\times\boldsymbol{E}=-\frac{\partial}{\partial t}\boldsymbol{B} \tag{7}$$

Electric field lines extend between the conductors and are normal to their surfaces. These electric and magnetic fields produce a transverse electric-magnetic wave traveling along the line in the space between the two conductors. In ideal coax this wave travels in the vacuum dielectric without attenuation and with velocity the same as that of light in a vacuum.

Waves propagating on transmission lines can be described in terms of the axes of the electric and magnetic fields and a mode number. One or both of the electric and magnetic fields must be transverse to the direction of propagation. The corresponding modes are TE, for transverse electric field, TM for transverse magnetic field and TEM when both field types are transverse. A pair of mode numbers, **n** and **m**, can be associated with these which represent the order of the mode in the transverse and longitudinal directions, respectively. Values of zero for each of these describe a principal mode in the corresponding direction.

For a coax line of infinite length and for wavelengths large compared to the inner circumference of the outer conductor

$$\lambda \gg 2\pi b \tag{8}$$

there is radial symmetry and the coaxial line exhibits a principle TEM_{00} propagation mode. The impedance presented to the source by the line can be written as,

$$Z_{TEM} = \frac{1}{2\pi}\sqrt{\frac{\mu}{\epsilon}}\ln\left(\frac{b}{a}\right) \quad \approx 60\ln\left(\frac{b}{a}\right) \tag{9}$$

where

$\mu \stackrel{\text{def}}{=} 4\pi x\, 10^{-7}$ Henry/meter $\approx 1.2566\,\mu$H/meter, permeability of a vacuum

$\epsilon = \dfrac{1}{c^2 \mu}$ Farad/meter $\approx$ 8.8542 pF per meter, permittivity of a vacuum

For the matched condition described the voltage produced by the wave at a position separated from the source by distance, l , along the line can be described as

$$V = \frac{A}{2}\sin(\omega t)\, e^{-\gamma l} \tag{10}$$

where

$$\gamma = \alpha + j\beta$$

is the propagation constant. α describes the attenuation while β describes the phase, per unit length of line. The propagation constant for the principle mode can be shown to relate to the components in Illustration 2 by

$$\alpha + j\beta = \sqrt{(R + j\omega L)(G + j\omega C)} \tag{11}$$

which for the lossless case is purely imaginary and the same as that of the enclosed medium[3].

Practical cables require dielectric supports and use imperfect conductors which complicate the model but more than a century of use has validated this basic understanding of coaxial line and its application to the solution of real world problems. For most applications from RF through upper microwaves, conveniently dimensioned coaxial cable has proven to be an excellent device for transferring electromagnetic energy between different locations without significant radiation; effectively shielding the internal wave from external components and circuitry.

The Propagating TM Wave in Coax

A homogeneous plane wave in an isotropic medium has an intrinsic impedance[4]

$$Z=\sqrt{\frac{\omega\mu}{\omega\epsilon+i\sigma}} \qquad (12)$$

in free space where

$$\sigma=0$$

this reduces to

$$Z=\mu c \quad =\sqrt{\frac{\mu}{\epsilon}} \quad \approx 120\pi \text{ ohms} \qquad (13)$$

In coax, as the geometry $\frac{b}{a}$ increases, the impedance of the TEM_{00} mode increases logarithmically and the real current density, **J**, tends toward zero. Equating (2) and (3) shows that the impedance of the TEM_{00} mode in coax equals that of free space when

$$\ln(\frac{b}{a})=2\pi \quad \Rightarrow \quad (\frac{b}{a})\approx 535 \qquad (14)$$

However, energy may not propagate faster than the speed of light in a vacuum.

$$c=\frac{1}{\sqrt{\mu\epsilon}}=\frac{Z}{\mu} \qquad (15)$$

Just as for a planar wave in free space, energy propagating through a lossless coaxial transmission line having vacuum dielectric and no magnetic materials is subject to this constraint. The impedance associated with the propagating energy in a transmission line is bounded by the permeability and permittivity of space. Energy may propagate simultaneously by way of a hybrid of multiple modes but the combined impedances and the combined admittances of the propagating modes are bounded such that for the total propagating wave

$$Z_{total}=\frac{1}{Y_{total}}\leq\sqrt{\frac{\mu}{\epsilon}} \qquad (16)$$

For a line with dimension meeting (8), due to symmetry, only modes with a transverse magnetic component, either TEM or TM, are possible since any asymmetric modes that would produce a longitudinal magnetic component will be immediately damped out[5]. Only TEM_{0m} or TM_{0m} modes can propagate. Additionally, for perfect conductors only the principal modes are supported[6]. Therefore only TEM_{00} or TM_{00} are possible. In coax of this type the combined admittances of these must be bounded such that

$$Y_{total} = Y_{TEM_{00}} + Y_{TM_{00}} \geq \sqrt{\frac{\epsilon}{\mu}} \approx 2.65\,x\,10^{-3}\,\text{mho} \tag{17}$$

The admittance due to the TEM_{00} mode

$$Y_{TEM_{00}} = \frac{1}{Z_{TEM_{00}}} = \frac{2\pi\sqrt{\frac{\epsilon}{\mu}}}{\ln\left(\frac{b}{a}\right)} \tag{18}$$

is positive, finite and continuous over the range

$$1 < \left(\frac{b}{a}\right) < \infty \tag{19}$$

So at least for the case where

$$\ln\left(\frac{b}{a}\right) > 2\pi$$

A propagating TM_{00} mode must also exist and provide a finite admittance

$$\sqrt{\frac{\epsilon}{\mu}} \leq Y_{TM_{00}} < \infty \tag{20}$$

All propagating modes are solutions to the wave equation which results from Maxwell's equations and satisfy the requirements for continuity of fields at the conductor-vacuum boundary. Combinations of Bessel functions are used to describe the fields and impedances associated with these solutions. These functions and their first derivatives have singularities only at zero and infinity and are continuous in between. Therefore, the fields and waves they describe also are without discontinuities over the intermediate region.

As a result, contrary to longstanding belief to the contrary, in coax there exist simultaneous propagating TEM_{00} and TM_{00} modes over the entire range of geometries

$$1 > \left(\frac{b}{a}\right) > \infty \tag{21}$$

The Propagating TM Wave as a Surface Wave on a Single Conductor

The TM wave on a single conductor embedded in a dielectric medium, can be viewed as a surface wave along the inner conductor of a coax line having infinite geometry. In this view, for finite V_s , real current density vanishes:

$$\boldsymbol{J_r} \Rightarrow 0 \text{ as } (\frac{b}{a}) \Rightarrow \infty \tag{22}$$

However from the Maxwell-Heaviside equations, the total magnetic field is due to *both* real current $\boldsymbol{J_r}$ involving moving charges and to displacement current due to the time rate of change of the electric field $\frac{\partial \boldsymbol{E}}{\partial t}$,

$$\nabla \times \boldsymbol{B} = \mu \boldsymbol{J}_r + \mu \frac{\partial \boldsymbol{D}}{\partial t} = \mu \boldsymbol{J}_r + \mu \epsilon \frac{\partial \boldsymbol{E}}{\partial t} \tag{23}$$

As the geometry of coax increases without bound, the component of the magnetic field due to the longitudinal component of the displacement current increases at the same time that the component due to real current decreases.

$$(\frac{b}{a}) \Rightarrow \infty , Y_{TEM} \Rightarrow 0$$

and

$$Y_{total} \Rightarrow Y_{TM} = \sqrt{\frac{\epsilon}{\mu}} \approx 2.65 \, x \, 10^{-3} \, \text{mho} \tag{24}$$

In the limit, the amount of real current in the outer conductor falls to zero and the total admittance is due entirely to displacement current which produces a single principal TM_{00} mode with the same impedance as a wave in free space.

