High-Frequency Electromagnetic Waves on Unshielded Twisted Pairs: Upper Bound on Carrier Frequency

Ergin Dinc¹,*, Syed Sheheryar Bukhari¹, Anas Al Rawi¹,², and Eloy de Lera Acedo¹

¹Dept. of Physics, Cavendish Laboratory, University of Cambridge, CB3 0HE, Cambridge, UK.
²BT labs, Adastral Park, Martlesham Heath, IP5 3RE, UK.
*ed502@cam.ac.uk

ABSTRACT

This paper explores the behaviour of the ubiquitous twisted pairs at high frequencies and wideband excitation of twisted pairs up to 12GHz. Although there is a large quantity of papers on twisted pairs, the papers in the literature mostly focus on the sub-1GHz spectrum, where the current digital subscriber line technologies operate. Higher carrier frequencies on twisted pairs can enable the data rates required by the future communication networks; hence, the existing copper infrastructure can be utilised on the last mile complementing the fibre networks. Towards this objective, we derive analytical expression for the electromagnetic fields and characteristic equation of twisted pairs. With these derivations we show a fundamental limit on the operating frequency of twisted pairs beyond which twisted pairs start to radiate and behave like an antenna. To validate our theory through measurements, we designed a microstrip balun in order to excite the differential mode on the twisted pairs. Unlike off-the-shelf devices, this balun has a nearly linear transmission curve across 1-12GHz. This linearity allows us to detect the radiation which would not have been possible with an off-the-shelf device. At the end, we demonstrate that the standard twisted pairs used in the UK can be used up to 5GHz carrier frequency without any radiation effect and this upper-bound can be moved to higher frequencies by decreasing the twist lengths.

Introduction

Delivering high-speed broadband access necessitates replacing old copper infrastructure with fibre optic cables. However, full fibre broadband for everyone is still not a feasible solution due to its high deployment cost especially in the metropolitan and historical cities. In addition, rewiring of existing buildings and sparsely populated areas are extremely costly and this situation is an important limitation for the fibre deployment. For these reasons, the existing copper infrastructure based on twisted pair (TP) wires will continue complementing the connection between subscribers and the closest fibre-to-the-premises as also discussed by Maes et al.¹. Thus, it is essential to improve achievable data rates over the copper infrastructure so that the bottlenecks over the last mile can be successfully avoided while satisfying the future data demands.

The advancements in the digital signal processing (DSP) have enabled digital subscriber line (DSL) speeds beyond 100Mbps as described by Ginis and Cioffi². In addition to the improvements in DSP, the current efforts mostly target increasing the available bandwidth to boost the capacity over TP. The available technology standard, G.fast³, operates in the frequency spectrum up to 212MHz and can achieve data rates up to 2Gbps. The emerging DSL technology, MGFAST⁴ aims to increase the available bandwidth by further extending the frequency spectrum up to 848MHz and targets data rates up to 10Gbps. As noticed, the available and emerging technologies only consider sub–1GHz frequency spectrum. We believe that the DSL technologies can further boost the achievable rates beyond 10Gbps by utilising carrier frequencies higher than 1GHz. In this way, the existing copper infrastructure can be further used to satisfy the data rates required for the future communication networks without any need of replacing all of the copper infrastructure. Towards this objective, this paper investigates electromagnetic waves on TPs and reports novel results about the upper-bound on the carrier frequency that can be exploited by the DSL technologies.

TP wires were invented by Alexander Graham Bell⁵ in 1881 to reduce radiation from cables, lower crosstalk between pairs and provide robustness against electromagnetic interference. Several research papers have been published demonstrating these benefits offered by TPs. Interference from radio stations (at 100-500KHz) on TP wires was modelled by Stolle⁶. Cross-talk between multiple TP cables in the frequency range below 100MHz was investigated by Paul and McKnight⁷ based on the equivalent-circuit model of TP transmission lines. The references⁸–¹⁰ include experimental and theoretical results on the magnetic fields around TP and conclude that lower twist pitch, i.e., the length of a twist, results in lower magnetic field around the wire. Therefore, TPs with lower pitch length can be placed closer to each other. Yan et al.¹¹ proposed a finite-difference
time-domain (FDTD) algorithm to estimate the coupling of external electromagnetic (EM) waves (1.5-6GHz) to TP cables. In this reference, it is observed that the interference is maximum when the wavelength is half-integer or integer multiple of the twist pitch.

