Pushing the wireless data rate to the Internet speed

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Pushing the Wireless Data Rate to the Internet Speed

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ABSTRACT
This paper intends to answer the question how to achieve wireless data rates that can catch up with current Internet speed, from a basic physics point of view. It is shown that the traditional electric circuit theory and design methodology that have been used for generations are unfortunately not adequate for wireless communications in the future. Instead, disruptive approaches, such as six-port modulators for processing of electromagnetic waves and optical pulses, should be employed to push up the wireless data rate above 100 Gb/s. The key variables to consider for high-speed digital communications are bandwidth, modulation order, and signal-to-noise ratio. In principle, it should be possible to achieve a wireless data rate at 100 Gb/s within the frequency spectrum below 20 GHz.

INDEX TERMS
Circuits and systems, communication technology, direct optical-to-wireless data conversion, full wave approach, high modulation order, internet speed, parallel processing, six-port modulator, wireless communication, wireless data rate.

I. INTRODUCTION
The data rate in wired networks has been increasing rapidly after the creation of the Internet infrastructure, enabling high speed communications in a single global network. Today, data speed of above terabits per second (Tbps) is available with fiber-optic data transmission [1]. Another driving force to push up the data rate is the wireless mobile communication technology. For instance, the 5th generation (5G) mobile telephony [2] is aiming at market introduction in 2020 and it has been specified with a data rate at 10 Gbps which is 10 times higher than the LTE/4G mobile communication technology used today.

It should be noted that the wireless data rate of 10 Gbps for 5G is already appearing within industrial research and development projects. In the academic research, wireless data rate at 100 Gbps is already on the agenda. For instance, in [3] the authors have reported that in the frequency band from 250 to 400 GHz, they have been pushing the highest data rate to 50 Gbps with a single channel, and 100 Gbps with a multi-level modulation and a multiband/multicarrier operation. Now the question is: Can we achieve 100 Gbps and above for wireless data transmissions within the frequency spectrum below 20 GHz for instance? If we can avoid using sub-terahertz bands, it would be much more interesting for practical applications in the future.

The objective of this article is to answer the question: How can we achieve high speed wireless data transmission at the physics limit and what are disruptive methods that may be used to implement new solutions in the future, besides photonics-based solutions? The answers are essential since any technological solution can never go beyond what the principles of physics permit. However, we want to maximize the utilization of nature principles in order to develop new technologies and products with the most advanced but least complex solutions.

II. ANALYSIS
We start with an analysis of the fundamental theory governing high speed data transmission and then essential methods to achieve the maximum data rate that physics permits. In addition, we examine a conventional modulator for wireless digital communications. Furthermore, it is pointed out that the traditional circuit theory based on Kirchhoff’s electrical current and voltage laws has a fundamental problem to treat electrical circuitry of high frequency and high speed needed in the future, whereas a unified electromagnetic wave approach for digital modulations should be adopted for high speed data transmission at high frequency.

A. FUNDAMENTAL COMMUNICATION THEORY
We start with the basic theory governing high speed data transmission, given by the Nyquist criterion and Shannon’s
TABLE 1. Methods to increase frequency bandwidth, modulation order and signal-to-noise ratio.

<table>
<thead>
<tr>
<th>Method 1</th>
<th>Method 2</th>
<th>Method 3</th>
</tr>
</thead>
<tbody>
<tr>
<td>Large B</td>
<td>Find ways to increase the bandwidth of modulator/demodulator.</td>
<td>Multi-bands for parallel RF processing.</td>
</tr>
<tr>
<td>Large M</td>
<td>High modulation order.</td>
<td>Increase the linearity of modulator/demodulator.</td>
</tr>
<tr>
<td>Large SNR</td>
<td>Reduce Flicker and Shot noises, i.e., avoid CMOS/BiCMOS circuits for modulator/demodulator.</td>
<td>Reduce signal loss in antenna, filter and modulator/demodulator.</td>
</tr>
</tbody>
</table>

theorem [4]:

\[ D = 2 \log_2 M \]  

-Nyquist criterion (1)

\[ D = \log_2 (1 + \text{SNR}) \]  

-Shannon’s theorem (2)

where D represents data rate or channel capacity, B available frequency bandwidth, M modulation order, e.g., M=64 for 64 Quadrature Amplitude Modulation (64-QAM), and SNR signal to noise ratio. It is seen from (1) and (2) that to achieve a high data rate, there are basically three variables to play with, i.e., increased bandwidth B, increased modulation order M, and reduced noise level of the system for a large signal-to-noise ratio SNR. For instance, if we choose B=7.5 GHz within the entire ultrawideband bandwidth from 3.1 to 10.6 GHz [5] and M=256, D=2\log_2 M=2\times7.5\times8=120 \text{ Gbps}, according to (1). The required minimum signal to noise ratio is SNR_{min} = 48 dB at this data rate according to (2). It is apparent that with a large bandwidth (≥7.5 GHz), a high modulation order (256 QAM [6]) and a large signal-to-noise ratio (≥48 dB), wireless data transfer rates above 100 Gbps are theoretically feasible, within the frequency spectrum below 20 GHz.