For intermediate geometries, the total admittance is due to contributions from each mode. The outer conductor provides a path for real return current which increases the total admittance. This increase in admittance reduces the potential on the line and causes an associated reduction of longitudinal displacement current and a corresponding decrease in the portion of the total power propagated via the TM mode.

Thus, conventional coax cable always propagates power by a hybrid of a principal TEM mode and a principal TM mode over the entire range of coax geometries. Both of these modes have the same propagation velocity which is determined by the relative permittivity of the enclosed dielectric. For the case of perfect conductivity and vacuum dielectric both waves travel without attenuation at the speed of light.

History

The existence, practicality and impact of the surface wave TM mode seems to have been generally overlooked. This is perhaps not so surprising in view of the small effect it has on propagation in coax of convenient geometry and common impedance, as previously described. Sommerfeld investigated surface waves[7] as did Zenneck[8], particularly involving lossy conductors as part of better understanding beyond-the-horizon radio propagation during the early 1900's. Solutions for the wave around a perfectly conducting center cylinder embedded in a dielectric were presented by Stratton in 1941. There it was found that only a single modal low-attenuation solution describing a TM_{00} wave having the same propagation constant as that of the enclosing dielectric exists[9]. Solutions for coax were also investigated but only the single principle TEM_{00} mode was described as being consequential for line dimensions that are common in communications practice[10]. In the coaxial solutions nly a single principle TEM_{00} mode was considered.

More recent characterization of precision coaxial line standards in slightly lossy line for use in vector network analysis with a reference impedance of 50 ohms also found the effect of the TM_{00} mode to be small. However, in calculating it's effect on line impedance, the H-Field and wave admittance associated only with the radial component of the electric field were included[11]. Apparently this was due to an a priori assumption that no propagating TM mode, or at least no significant mode, exists in coax and any longitudinal component of the E field would be only evanescent or so small that it could be neglected.

Perhaps most surprising is that during the 1950's the initial practical application of surface wave transmission involving only a single conductor and applications of that same work since have not uncovered the existence and usefulness of this TM mode. The seminal application of surface wave propagation over a single conductor was presented by Goubau[12]. This application, called “G-Line”, provided methods to build a practical transmission system by using special conductor surface conditioning or a surrounding dielectric material along with special launchers to convert from coax or waveguide modes to a surface wave mode on the line. In spite of the prior work by Sommerfeld and Stratton, as part of patenting[13] this system Goubau posited that a reduction of the wave velocity on the conductor was *required*, both to prevent radiation and to allow a launcher of convenient size. Adaptations of his work, including recent variations[14], have continued along these same lines of thought and this opinion seems to have persisted until the present day.

Implementation of Practical Single Conductor Lines

Single Conductor Application of the TM Wave

Since this article is intended as an introduction to the usefulness of the TM mode rather than as a complete solution to the general case of hybrid propagation of the TEM and TM modes with lossy conductors and imperfect dielectrics in coax, it will now turn toward providing some insight into practical non-coaxial application of the TM mode in real world situations.

As it is necessary to couple to and from the mode in order to access it and take advantage of it in conjunction with other traditional transmission lines such as coax and waveguide, developing a visualization of the associated electric field at this point seems useful.

E-field direction

The solution to the wave equation for the propagating TM mode produces a non-zero longitudinal component of the E-field. This is in contrast to the solution for the TEM mode in coax which produces only a transverse E-field.

Whereas the TEM mode is excited by real current, the TM wave is excited by the displacement current. The potential on the central conductor increases as line impedance increases. As these increase, the magnitude of the E-field increases as well. However, a nearby conductor other than the line itself may provide a termination point and thereby reduce energy coupled into the TM wave. This is the case with the shield of conventional coaxial cable of common geometry. The proximity of a shield reduces the TEM impedance, provides a return path for E-field lines, increases real current, reduces displacement current and correspondingly reduces the power coupled into the TM wave. The result is that as the geometry is reduced, propagation in coaxial cable rapidly becomes dominated by the TEM mode to the exclusion of the TM mode. When the geometry has reached 50 ohms in ideal coax, $\frac{b}{a} \approx 2.3$ (Illustration 1) the TM mode has been almost entirely suppressed. To examine the mode it is necessary to consider a central conductor apart from nearby shielding or conductors which can suppress it.

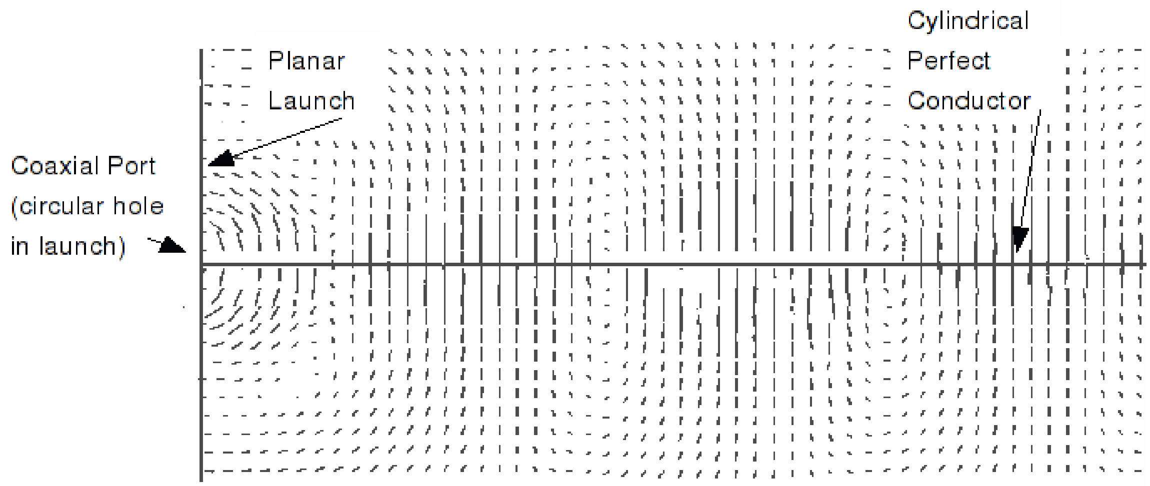

Illustration 3: E-field directions in vicinity of a perfect conductor and a planar launcher

Illustration 3 shows a plot of electric field generated from a numeric solution of Maxwell's equations performed by a 3D E-M solver (HFSS). The model is of a thin, perfectly conducting circular disk, on the left, having a central hole through which passes a perfectly conducting wire that extends continuously from left to right. The short region inside this hole is equivalent to a section of ideal coax and excitation of this port is configured to be coaxial at this location. The rest of the region in the illustration is vacuum wherein the short lines indicate the direction of the E-field that result when the port is driven by a sinusoidal signal through a port impedance equivalent to that of the TEM mode at the coaxial input at the plane of the disk.

It is important to recognize that because the TM mode has not previously been known to existent, computer analysis tools may make assumptions about the conditions at the port of a model. Even though in the analysis itself, a full numerical solution of Maxwell's equations may be performed, the port excitation for the model does not necessarily include this. For the model and plot shown above, the analysis was performed with the assumption that conditions to the left of the launcher port, that region “inside” the modeler, is a TEM extension of the port. No longitudinal E-field component is present there and as such it only models excitation from a TEM source. Because a TM wave actually does exist this causes some error. However, in this example the port geometry has been chosen to provide a relatively low impedance, in the vicinity of 50 ohms, and the TM contribution to the propagating wave is so small that the error is negligible.

This same problem exists with conventional scalar and vector network measurement and analysis of coaxial systems in general. All commercial systems of which the author is aware presently make the implicit assumption that in coax only a TEM wave exists. For fifty ohm systems this assumption has been, and continues to be, almost entirely adequate with the possible exception of characterization of precision coaxial calibration standards for vector network analysis, as previously cited. The TM mode is so well suppressed that for almost all practical measurements and applications the errors due to this assumption are of no consequence.