EM modelling of twisted and untwisted pairs has been extensively studied in the literature. However, parallel with the existing DSL technologies, the papers on modelling EM waves on TP mostly focus on sub-1GHz frequency spectrum and use relevant approximations that makes them invalid for higher frequencies. In the 1950s, helically wound conductors became popular for travelling-wave tubes to amplify RF signals thanks to their wide bandwidth levels. As a result, there is a large amount of research on modelling helically wound structures, which has a very close geometry with TP. The very first analytical model for a single helical conductor was presented by Sensiper for a finite width tape around a cylindrical geometry. This geometry was specifically selected such that variables can be separated in the cylindrical coordinate system. The exact modelling of the single tape helix model (including dielectric rods near the helix) was proposed by Chernin et al., where the field equations and dispersion relation were derived. Cross-wound double helices, the closest geometry to TP, was investigated by Chodorow and Chu. However, this reference investigated symmetric mode on the geometry where the current directions are the same on both conductors, whereas the existing DSL technologies use asymmetric mode, where the currents are flowing in opposite directions. In this way, the EM field is concentrated between the wires such that this mode is less affected by the objects in the close proximity and enables placement of multiple wires in a confined space. More details on the symmetric and asymmetric modes on untwisted pairs has been recently investigated by Molnar et al. At the end, the modelling of TPs at higher frequency ranges (>1GHz) stands as a significant open research problem.

We have developed an FDTD simulation in CST Studio Suite in order to validate the proven benefits of TP cables and also observe the propagation of high frequency EM waves on TPs. For this purpose, a cable with varying twist pitch length is designed. The cable starts as an untwisted pair and twisting starts gradually reaching the lowest twist pitch length of 5mm in the middle section. As noticed in 2D E-field results presented in Figure 1(a) for 0.5GHz and 1.5GHz, the fields become more confined around the wire when the twist pitch length is lower as also discussed in several references. More importantly, the TP wire starts radiating after a certain carrier frequency as seen in the E-field results for 7GHz in Figure 1(a) and the $S_{21}$ parameters in Figure 1(b). If TP starts radiating after a certain frequency, it means that there would be an upper-bound on the carrier frequency that can be exploited by the DSL technologies. To the best of our knowledge, this radiation effect has not been reported in the literature and the exact radiation frequency depends on factors such as twist pitch length and separation between wires. Therefore, this paper provides in depth-investigation of this effect by providing a theoretical derivation of the radiation frequency with numerical and experimental justifications.

**Figure 1.** Numerical results for a TP with varying twist pitch. The investigated wire starts as an untwisted pair and the twist rate is gradually increasing. Lowest twist pitch length of 5mm is observed close to the middle section. The wire unwinds into an untwisted pair at the end. (a) E-field plots for different frequencies. The fields become more confined for lower twist pitch lengths as in 0.5GHz and 1.5GHz. However, the wire radiates at higher frequencies as in the 7GHz case. (b) $S_{21}$ results for the wire. The effect of radiation can be also seen in the $S_{21}$ results.

This paper also includes another significant contribution by proposing a microstrip balun which has a nearly linear transmission response to excite asymmetric mode on TP. The asymmetric mode is also known as differential mode, i.e., the fundamental transverse electromagnetic mode (TEM) on untwisted and twisted pairs. Most of the state-of-the-art baluns used in the DSL networks are based on classical transformers and generally operate on sub-1GHz frequencies. This type of baluns provides impedance transformation from 50\Ohm input impedance to 100\Ohm differential impedance. However, especially at higher frequencies, the classical transformer type baluns do not provide an efficient impedance matching as investigated by Schaich et al. In this reference, the authors improved the transmission loss up to 10dB by adding a matching network based on lumped circuit elements. A differential mode launcher based on lumped circuit elements is not an option for us because classical lumped circuit elements have frequency-dependent characteristics, which will make wideband impedance matching impossible. The second option is using a monolithic microwave integrated circuit (MMIC)-type balun. MMIC-type circuit components can
efficiently operate up to 300GHz. These types of baluns are widely available on the market at a higher cost than the classical baluns. MMIC-type baluns also provide 100Ω differential output impedance, which is required to be matched to the wire as highlighted in\textsuperscript{17}. For our experimental setup, the transmission curve of the balun should not have any large oscillations across the frequency band of interest. If there are large oscillations, it is likely to miss any radiation effect having similar level of loss with the oscillation magnitude. As will be discussed in the launcher design section, MMIC-type baluns are not suitable for our experimental setup due to large oscillations on the transmission band. That’s why, we also design a microstrip balun that can provide wideband excitation of TPs up to 12GHz and achieve nearly linear transmission curve. The microstrip balun is referred to as the differential mode launcher in the rest of the paper. Therefore, our theoretical model together with the novel differential mode launcher will enable unlocking high carrier frequencies up to 5-10GHz, which can provide multiplicative increase in the achievable DSL capacities.