In Table 1, we do an analysis based on the three basic variables, i.e., B, M and SNR in equations (1) and (2), and try to find methods to increase them for a large D, that is, high data rate. In order to achieve the maximum data rate derived from (1) and (2), all the three methods listed in Table 1 should be considered. Note that to have a large SNR by reducing noise levels is equally effective as increasing the modulation order M, therefore the traditional CMOS/BiCMOS process technologies are better avoided for achieving a high SNR, for instance, SNR_{min} = 48 dB as required for the above derived data rate 120 Gbps. The reason is that CMOS transistors and p-n junctions generate Flicker and Shot noises and the noise power is proportional to the static current [7]. It may be argued that Flicker noise, referred to as 1/f noise, will be

overwhelmed by the thermal noise above the corner frequency such that it is not significant at high frequency in the gigahertz frequency spectrum. However, in a digital modulator as the one shown in Figure 1, low frequency Flicker noise will be up-converted to the so-called oscillator phase-noise near the carrier frequency, which reduces SNR [8].

![FIGURE 1. Conventional modulator for direct digital modulation.](image)

Table 1 also lists three methods to increase frequency bandwidth. It should be pointed out that the thermal noise power, equal to kT where k is the Boltzmann constant and T the absolute temperature and B the bandwidth [9], is proportional to the chosen bandwidth, which then reduces SNR as well. This property indicates that to use large bandwidth in the sub-terahertz frequency spectrum, e.g., 100-300 GHz, is a tempting but not a simple solution to increase data rate. Other methods such as using multi-bands for parallel RF processing [6] and fully differential design [10], as well as massive MIMO (multiple-inputs and multiple-outputs) [11] should be utilized.

B. CONVENTIONAL MODULATION

As illustrated in Figure 1, a typically used conventional modulator [7] for direct digital modulation consists of two digital-to-analog (DAC) convertors for the In-band (I[n]) and Quadrature-band (Q[n]) baseband data, one local oscillator (LO) to generate the sinusoidal signal, two mixers for modulation of the multi-level I(t) and Q(t-τ) signals with the LO signal having a π/2 phase shift, that is, A_{LO}\cos(\omega t) and A_{LO}\sin(\omega t), respectively. The purpose for the digital clock (CLK) to have the bit delay time τ and for the LO to have a π/2 phase shift is for so-called orthogonal QAM which generates a two-dimensional constellation pattern, for instance, 64 QAM used by the WiFi technology. The power combiner Σ is used to generate the modulated RF signal to the radio front-end. However, there are several problems with this traditional digital modulation technique, when the baseband data rate increases to an extremely high level, for instance, above 10 Gbps. Firstly, it requires extremely high digital processing capacity of the DAC, resulting in unacceptably high power consumption of the circuitry. Secondly, there are other problems when the two mixers are designed with an integrated circuit consisting of transistors and diodes, due to Flicker and Shot noises and switching noises from the DAC. Moreover, there is another fundamental problem associated with conventional RF circuit design as described in the next section.
TABLE 2. Calculation of relationship between frequency, wavelength, and the ratio of propagation delay per centimeter to wave period.

<table>
<thead>
<tr>
<th>Frequency (GHz)</th>
<th>Wavelength in air (cm)</th>
<th>Wave period (ps)</th>
<th>Propagation delay (ps/cm)</th>
<th>Ratio of propagation delay to wave period</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.1</td>
<td>300</td>
<td>10</td>
<td>0.033</td>
<td>0.0033 (0.3%)</td>
</tr>
<tr>
<td>1</td>
<td>30</td>
<td>1</td>
<td>0.033</td>
<td>0.033 (3.3%)</td>
</tr>
<tr>
<td>3</td>
<td>10</td>
<td>0.33</td>
<td>0.033</td>
<td>0.10 (10%)</td>
</tr>
<tr>
<td>10</td>
<td>3</td>
<td>0.10</td>
<td>0.033</td>
<td>0.33 (33%)</td>
</tr>
<tr>
<td>20</td>
<td>1.50</td>
<td>0.05</td>
<td>0.033</td>
<td>0.66 (66%)</td>
</tr>
<tr>
<td>100</td>
<td>0.30</td>
<td>0.010</td>
<td>0.033</td>
<td>3.3 (330%)</td>
</tr>
<tr>
<td>150</td>
<td>0.20</td>
<td>0.0067</td>
<td>0.033</td>
<td>5.0 (500%)</td>
</tr>
</tbody>
</table>

C. MAXWELL’S EQUATIONS VS KIRCHHOFF’S LAWS

Design of integrated circuits with lumped element models is in principle based on Kirchhoff’s current law (KCL) and voltage law (KVL) in electronic circuit theory. However, as shown in the Appendix, KCL and KVL are simplification from Maxwell’s equations under the time-invariant electrostatic or magnetostatic field condition. This is the direct current (dc) condition in the circuit theory, not the alternating current (ac) condition. It is also seen in the Appendix that KCL and KVL for ac signals are only valid when the signal propagation delay time is not considered, that is, under the condition that the signal propagation speed \( v_p = \infty \), which is fundamentally not true since \( v_p = \frac{1}{\epsilon \mu} \) and neither \( \epsilon \) nor \( \mu \) can be zero, where \( \epsilon \) is permittivity and \( \mu \) permeability.