The conductive planar disk with the coaxial port, on the left in Illustration 3, is called a “launcher” and serves to couple energy from the coaxial stimulus into the TM wave propagating along the central conductor.

From the plot it can be seen that close to the excitation port, the E-fields extend from the central conductor to the launcher and are normal to the surfaces of each conductor immediately adjacent to the conductor. Perfect conductivity forces tangential components of the electric field to be zero and only a field component at right angle to the conductor surface is possible. In this region near the port, real current flows in the plane and returns by way of the outer conductor of the input coax port. Further to the right, away from the launcher, close examination of the illustration will reveal that E-field lines terminate along the conductor. Here also they leave the conductor normal to it's surface but curve in the enclosing (vacuum) dielectric and return at a different location along the same conductor, up to one half wavelength away. In this region the resulting wave is TM. In essence, the launcher serves as a transition between the predominantly TEM mode in the coax and the predominantly TM mode on the conductor in the region far from the launcher.

The field solution to the wave equation for coax shows that the peak magnitude for the longitudinal E-field is displaced from the peak magnitude for the radial field by one quarter wavelength. The peak longitudinal fields occur at the locations of voltage minima on the central conductor. The phase of the excitation in Illustration 3 has placed the voltage maximum at or near the input port. Careful examination of the field lines will show that the first clearly discernible maximum of the longitudinal E-fields occurs slightly to the left of the center of the central conductor and approximately three quarter wavelengths away from the maximum occurring near the excitation port. The first longitudinal maximum occurs one quarter wavelength from the port but is difficult to discern because of the other field lines returning to the launcher.

E-field magnitude

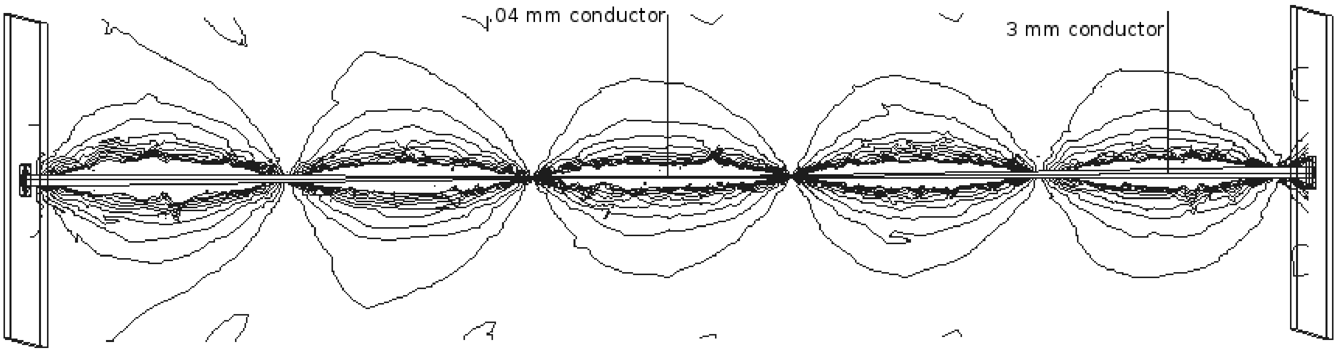

Illustration 4: Contours of constant E-field magnitude

Although Illustration 3 gives insight into E-field direction, it gives almost no information about E-field amplitude or even relative magnitude. To help provide this, contours of constant E-field magnitude for a different modeled two-port system are shown in Illustration 4. These lines are contours of constant magnitude so Illustrations 3 and 4 must be taken together in order to visualize the complete E-Field vectors, which contain both amplitude and direction information. The launchers in this illustration are 100 mm square rather than round and the central conductor is 400 mm long, also square but tapered from 4 mm at each end to .04 mm at the center. The stimulus frequency is 1875 MHz where the structure is 2.5 wavelengths long.

It is noteworthy that the radial extent of the E-field is dependent only on line impedance and not on conductor diameter or wavelength. Because displacement current is constant, conductor diameter affects the E-field magnitude at the surface of the conductor but not the contour it follows in the surrounding dielectric medium.

Contrary to previous belief in regard to surface waves on G-Line, a launcher need not be large. Because most of the E-field is quite close to the conductor, both in the TEM region and in the TM region, the majority of the terminating field lines and current also occur quite close to the conductor surface. The field solutions show that the magnitude of the radial component follows a $\frac{1}{r}$ curve and that the majority of the propagated energy is within a few conductor diameters of the center axis.

Longitudinally the E-field is dependent on wavelength since each field line must have a termination point, which can be up to one half wavelength away. Therefore the conductor must be at least a half wavelength long in order to support the TM mode.

The line impedance in the vicinity of the launcher is lower than that of the the line in free space. This is because field lines producing real current in the coaxial (TEM) region are present along with the lines terminating on the conductor in the TM mode.

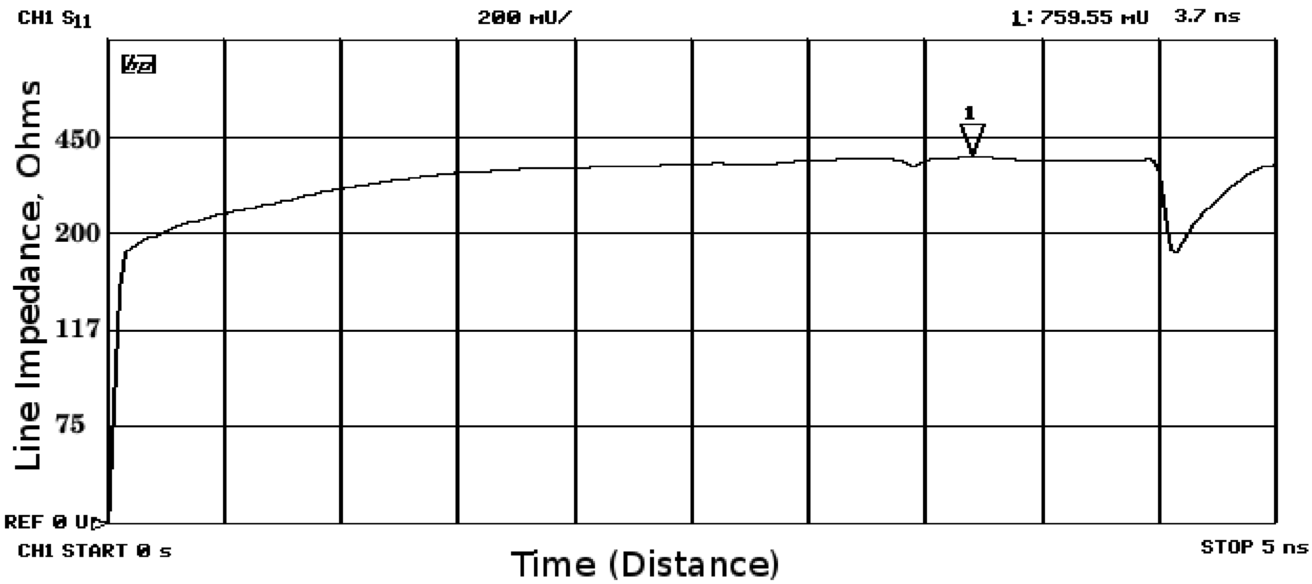

Illustration 5: Time domain measurement of impedance of 680 mm length of TM line stretched between two 68 mm diameter planar launchers.

Illustration 5 shows a VNA time domain measurement of a simple system constructed with a pair of circular brass planar launchers 68 mm in diameter and spaced 680 mm. The conductor is cylindrical, made of .5 mm diameter bare copper conductor (burnished #24 copper wire) and connected between the center pins of SMA connectors each mounted at the center of one of the launchers. The left Y axis has labels for the equivalent line impedance, as calculated from the real part of the reflection coefficient plotted over a range from 0 to 1 when the system reference impedance is fifty ohms.