The remainder of the paper is organised as follows. First, we provide the theoretical explanation for the radiation. The derived radiation frequency is also compared with the numerical FDTD results to justify the developed model at the end of the first section. Then, the design guidelines for the differential mode launcher is provided. The next section after the launcher design includes the comparison of experimental and numerical FDTD results. Lastly, a detailed discussion on our findings and its importance are included in the conclusion.

**Derivation of Electromagnetic Fields and Characteristic Equation**

For the derivation of EM fields and the characteristic equation for unshielded TP, we followed the methodology presented in\textsuperscript{19}. This is due to the fact that, even though the cable presented in\textsuperscript{19} is shielded and geometrically substantially different, the modelling technique in\textsuperscript{19} has been verified with numerical simulations and experimental measurements. For this reason, we decided to utilise a similar methodology for the theoretical derivations. Figure 2 presents the TP geometry used in the derivations. The separation between Conductor 1 (C\textsubscript{1}) and Conductor 2 (C\textsubscript{2}) is 2r\textsubscript{c}. Twist pitch length, i.e., the full 360\degree turn, is denoted as p and the twist wavenumber k\textsubscript{p} = 2\times\pi/p. Twist angle is found as \cot(\psi) = k\textsubscript{p}r\textsubscript{c}, where \psi = \pi/2 is associated with the untwisted pair as p → ∞. In addition, the conductors are assumed as infinitesimal current filaments without a dielectric coating in the theoretical calculations. This assumption helps with the tractability of the derivations whilst still producing accurate calculations, which are justified with the numerical simulations.

![Figure 2. TP geometry.](image)

**Derivation of Fields in the Twisted Pair**

The fields around TP need to satisfy the source-free Maxwell equations

\[
\begin{align*}
\nabla \times \mathbf{E} &= i\omega \mu \mathbf{H}, \\
\nabla \times \mathbf{H} &= (\sigma - i\omega \varepsilon)\mathbf{E}.
\end{align*}
\]

where the space between the conductors are filled with a medium with conductivity \(\sigma\), permittivity \(\varepsilon\) and permeability \(\mu\). \(\omega\) is the angular frequency.

The cylindrical coordinate system is used for modelling TP. Therefore, the Helmholtz equation for the axial component of the electric field (\(\hat{z}\) axis) can be derived as\textsuperscript{19}

\[
\nabla^2 E_z + k^2 E_z = 0,
\]

where \(k = \sqrt{i\omega \varepsilon / \mu}\) is the wavenumber of the surrounding medium. According to the Floquet’s theorem, the axial component of the periodic structure can be expressed as\textsuperscript{19}

\[
E_z(r) = \sum_{m=-\infty}^{\infty} e_{zm}(r)e^{-im(\phi-k\rho c)\mu \hat{ho}z},
\]

where \((r,\phi,z)\) represent the cylindrical axis system.
The axial electric field component can be further simplified by substituting (4) into (3)

\[
\frac{1}{r} \frac{d}{dr} \left( r e_{\text{ex}} \right) - \left( \frac{\gamma^2_{\text{ex}} m^2}{r^2} \right) e_{\text{ex}} = 0.
\] (5)

The lateral wavenumber is \( \gamma_{\text{ex}} = \sqrt{(h_0 + mk_p)^2 - k^2} \), where \( h_0 \) is the unknown parameter that will be calculated with the characteristic equation. (5) obeys the Bessel equation and solutions to the axial field components can be represented with the modified Bessel functions. The axial fields at the centre \( (r = 0) \) should be finite; thus, the fields for the interior \( (r < r_c) \) can be expressed with the modified Bessel function of the first kind

\[
e_{\text{ex}}^{\text{in}}(r) = E_m^{\text{in}} I_m(\gamma_{\text{in}} r), \tag{6}
\]

\[
h_{\text{ex}}^{\text{in}}(r) = H_m^{\text{in}} I_m(\gamma_{\text{in}} r). \tag{7}
\]

The fields at the exterior region \( (r > r_c) \) are vanishing at infinity, so they can be represented as the modified Bessel function of the second kind

\[
e_{\text{ex}}^{\text{ex}}(r) = E_m^{\text{ex}} K_m(\gamma_{\text{ex}} r), \tag{8}
\]

\[
h_{\text{ex}}^{\text{ex}}(r) = H_m^{\text{ex}} K_m(\gamma_{\text{ex}} r). \tag{9}
\]

The current density on the wires in the Floquet form is written as

\[
K(\phi, z) = \frac{\cos(\psi) \hat{\phi} + \sin(\psi) \hat{z}}{2\pi r_c} \sum_{m=\infty}^{\infty} (i_1 + (-1)^m i_2) e^{-im(\phi - \psi)} e^{i\omega z}, \tag{10}
\]

where \( i_1 \) and \( i_2 \) are the current flowing through filaments as presented in Figure 2.

The relationship between the coefficients of the interior and exterior of the fields can be derived by using the following boundary conditions

\[
\hat{r} \times (E^{\text{ex}} - E^{\text{in}}) = 0, \tag{11}
\]

\[
\hat{r} \times (H^{\text{ex}} - H^{\text{in}}) = K, \tag{12}
\]

which are valid on the TP filaments at \( r = r_c \). In addition, the derivative of the tangential magnetic field is continuous on the boundary.

By applying the boundary conditions, the expressions can be simplified as

\[
E_m^{\text{in}} = E_m^{\text{ex}} \frac{K_m(\gamma_{\text{in}} r_c)}{I_m(\gamma_{\text{in}} r_c)}, \tag{13}
\]

\[
H_m^{\text{in}} = H_m^{\text{ex}} \frac{K_m'(\gamma_{\text{in}} r_c)}{I_m'(\gamma_{\text{in}} r_c)}, \tag{14}
\]

where \( (') \) shows the derivative of the modified Bessel functions.

By applying the boundary conditions and using (10), (13) and (14), the field coefficients for the exterior region can be derived as

\[
E_m^{\text{ex}} = I_m(\gamma_{\text{ex}}) \frac{m(h_0 + mk_p) \cos(\psi) + \gamma_{\text{ex}}^2 r_c \sin(\psi)}{i 2\pi r_c (\sigma - \omega e)} \sum_{m=\infty}^{\infty} (i_1 + (-1)^m i_2), \tag{15}
\]

\[
H_m^{\text{ex}} = I_m'(\gamma_{\text{ex}}) \frac{\gamma_{\text{ex}} \cos(\psi)}{2\pi} \sum_{m=\infty}^{\infty} (i_1 + (-1)^m i_2). \tag{16}
\]

The other components of the fields can be calculated by using the Hertz vectors \( (Z_E, Z_H) \). The Hertz vectors satisfy the Helmholtz equation and can be represented as

\[
Z_E^m = -\sum_{m=\infty}^{\infty} \frac{E_m^m}{\gamma_m^2}, \tag{17}
\]

\[
Z_H^m = -\sum_{m=\infty}^{\infty} \frac{H_m^m}{\gamma_m^2}. \tag{18}
\]
At the end, the remaining field components can be found in terms of the Hertz vectors as

\[
\begin{align*}
E_r &= \frac{\partial^2 Z_E}{\partial r \partial z} - \frac{1}{cr} \frac{\partial^2 Z_H}{\partial \phi \partial t}, \\
E_\phi &= \frac{1}{r} \frac{\partial^2 Z_E}{\partial \phi \partial r} + \frac{1}{c} \frac{\partial^2 Z_H}{\partial r \partial t}, \\
H_r &= \frac{\partial^2 Z_M}{\partial r \partial z} + \frac{1}{cr} \frac{\partial^2 Z_E}{\partial \phi \partial t}, \\
H_\phi &= \frac{1}{r} \frac{\partial^2 Z_M}{\partial \phi \partial z} - \frac{1}{c} \frac{\partial^2 Z_E}{\partial r \partial t}.
\end{align*}
\]

Characteristic Equation

The tangential component of the electric field is required to be vanishing on the TP filaments \((r = r_c)\); thus, we can write the following

\[
(\cos(\psi)\dot{\phi} + \sin(\psi)\ddot{\xi})E = 0.
\]