Within approximately the first centimeter from the excitation port, approximately 20 wire diameters, the impedance rises very rapidly from the initial 50 ohm value at the SMA connector. Beyond that it rises much more slowly and asymptotically approaches the free space value of 377 ohms. The value of the reflection coefficient at the marker corresponds to a line impedance of about 366 ohms. The discontinuity at 4.5 ns is at the location of the second SMA connector.

Practical Launchers

A practical launcher should provide the transition from TEM_{00} to TM_{00} waves as effectively as possible. Generally this transition is between different impedances as well as between different modes.

Any launcher represents a discontinuity to the propagating waves. This discontinuity may produce radiation away from the region. In the TM portion of a system such as is shown in Illustrations 4 and 5, there is complete symmetry of E-field; every field line is one of a pair of lines of equal magnitude but opposite sign. This symmetry is present both axially and longitudinally. Therefore at distances of more than a few wavelengths, these fields add to zero and no net field and no radiation results. However, for the region near a launcher, there is no longer longitudinal symmetry and incomplete cancellation of fields may result at large distances. This produces radiation away from the launcher with the radiated wave linearly polarized parallel to the conductor.

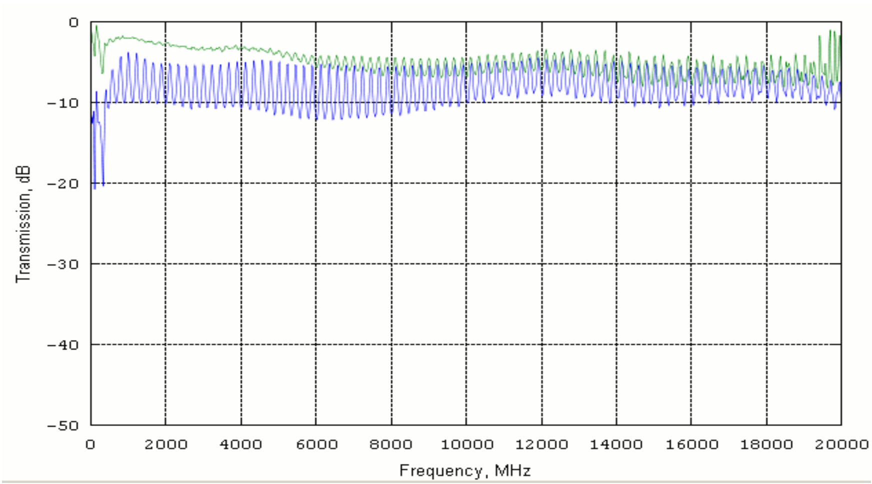

Illustration 6: Measured S_{21} and GA_{max} for the two port TM system of Illustration 5

Illustration 6 shows a frequency domain measurement of the same system with planar launchers that was measured in Illustration 5. The lower trace is of S_{21} which displays the ripple or "beat" between the discontinuities produced by the launchers at each end of the line. In addition to the ripple there is a large amount of mismatch loss between the 50 ohm impedance of the VNA and the impedance presented by the TM system at each port. The upper trace is a calculation of GA_{max} [15] which effectively removes the extra attenuation due to port mismatch and allows just the ohmic and radiation losses to be evaluated. In addition to attenuation due to ohmic losses in the copper conductor, approximately 2 dB loss is apparent near 1 GHz. This is almost entirely radiation loss due to the discontinuities at the launchers and occurs over the entire measurement range. Because of the large standing waves present on the line due to mismatch, the radiation loss is greater than it would be for the situation of a perfectly impedance-matched launcher.

While a planar launcher of the type shown in these illustration is useful for analysis, even with impedance matching added at the ports, it is not generally the best design for minimum system attenuation. The modal discontinuities of this type of launcher generally produces both unwanted radiation and reflection.

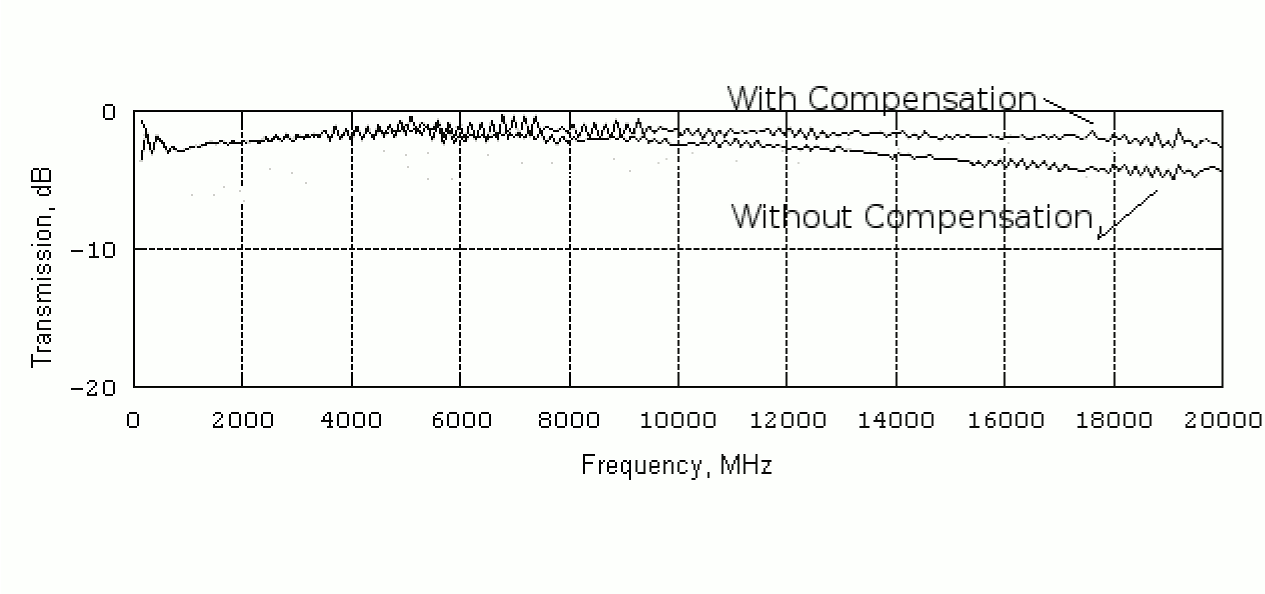

Illustration 7: GA_{max} of "Forward horn" launchers from measurement on 680 mm line, with and without compensation.

Measurement of a system with somewhat better launchers is shown in Illustration 7. These are also 68 mm in diameter but of the forward conical horn rather than the planar type. These were fabricated from a section of a circular brass disk folded and soldered so as to create a ninety degree cone. An SMA bulkhead connector was soldered to the narrow end of the cone and the same type and length of bare copper conductor used for Illustrations 6 and 7 was soldered to the center pin of the connector. Two measurements of GA_{max} are shown; these are with and without a small polyethylene dielectric compensator added to help reduce the discontinuity and consequent reflection and radiation. The compensator was fabricated from an approximately 30 mm long section of polyethylene dielectric removed from conventional RG/8 coaxial cable and placed a few mm away from the SMA connector. Material was removed so as to taper the diameter of the compensator linearly from the wire diameter at each end to a maximum diameter of about 8 mm at its middle. As can be seen by the measurement, this small amount of compensation is only sufficient to make significant improvement above about 5 GHz where the compensator is one half wavelength long.