By using the field equations for the exterior of the wire at \(r = r_c\), the characteristic equation can be derived as

\[
\sum_{m=-\infty}^{\infty} \left[ \frac{1}{(-1)^m} \frac{(-1)^m}{1} \right] [\hat{t}_1 \hat{t}_2] S_m(\omega, h_0) = 0,
\]

where

\[
S_m(\omega, h_0) = (m(h_0 + mk_p) + \gamma_p^m r_c \tan(\psi))^2 + (ki_c \omega \varepsilon_0 r_c) \frac{K_m(\gamma_m r_c)}{I_m(\gamma_m r_c)} K_m(\gamma_m r_c) I_m(\gamma_m r_c).
\]

In the differential mode, the currents on TP flow in opposite directions such that \(i_1 = -i_2\). Therefore, the current density can be nonzero for only odd multiples of \(m\). The even multiples of \(m\) are associated with the symmetrical mode, which is also known as the surface wave mode.

Poynting Vector

In order to predict the radiation from the cable, we can check the radial component of the Poynting vector. If there is a power flow in the radial direction, it means the wire is behaving like an antenna such that some of the power will be leaking away from TP. For this purpose, the radial component of the Poynting vector is calculated as

\[
\langle S_r \rangle = \frac{1}{\mu} Re\left[\frac{1}{(E \times H') \cdot \hat{r}}\right] = \frac{1}{\mu} Re\left[\frac{1}{E_\phi B_z^* - E_z B_\phi^*}\right].
\]

where \(Re(\cdot)\) returns the real part of \(x\).

By substituting (8) and (9) into (17), (18), (20) and (22), the radial component of the Poynting vector can be calculated as

\[
\langle (S_r) \rangle_m = \frac{1}{\mu} Re\left[\frac{i \omega}{c \gamma_m^2} B_m^x B_m^x K_m^x(\gamma_m r_c) K_m^x(\gamma_m r_c) + \frac{i \omega}{c \gamma_m^2} A_m^x A_m^x K_m(\gamma_m r_c) K_m(\gamma_m r_c)\right].
\]

As noticed \(\langle (S_r) \rangle_m\) will be zero when \(\gamma_m\) is real-valued as the term inside the \(Re(\cdot)\) function will be purely imaginary. However, if \(\gamma_m\) is imaginary, the radial component of the Poynting vector will be nonzero and this causes the TP to radiate. \(\gamma_m = \sqrt{(h_0 + mk_p)^2 - \beta^2}\) can be imaginary for \(m = -1\) beyond a certain frequency range, which is associated with the radiation effect. The lower indexes \((m < -1)\) can also contribute to the radiation, but the magnitude of \(\langle (S_r) \rangle_m\) reduces with the square of \(\gamma_m\). For this reason, the most significant contributor of the radiation is the \(m = -1\) index. To show this relationship, we solved (24) by using a simplifying assumption as in21 such that the fundamental component \(m = 1\) of the differential mode is assumed as the only contributing index. In this way, we can numerically solve the characteristic equation for \(m = 1\) and then calculate \(\gamma_m\) term. This condition can be interpreted with the leaky wave antenna theory22 as well. Even though the fundamental mode \((m = 1)\) is a slow wave, the \(m = -1\) space harmonics is a fast wave with \(h_{-1} = h_0 - k_p < k\), and this causes the system to behave like an antenna. Thus, the EM wave propagating on TP radiates along the wire.
Analytical Results: Upper Bound on Carrier Frequency

Figure 3(a) presents the real and imaginary part of $\gamma^{-1}$ for twist pitch length of 15mm and $r_c = 0.5$mm. As noticed, the imaginary part of the wavenumber becomes nonzero at frequencies higher than 9.1GHz for this wire geometry and this frequency is associated with the radiation effect. The radiation frequency can be also estimated by the periodicity of the geometry as $c/(2 \times p)$. This simple estimation predicts 10 GHz for the twist pitch length of 15mm; however, this estimation is significantly higher than the value calculated with our technique. Therefore, we show that the periodicity of the structure is not enough to estimate the frequency of radiation.

The effects of twist pitch length ($p$) and wire separation radius ($r_c$) are presented in Figure 3(b). The radiation frequency decreases as the twist pitch length increases and the pitch length is the dominant factor, which determines the radiation frequency. Figure 3(b) also suggests that the radiation frequency is higher for wires that are closely spaced. These radiation frequencies are also upper bounds on the carrier frequency for the differential mode. Communication systems are required to utilise the spectrum below the radiation frequency as the transmitted power beyond this upper bound is radiated to the medium rather than guided along the transmission line. In the next subsection, our analytical results are validated through numerical simulations.

![Figure 3](image)

**Figure 3.** a) Analytical results for the lateral wavenumber $\gamma^{-1}$. $h_0$ and $\gamma^{-1}$ are calculated by solving the characteristic equation numerically. The wire starts radiating in the frequency, in which the imaginary part of $\gamma^{-1}$ is non-zero. b) Radiation frequency results for different wire geometries.

Numerical Validation

Table 1 includes the results for different twist pitch lengths and wire separation radius $r_c$. The numerical results are calculated with the FDTD simulations on CST. The radiation frequencies are determined from the $S_{21}$ plots and these frequencies are associated with the start of significant drop on $S_{21}$ levels (Similar simulation results are presented in Figure 6, which is for TPs with dielectric layer). As noticed, the developed model is able to predict very close radiation frequency for each geometry. The gap between theory and numerical results are getting larger for lower pitch lengths. The main reason for this is the infinitesimal filament approximation in the theoretical modelling while the wires of TP in the numerical simulations are 3D objects with 0.25mm radius. The details of the simulation environment and excitation of TPs are discussed in the Methods and the next section. In conclusion, the theoretical derivations have shown the fundamental mechanism behind this effect and the radiation frequency can be estimated with the derived characteristic equation.

![Table 1](image)

**Table 1.** Theoretical and numerical results for the radiation frequency.
Design Guidelines for Differential Mode Launcher and Simulation Environment

Validating the radiation effect observed in the numerical simulations and our theoretical calculations in the previous section requires a differential mode launcher, which is working up to $\approx 12$GHz. In this way, we can investigate the radiation effect at twist pitch lengths of 10mm and higher. The most common telephone wires that are deployed in the UK$^{23}$ have twist length of 25mm complying the BT’s Specification of CW1423, but shorter twist lengths of 10-15mm are commonly used in Category 5-8 cables. That’s why, twist lengths of 10-25mm are investigated in this paper.

We designed a microwave balun as a differential mode launcher as seen in Figure 4(a). The DM launcher starts as a microstrip line matched to 50$\Omega$. Both top and bottom traces are tapered down to provide a smooth impedance transformation from 50$\Omega$ to the impedance of the connected wire. Individual wires of TP are separated from each other in order to solder them to the top and bottom traces of the DM launcher as in the bottom sketch of Figure 4. Thus, the part of the TP, which is close to the launcher, is modelled as a double line in the launcher design.

The fundamental mode on the end of the DM launcher and the double wire are both quasi-TEM and the port impedances are nearly constant across the frequency spectrum. Therefore, matching the impedance of these two ports will give the highest performance. Figure 4(c) includes the port impedance values for the end of the DM launcher and double wire. The port impedances are calculated with the frequency-domain solver of CST$^{16}$. The following parameters are used for the double line: center-to-center separation 2.18mm (This value is higher than the TP’s center-to-center separation distance of 1mm due to additional space for soldering.), conductor radius 0.25mm, dielectric thickness 0.25mm and $\varepsilon_r = 2.7$. The DM launcher is designed on an FR-4 substrate (thickness 1.6mm, $\varepsilon_r = 4.3$ and $\tan\delta = 0.021$). As noticed from the figure, the best match is achieved when the end trace width of the launcher is 0.2mm. However, this trace width is not practical as it was not possible to solder wires with conductor radius of 0.25mm without extending the trace width. For this reason, we fabricate DM launchers with 1.25mm in our experiments. The fabricated launchers have an impedance mismatch, but they are still good enough to observe the radiation effect on TPs and validate our theory. The width and length of the substrate along with the type of the taper is optimised by using the time-domain solver of CST. Substrate width and length are selected as 50mm and 100mm, respectively. The taper is designed as a raised-cosine taper with ($\beta = 0.08$). The substrate length is especially important for the return loss of lower frequencies as the smoothness of the taper depends on the wavelength. The designed launcher is targeted to operate between 1-12GHz to observe the radiation from the TPs having 10-25mm twist pitch lengths. Figure 5(a) includes the photos of the fabricated devices.