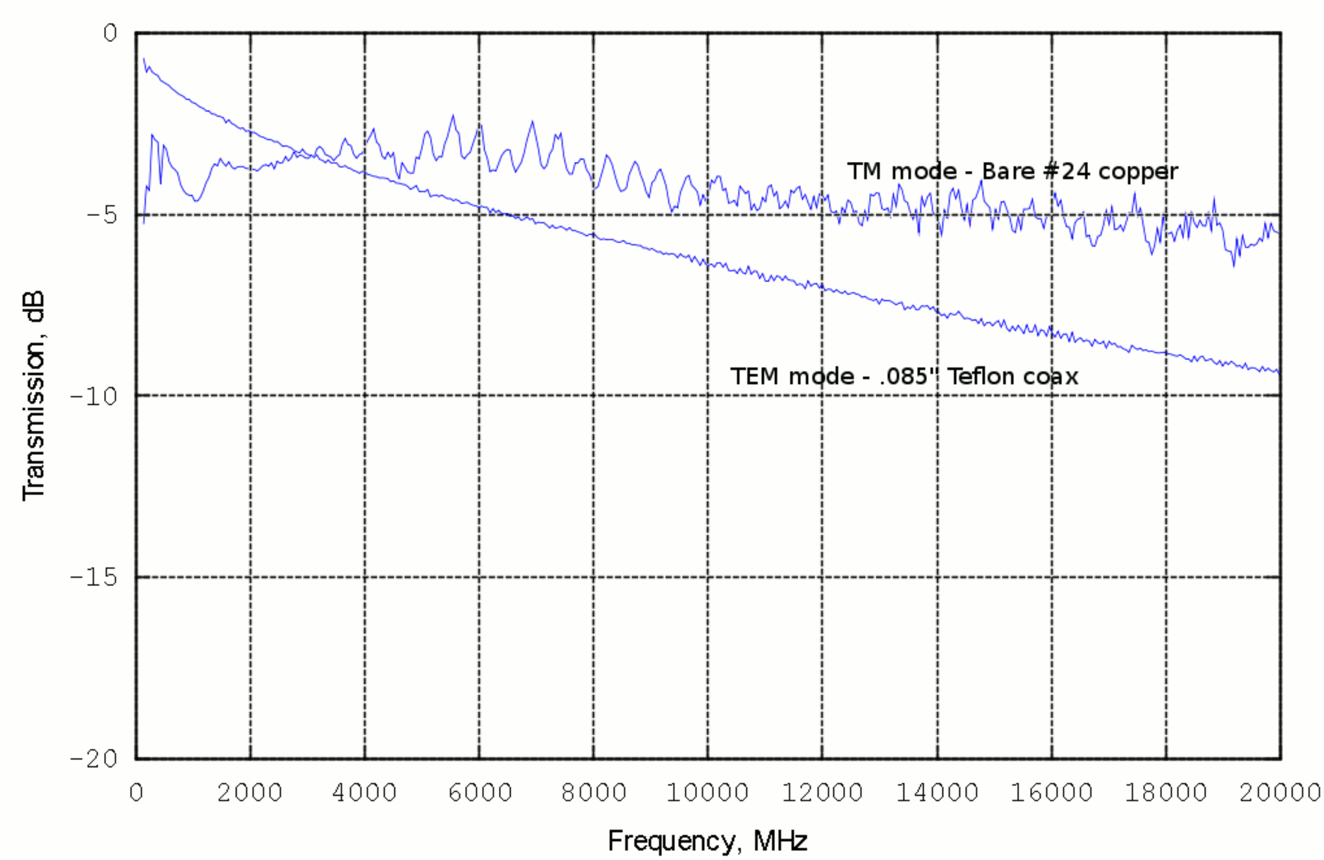

Illustration 8: 140 MHz - 20 GHz measurement of 3.4 meter lengths of .085" Teflon dielectric semi-rigid coax (TEM) and #24 bare copper (TM) mode line.

Illustration 8 shows a measurement of GAmax for the same type of compensated launcher but the line length has been increased to 3.4 meters. Additionally a measurement of an equal length of .085" Teflon dielectric semi-rigid coax has been included. The coax center conductor is of about the same diameter as the conductor of the TM line but is silver plated. In spite of the better conductivity of the coax and the radiation due to the launchers, the lower attenuation of the TM wave system is obvious. A better launcher design can provide even more contrast between the attenuations of the TEM and TM modes. Even with only crude techniques, it is not difficul to reduce total loss for a single launcher to less than .25 dB. These and other launcher possibilities and designs have been described elsewhere[16]. Because the displacement current in TM line is much less than the real current in conventional coax, impedance is higher and the ohmic losses in TM line are dramatically less than for coax. The smaller slope of the TM attenuation versus frequency gives an indication of this superior TM performance.

Broadband

The broadband nature of this transmission system is obvious from these measurements. With even relatively simple launchers it is possible to achieve three or more decades of low-attenuation performance. The lower frequency limit is primarily determined by the diameter of the launcher and by the ability to effectively match to the line impedance. The launcher acts as a sort of “capacitor to space” in that it provides a return path for displacement current. As the launcher gets very small, the reactance of this capacitor increases and gets large compared to load presented by the line-plus-launcher. This higher Q makes broadband impedance matching more difficult. However a 60 cm diameter planar launcher has proven quite usable to below 20 MHz.

The upper frequency limit is affected mainly by the detail of the transition from the coax connector to the line itself. The same 60 cm diameter planar launcher described above easily provides good performance from 20 MHz to 20 GHz, which is the upper limit of the HP8720 VNA used for this measurement. It is very probable that performance was excellent well beyond this.

As the line diameter becomes large compared to a wavelength more care needs to be taken to assure that unwanted discontinuities and resultant radiation do not occur. However it is possible to support the TM mode on lines that are large compared to a wavelength. Work between 30 GHz and 500 GHz indicates that the mode is useful at least that high[17] using conductors having circumferences which are large compared to a wavelength.

Practical Applications

Overhead Power Lines

An obvious and very promising class of applications of this transmission mode is in use of existing overhead electric power lines for last-mile information services. The low attenuation and broadband nature of the mode operating on preexisting infrastructure can provide a basis for very low cost information transmission in much of the populated world. Because the underlying hardware, rights-of-way, support and maintenance for power grids are already in place, the addition of high capacity information transport can be quite inexpensive, particularly when compared to other candidate transmission methods such as DSL, CATV, fixed or mobile wireless systems or fiber optic cable.

The previous practical examples and measurements of TM structures used relatively small conductor diameters. Common power distribution and transmission line conductors range in diameter from about 4 mm up to 25 mm or even 50 mm. Modern power conductors are often constructed by winding multiple bare aluminum or copper wires around a central steel carrier wire which produces a multi-strand cable with extra strength and resistance to stretching. Two or more of these cables are then placed under tension and supported by separate insulators mounted on periodic supports in order to form multi-span segments of overhead power line. In much of the world these supports are wooden “power poles” and may be 10-20 m tall and spaced 30-100 m. It's not uncommon for a single system of poles to provide support for multiple sets of lines, with higher voltage distribution line near the tops of the poles, possibly in conjunction with a step-down transformer, and a second set of supports lower down for lower voltage lines that provide delivery to residential or business end-use sites located adjacent to the line. These lines are prevalent in much of the inhabited world, are located in areas associated with human activity, have systems in place to ensure that they are kept operating and maintained, and as such they are good candidates for last-mile information delivery systems. Because of the capability for very large bandwidth and low attenuation of the TM mode it's useful to examine the characteristics of practical TM mode power line systems. RF and microwave transmission systems using the TM mode that utilize overhead power transmission, distribution or delivery infrastructure have been dubbed “E-Line”.

Illustration 9: A slotted E-Line launcher mounted to an aluminum power line conductor.

An example of a special slotted launcher[18] adapted to mount on an existing power conductor is shown in Illustration 9. This launcher has a special tri-axial adapter section included to allow coupling between coaxial line and the surface wave mode propagating along the aluminum power conductor. The slotted design allows the entire assembly to be placed on the line without requiring any modification of the line conductor. The coaxial port is connected to bi-directional amplifiers, which are solar-powered in this example, located behind the launcher and directly above a mechanical clamp which attaches the entire assembly to the power line conductor. The launcher in this photograph does not include any dielectric compensation to improve the impedance and mode match between the coaxial and TM modes.