At the bottom of Figure 4(b), the block diagram of an end-to-end system without a TP is given. In order to accurately determine the radiation frequency, the back-to-back connected baluns are required to have a nearly linear transmission curve without large oscillations. Otherwise, it is not possible to observe the radiation effect when the radiation loss is on a similar level with the oscillations. This is especially expected for larger twist pitch lengths as will be shown in the next section. For instance, we show a comparison of our launcher with a high-end balun in Figure 4(b). Even though the balun (BALH-0012SSG from Marki Microwave$^{24}$) has a better transmission response at higher frequencies, the sinusoid-like nature of its transmission...
makes it not practical to use to measure radiation due to aforementioned reasons. Figure 4(b) also presents the back-to-back $S_{21}$ for the designed DM launcher on FR-4. Note that the transmission loss of this system is equivalent to $-S_{21}$ in dB. As in Figure 4(b), both the numerical simulation and measurement results suggest that the designed DM launcher on FR-4 has a nearly linear response across 2-12 GHz without any large oscillations. The loss of the proposed DM launcher is higher than the off-the-shelf device, but this is primarily caused by the high dielectric losses introduced by FR-4 substrate, which is preferred due to its easy fabrication, availability and low price. However, an optimised launcher for communication applications can be designed by following the guidelines provided in this section.

![Figure 5](image1.png)

**Figure 5.** a) Photos of fabricated DM launcher, and b) measurement setup.

**Experimental and Numerical Results**

The measurement setup can be seen in Figure 5(b). A custom designed jig is used to stabilise the DM launchers 8-10cm above the bench such that any proximity effect due to the surrounding objects can be avoided. Twist lengths are not perfectly uniform along the cables and show 1-3mm variations. The photo of the different twist lengths used in the measurements are included in Figure 6 and the twist lengths are indicated in the subtitles of each measurement. All measurements are performed for 0.5m cables. The cables have properties of 0.5mm diameter copper conductor covered with a 0.25mm thick cylindrical dielectric. The dielectric properties of the wire are assumed as $\varepsilon_r = 2.7$ and $\tan \delta = 0.01$.

Figure 6 includes the S-parameters for the numerical simulations and the experimental measurements, which are performed with the designed DM launcher. Both $S_{11}$ and $S_{21}$ for different wire pairs are highly consistent with the measurements. $S_{11}$ results show that the EM power is successfully enters the system. This transmitted power cannot reach to the second port beyond the radiation frequency as noticed in the $S_{21}$ results. The same radiation frequency calculated in the numerical simulations are observed in the experimental measurements. There is a small mismatch in the radiation frequency of Wire 2 in Figure 6(b). This is probably caused by the uneven twist length along the wire. In addition, accurate measurement of exact twist lengths becomes harder as the dimensions are small.

The power loss due to radiation effect depends on the twist length as well. Wire 4(p=10mm) experiences more than 30dB decrease in the transmitted power after 11GHz. The additional loss in $S_{21}$ is lower for lower twist rates, but even Wire 1 (p=25mm) has additional 3dB loss over 0.5m. Since the twisted pair behaves like a leaky-wave antenna at these frequencies, the transmission loss in dB ($-S_{21}$dB) linearly scales with the length of the wire. Therefore, we can conclude that any future DSL technology, which will operate at high-carrier frequencies, needs to operate below the calculated radiation frequency. Above this frequency, the transmitted power will be radiated away from the line and cause interference in the communication systems nearby.

**Conclusion**

This paper presents two major contributions towards enabling the exploitation of higher carrier frequencies in the DSL technologies. First, electromagnetic fields and characteristic equations of TPs are derived for a helically wound double infinitesimal current filaments. This approach generated an analytically tractable characteristic equation. The individual components of the electric and magnetic fields are calculated. By using these derivations, we calculated the radial component of the Poynting vector. It is found that $m = -1$ space harmonics of TP has a nonzero radial power flow away from the transmission
Figure 6. Experimental and numerical S-parameters of the end-to-end system for different twist lengths.

The second contribution of this paper is the design of a wideband differential mode launcher that can excite TP in the frequency spectrum of $1 - 12\text{GHz}$. Unlike off-the-shelf devices, the proposed design has a nearly linear transmission curve across the frequency spectrum, which makes the detection of any radiation loss possible. The differential mode launcher used in
our experimental setup is designed for validating our theory. Therefore, it was not perfectly optimised for operations requiring low-losses. The complete design guideline of the differential mode launcher is provided in the paper. In conclusion, we believe that our results and design guidelines will help scientists and engineers to better understand the wave propagation on TPs and enable them to design wideband communication systems operating at higher carrier frequencies.