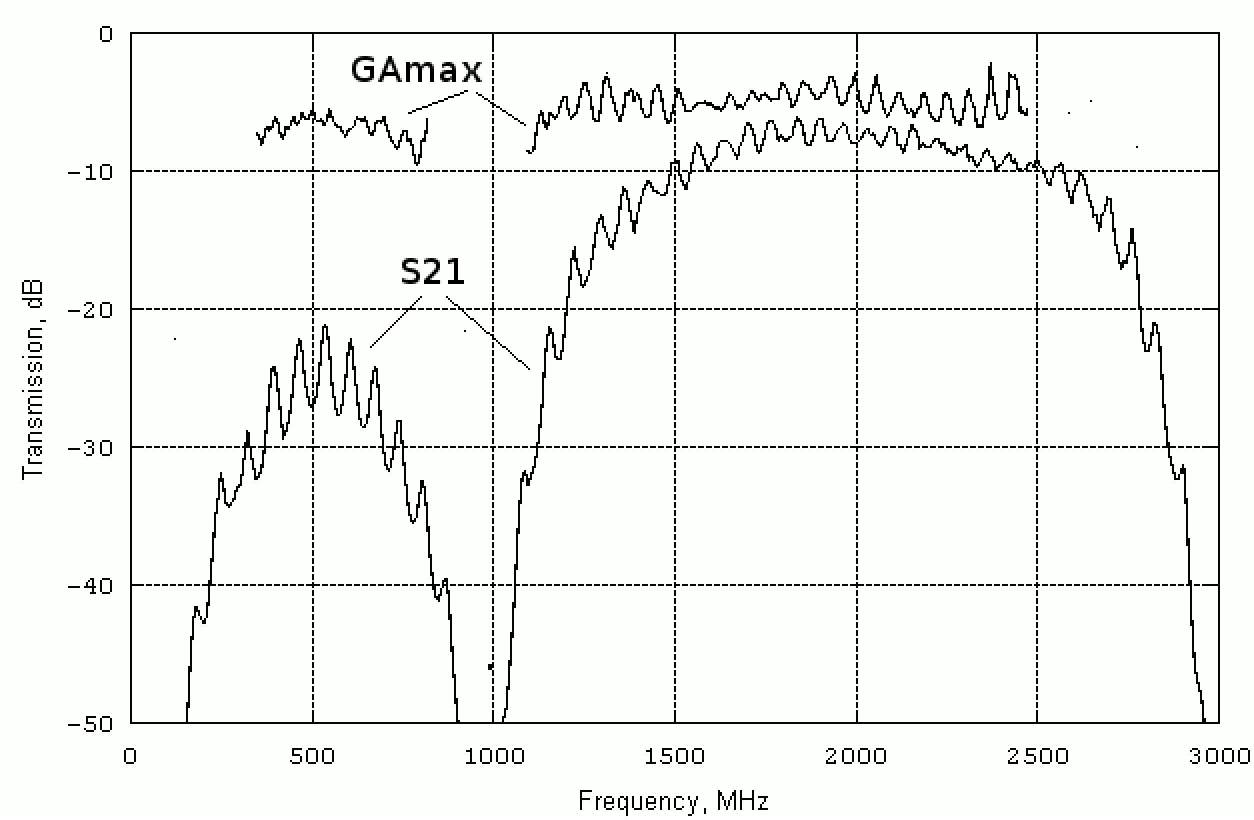

Illustration 10: Measurement of GA_{max} *and* S_{21} *on 18 meters of 4 mm stranded copper power line conductor used in conjunction with the uncompensated launcher shown in Illustration 8.*

A measurement of S_{21} and S_{21} for a pair of launchers of the type shown in Illustration 8 mounted on approximately 18 meters of #4 stranded copper power conductor is shown in Illustration 10. The bandpass nature of the tri-axial coupler is made evident by the transmission response centered at approximately 2 GHz. A second incidental response which is attenuated considerably exists at about 500 MHz. The impedance match of this second response is very poor and results in a great deal of mismatch loss. The degree of this mismatch can be appreciated by comparing the GA_{max} measurements to the S_{21} response. At 1900 MHz, of the 7 dB total system insertion loss shown about 3 dB is due to port mismatch. Approximately another 3 dB is due to radiation loss from modal discontinuities of the uncompensated launchers and the remaining 1 dB loss is due to ohmic losses in the 18 m length of copper conductor.

While these particular launchers are not ideal, their measurement is useful to develop an appreciation of system characteristics. Of course, in power line transmission and distribution systems, other factors contribute to attenuation, reflection and radiation.

Impairment Type	2 GHz	5 GHz	Notes: Standard 7 strand 4ACSR conductor
Line Attenuation	2.2 dB	2.5 dB	Ohmic attenuation per 100'
Saddle Insulators	5 dB	6 dB	Approx. 1 dB variation depending upon “tail” on end of tie
Splices	.5-5 dB	1-5 dB	Finger trap style, larger (step) diameter slightly worse. Quite flat with frequency
Tap Line	3 dB		Function of connection hardware, in particular first with1” from line. Quite flat with frequency.
Rain	-		Too small to measure on 1100' run.
Sag	-		No measurable variation for any practical tension
Bends	0- 20 dB		Saddle insulator $Loss, dB = 0.0192\alpha^2 + 0.017\alpha$, for $0 < \alpha < 25$
Birds	small		Single bird, very large flock may approach -6 dB

Table 1: Impact of impairments common to overhead power lines

Table 1 lists some common impairment factors and their characteristics at 2 and 5 GHz. Insulators normally account for no more than a few dB additional attenuation. Tap lines which connect to a conductor and lead directly away from the line, such as those at a step-down transformer, interrupt the field lines in only one plane and usually cause about 3 dB of extra attenuation. In general, impairments located close to the surface of the conductor tend to have more influence than those even slightly removed. This is to be expected since this is the location of the largest fields. The effects of line bends generally depend a lot on the detail of the conductor and insulator close to the bend itself. A small radius bend is more influential than a slower bend having a larger minimum radius of curvature. Normal line sag has no measurable effect. Most of these impairments have a relatively uniform effect versus frequency and as a result produce rather low group delay perturbation of the transmitted wave.

Because the effects of impairments are generally stable and well-behaved, high Q resonances, sharp frequency domain notches and similar effects are relatively uncommon. As for other types of transmission lines it is possible to configure special structures in a way to create frequency dependent filtering from sections of TM mode line but these kinds of responses aren't common on typical overhead power line installations.

To use overhead power lines for transport of RF and microwave information-bearing signals, a link budget analysis can be made in much the same way as for other wired or wireless systems. To examine the capabilities of E-Line, it is useful to compare the underlying ability to transport signals with other methods, in terms of spectral bandwidth, attenuation and distance.

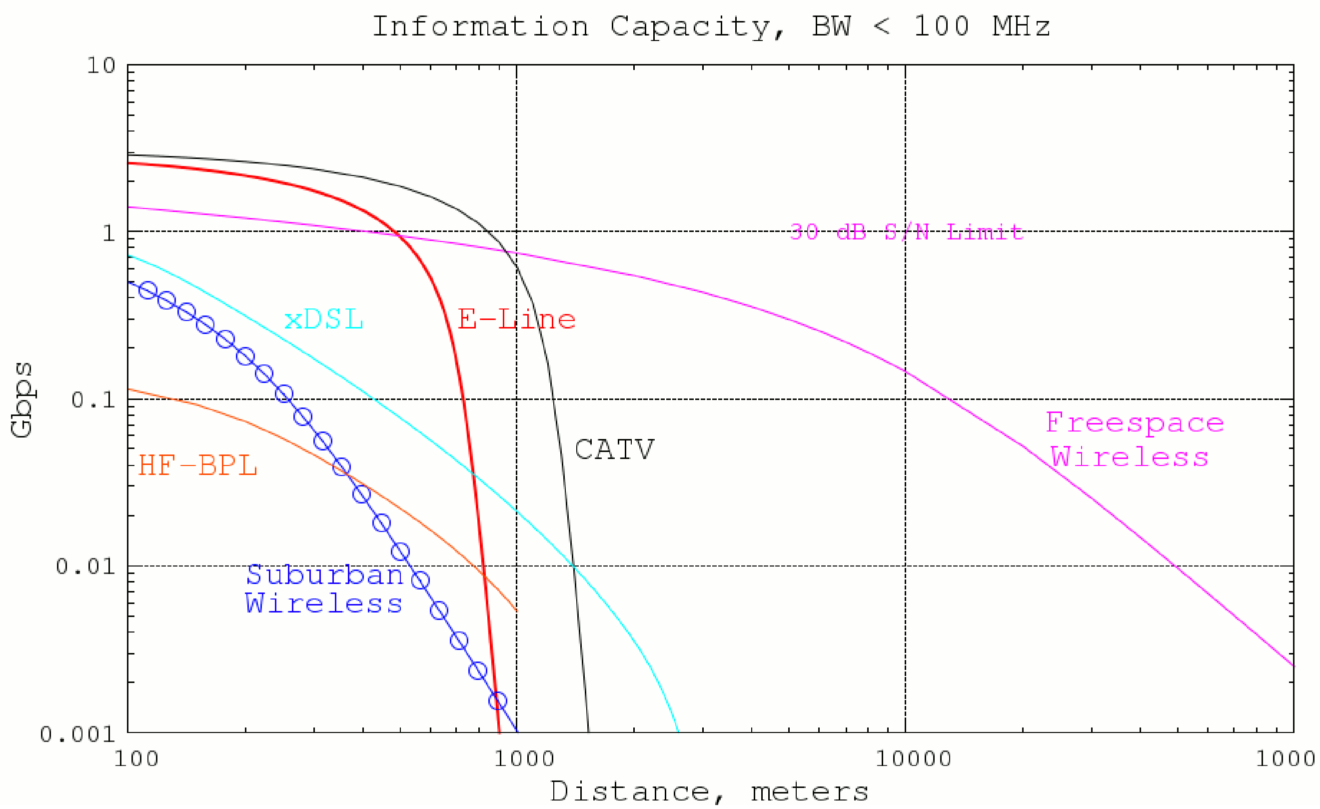

Illustration 11: Maximum information capacities within 100 MHz bandwidth for E-Line compared with other last-mile transport methods.

Transmission Method	Spectral BW, (Center Frequency)	Signal Power dBm (mW)	Noise dBm or Noise Figure (dB)	Attenuation	Notes
HF-BPL	26 MHz (17 MHz)	- 50 dBm/Hz (260)	OPERA[19]		Center Frequency limits available bandwidth
xDSL	100 MHz (50 MHz)	0 (1)	-120	.3 dB/m @ 100 MHz	Crosstalk limited
Suburban Wireless	100 MHz (2 GHz)	0 (1)	(3)	COST231/Hata propagation model	Antenna #1 $1\,m^2$ aperture, 20m elevation
Free space Wireless	100 MHz (2 GHz)	0 (1)	(3)	lossless	Antenna #2 dipole, 2m elevation
CATV	100 MHz (1 GHz)	0 (1)	(3)	Per data sheet	Times Wire LMR600
E-Line	100 MHz (2 GHz)	0 (1)	(3)	Line + Insulator Attenuation	Typical Installation, line loss plus effects of supporting insulator every 100m

Table 2: Conditions and assumptions used to calculate the information capacities in Illustration 11

Illustration 11 plots the maximum theoretical information capacity as a function of distance for several existing last-mile transmission methods along with that for E-Line. The types compared are

- HF-BPL, HF transport using two power line conductors
- xDSL, twisted pair copper telephone lines,
- Free space wireless, radio with completely line-of-sight propagation,
- Suburban wireless, radio within a typical suburban environment,
- CATV, low loss distribution coax
- E-Line, TM propagating mode on single conductor overhead power lines.

A comparison of this sort is almost impossible to perform fairly or completely accurately. Each transport medium has its own characteristics, strengths and weaknesses that make any common benchmark less than perfect. Assumptions necessary for one method are inappropriate or irrelevant for another. Because of these difficulties, Illustration 11 should be considered only a qualitative comparison and is provided to give a sense of relative performance rather than an absolute measure.

This approach calculates information capacity as a function of distance by use of Shannon's equation

$$C = B\log_2(\frac{S}{N}+1) \qquad (25)$$

where

C = maximum channel information rate in bits/second

B = bandwidth in hertz

S = signal power

N = noise power

For each method, the associated spectrum was subdivided into 100 segments and the information capacity for each segment was calculated based on distance, segment center frequency, signal power and noise power. The information capacities of all of these subsegments were then summed to produce an associated maximum capacity. These results describe the maximum information rate possible if a perfect encoding and protocol is used. No allowances or margins for variation have been included. These results are therefore the upper bound rather than a description of practical systems. Unless noted, a source power of 0 dBm (1 milliwatt) and information bandwidth of 100 MHz have been used. Other relevant attributes are as shown in Table 2. Limiting the bandwidth to only 100 MHz considerably understates the capability of E-Line.

In addition to the plots for each of the methods the maximum information capacity possible in 100 MHz bandwidth with C/N ratio limit of 30 dB is shown. This is an arbitrary limit but is similar to the minimum required C/N for protocols such as 802.11a, 802.11g, WiMax, LTE and other common communications standards. If greater spectrum were used, the potential information capacity of E-Line would easily exceed that of every technology,except optical fiber and free space wireless out to distances of several km.

In order to transport information over very large distances all of these methods require periodic amplification in order to overcome signal loss, possibly accompanied by demodulation and remodulation of information. Illustration 11 reveals the maximum distance allowable between such amplification if a specific information rate is to be maintained. For E-Line installed on typical distribution lines with pole spacings of 100 m, amplification every few poles is necessary to maintain the majority of the maximum possible capacity allowed by the assumptions. Line power levels larger than 1 milliwatt can allow increased spacing. Practical systems have been built with five to ten amplifiers per mile of line which have supported more than 2 Gbps information capacity using less than 100 MHz information bandwidth. A photograph of a prototype of one of these amplifying nodes installed on an operating power line is shown in Illustration 12. The launchers are considerably larger than necessary for many applications but allow operation from as low as 200 MHz to above 20 GHz.

Illustration 12: Photograph of one amplifying node in a prototype system installed on an existing medium voltage power line