Methods

Launcher Prototyping and Measurements

DM launchers were fabricated with 2D milling on FR-4 substrate (substrate thickness 1.6mm and copper thickness 35µm on both sides). The twisted pair used in the measurements is mostly known as jumper wire complying the BT’s Specification of CW1423. The wire has a single-strand copper conductor of 0.5mm diameter and dielectric thickness of 0.25mm. Twist pitch length of the wire is in the range of 25-28mm (Wire 4 in Figure 6). The wires with different twist pitches (Wire 1-3) are produced by increasing the number of twists of the cable. We measured the length of all twists along the wire in order to determine the twist pitch length ranges that are stated in Figure 6. All wires used in the measurements are 0.5m. S-parameter measurements were performed with a Vector Network Analyser (VNA - 8722D from Agilent Technologies). The VNA was calibrated in the range of 50MHz-20GHz with the calibration kit of 85052D.

Numerical Simulations

For the design and characterisation of DM launchers, CST Studio Suite is utilised. CST Studio Suite has both time-domain and frequency-domain solvers exploiting FDTD and Finite Element Method (FEM), respectively. CST’s time-domain solver is especially powerful in analysing broadband response of the end-to-end system and we utilise the time domain solver to calculate the numerical S-parameters. In the simulations, TP is excited by using the designed DM launchers. The frequency-domain solver of CST is much faster in calculating the modes on 2D cross section; thus, we mostly used the frequency domain simulations to determine the width of the traces on the DM launcher design. In the simulations, the following material parameters are used: Copper for all conductors (5.8 × 10^7 S/m), FR-4 substrate for DM launcher (ε_r = 4.3, tan δ = 0.021), and dielectric around the wire (ε_r = 2.7, tan δ = 0.01).

References

1. Maes, J., Strobel, R., Rawi, A. A. & Ben-Ghorbel, M. High-speed copper and coaxial broadband. *IEEE Commun. Mag.* 57, 12–12 (2019).
2. Ginis, G. & Cioffi, J. M. Vectored transmission for digital subscriber line systems. *IEEE J. on Sel. Areas Commun.* 20, 1085–1104 (2002).
3. ITU-T G, I. 9701 fast access to subscriber terminals (G.Fast)—physical layer specification. *Recomm. ITU-T G 9701* (2014).
4. Oksman, V. *et al.* MGFAST: A new generation of copper broadband access. *IEEE Commun. Mag.* 57, 14–21 (2019).
5. Bell, A. G. Telephone-circuit (1881). US Patent 244,426.
6. Stolle, R. Electromagnetic coupling of twisted pair cables. *IEEE J. on Sel. Areas Commun.* 20, 883–892 (2002).
7. Paul, C. R. & McKnight, J. W. Prediction of crosstalk involving twisted pairs of wires-part I: A transmission-line model for twisted-wire pairs. *IEEE Transactions on Electromagn. Compat.* EMC-21, 92–105 (1979).
8. Moser, J. R. & Spencer, R. F. Predicting the magnetic fields from a twisted-pair cable. *IEEE Transactions on Electromagn. Compat.* EMC-10, 324–329 (1968).
9. Alksne, A. Y. Magnetic fields near twisted wires. *IEEE Transactions on Space Electron. Telem.* 10, 154–158 (1964).
10. Shenfeld, S. Magnetic fields of twisted-wire pairs. *IEEE Transactions on Electromagn. Compat.* EMC-11, 164–169 (1969).
11. Yan, Y. *et al.* An fdtd method for the transient terminal response of twisted-wire pairs illuminated by an external electromagnetic field. *IEEE Transactions on Electromagn. Compat.* 60, 435–443 (2017).
12. Sensiper, S. Electromagnetic wave propagation on helical structures (a review and survey of recent progress). *Proc. IRE* 43, 149–161 (1955).
13. Chernin, D., Antonsen, T. M. & Levush, B. Exact treatment of the dispersion and beam interaction impedance of a thin tape helix surrounded by a radially stratified dielectric. *IEEE Transactions on Electron Devices* 46, 1472–1483 (1999).
14. Chodorow, M. & Chu, E. Cross-wound twin helices for traveling-wave tubes. *J. Appl. Phys.* 26, 33–43 (1955).
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