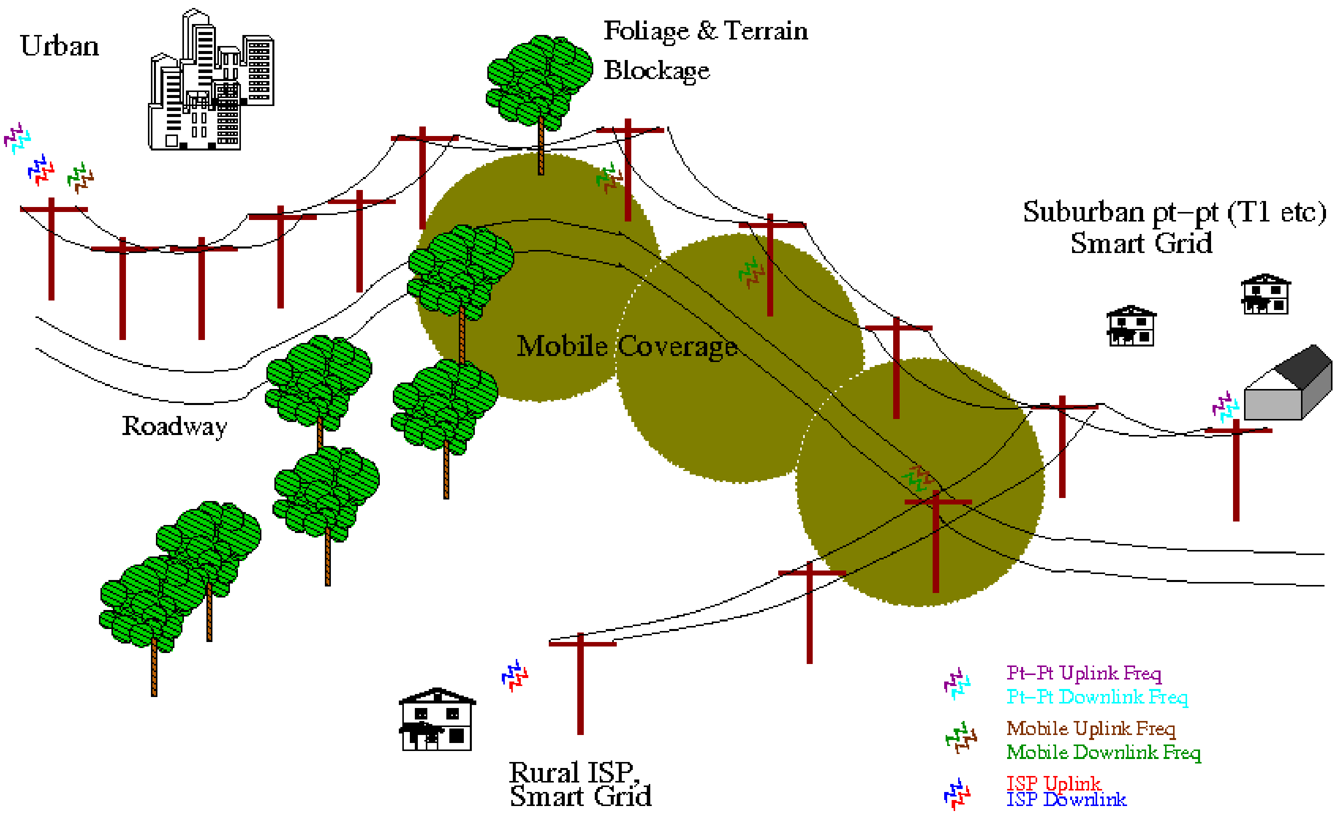

Illustration 13: E-Line (TM mode) system providing simultaneous transport and distribution of different information services.

Illustration 13 depicts an E-Line system capable of providing both high capacity end-to-end information transport as well as information distribution for end users near the line. Simultaneous usage of power poles as sites for "nano-cells" to provide access for adjacent users while enabling near line-of-sight (free space) radio paths allows very high user data rates along with small user antenna aperture and low transmit power. An E-Line distribution system can easily include both an antenna and active circuitry at selected poles in order to tailor a coverage footprint along and in the vicinity of the power line system. In a situation where the communications system is already frequency division duplex, such as in a mobile telephone system, this can be done with simple bi-directional amplification and filtering. In this way a single E-Line installation can provide back-haul (transport), front-haul (distributed antenna feed) and access (distributed antennas) for end users at 3G and 4G speeds, with an aggregate information capacity of many Gbps.

The relatively low attenuation of E-Line allows simple amplification to suffice at each amplifying node, rather than requiring demodulation, remodulation and the attendant delays (latency) produced by these processes. Since all hardware can be located on the line conductor itself and no pole attach is required the cost of this system is dramatically less than is the case for alternate technologies and methods. The result is that simple RF and microwave electronics, periodically included with pairs of launchers placed along a power line, can simultaneously provide and maintain both transport and distribution of high rate information services and content. Since the system follows the electric distribution grid, it can also be used to simultaneously provide “Smart Grid” communications with end-use locations for real time power management and billing.

Feed Line for High Altitude Antennas

The extreme simplicity and relatively small dimensions of a low attenuation and high bandwidth TM mode system make use as an antenna feed line between ground-located communications equipment and high altitude antennas attractive. For many practical terrestrial communications systems, coverage is severely limited by the presence of hills, buildings, foliage and other similar obstructions which are relatively close to the earth. The impact of these impairments can be appreciated by comparing the free space attenuation with that of the suburban environment attenuation of radio signals shown in Illustration 11. Forty to sixty dB of excess attenuation is commonplace for many practical and desirable path lengths. However, by locating at least one antenna well above the impairments, the total radio path loss rapidly falls and allows much higher carrier/noise ratios, performance and coverage. TM mode transmission line can be used to connect heavy ground-located equipment with high-altitude antennas.

To illustrate this application, a lightweight bi-conical antenna was fabricated and integrated with a small forward-horn type launcher. A small helium-filled balloon was used to lift the antenna and the entire assembly was tethered by means of small gauge copper wire which doubled as a lightweight TM feed line for the antenna.

Illustration 14: "Featherweight" bi-conical antenna, integrated TM mode launcher and supporting balloon tethered by feed line.

Illustration 14 is a photograph of the antenna with integrated TM launcher and supporting balloon.

To measure the improvement, the balloon was first allowed to support the antenna at about 2 meters above ground and a reference measurement of a distant commercial100 MHz FM broadcast signal was made. The transmitting antenna for this signal was approximately 150 km away and there was considerable intervening obstruction. As a result, the signal was at or near FM threshold and could not be fully demodulated by a standard FM stereo receiver. The copper wire tether was then allowed to play out and the balloon rose to approximately 60 meters. At that elevation, the antenna was well above local foliage and clutter. The received signal amplitude rose by more than 30 dB. As the feed line and antenna were passive structures, Lorentz reciprocity theorem applies and this antenna and feed line system could be expected to provide the same improvement at the distant location if the balloon supported antenna were used for transmitting rather than for receiving.

Although balloon and kite lifted antennas have been in use for about a century, the lightweight and low cross-sectional area of suitable TM line conductor allows the antenna feed point to be located at high altitude rather than at ground level, as was the case for previous aerially supported antennas which were generally end-fed. Thus, existing heavy communications equipment can remain ground-mounted but be easily used with temporary lightweight antennas located at very considerable elevation and the communications range and quality of common communications systems greatly increased.

An application of this sort might have particular value for emergency communications as well as in situations where temporary wide area communications is required, such as on a battlefield. In addition, because the attenuation of the TM_{00} mode is quite low, RF or microwave energy can be transmitted up to the elevated assembly and rectified to provide DC power for active electronics, signage or even for the lifting device itself. It should be possible, for example, to power an electric helicopter which supports the line and antenna which is simultaneously being used for communications purposes.

Summary

This article has described a previously unknown propagating TM_{00} surface wave mode which exists on a single unshielded conductor. Practical transmission lines utilizing this mode were not previously known to be possible. Descriptions of the associated fields and launchers useful for converting between this mode and conventional transmission lines have been provided and the broadband and low-loss nature of this mode has been illustrated through measurements of simple, practical systems. Some applications of this mode, including the use of the existing worldwide grid of overhead power lines for high rate last-mile information transport have been detailed.

In particular, this discovery allows very inexpensive implementation of wide area information services utilizing the pre-existing worldwide power distribution grid. Simple and inexpensive hardware can be installed on a single conductor of these ubiquitous lines and used to create a high capacity “3rd Pipe” for information distribution. The location and rights-of-way of these existing power systems allow them to be used simultaneously to provide 3G and 4G user access while they also provides back-haul and other point-point information transport. Of particular value, this system can easily be applied for use in Smart-Grid energy systems . The re-use of existing lines, rights-of-way and maintenance systems allow all of these information services to be deployed and operated at a small fraction of the cost of any other method.

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