However, at a relatively low ac signal frequency, for instance, below 1 GHz, when the ac signal wavelength is much longer than the circuit size, the phase delay due to limited phase velocity is usually neglected without any major issue.

Thus, one should be aware that KCL and KVL are only approximations when they are used for time-variant analog circuitry. Especially, caution must be made when applying KCL and KVL to radio frequency (RF) circuitry. Similarly, cautions must be made for high speed digital circuitry as well, with respect to its fundamental and harmonic frequencies.

To support the above analyses, quantitative calculations for the relationship between frequency, wavelength in air, and the ratio between the propagation delay per centimeter (assumed circuit size) and the wave period are listed in Table 2. At 0.1 GHz, the wavelength is 300 cm; the ration between the propagation delay per centimeter and the wave period is 0.0033, i.e., 0.33%. Thus, KCL and KVL can be used without significant errors for ac circuitry at 100 MHz. However, at 1 GHz, the ratio is 0.033, i.e., 3.3%, whereas at 3 GHz it is 10% and at 10 GHz it is 33%! Obviously, KCL and KVL cannot be used for circuit design at frequencies above 3 GHz. Moreover, considering common dielectric material on which electronic circuitry is built the issue increases further, since the relative permittivity of the medium surrounding conductors is typically >2 whereas it is equal to 1 in air, resulting in a larger propagation delay as compared to that in air. This is a dilemma. On one hand, engineers want to utilize frequencies above 3 GHz as required by 5G mobile telephony in order to achieve a large enough bandwidth. On the other hand, circuit theory and design methodology used earlier cannot be used in principle.

III. DISRUPTIVE APPROACHES

The above theoretical analyses indicate that we are facing some fundamental problems when traditional methods for signal processing are used for future wireless communications with extremely high data rate above 10 Gbps, requiring broad frequency bands at high frequency. These problems come not only from circuit architecture for digital modulation as the one shown in Figure 1, but also from the basic circuit theory based on Kirchhoff’s electrical current and voltage laws which are simpler than what is required in the future. However, a natural way to solve these problems is to go back to Maxwell’s equations, that is, the electromagnetic wave approach. We need a unified wave approach to do digital modulation and transmission. Note that the so-called RF electronics is a hybrid approach between the circuit theory based on Kirchhoff’s electrical current and voltage laws and the electromagnetic wave theory based on Maxwell’s equations [9].

A. SIX-PORT MODULATOR

During recent years, we have been doing research on the so-called six-port radio architecture for wireless communication purposes [6], [12], [13]. Our recent research results [12] have shown that binary baseband data, either electrical or optical, can be used directly to modulate the sinusoidal wave from the local oscillator to generate high order modulated wideband RF signal, using a six-port modulator without any digital-to-analog conversion. The modulation is realized by wave interferometry of the transverse electromagnetic mode (TEM) waves in a six-port correlator implemented using microstrip lines in the quasi-TEM mode [9]. The basic circuitry and principle of the six-port modulator are shown in Figure 2 and described with the following equations.

\[
b_{RF} = a_{LO}(\Gamma_I + j\Gamma_Q)/4 = A_{LOcos}\cos(\Gamma_I + j\Gamma_Q)/4 \\
\Gamma_I = (\Gamma_3 + \Gamma_4) \quad \text{and} \quad \Gamma_Q = (\Gamma_5 + \Gamma_6)
\]

where \( b_{RF} \) is the modulated and broadband outgoing wave from the RF port to the antenna, \( a_{LO} \) is the sinusoidal incident wave from LO, and \( \Gamma_3 \) to \( \Gamma_6 \) are reflection coefficients from the impedance loads. Modulation occurs when the single frequency LO signal, i.e., \( A_{LOcos} \) in (3), is reflected from the dynamically changing impedance loads through the six-port correlator.

Unlike conventional modulator implementation shown in Figure 1, the six-port modulator shown in Figure 2 performs digital modulation with electromagnetic wave interferometry in the six-port correlator, using both the incident LO wave and the reflected waves from the impedance loads, all in the TEM or quasi-TEM mode using microstrip lines [9]. This is a unique property, since if we push the data rate to the limit that physics permits, we must start from Maxwell’s...
equations treating electromagnetic waves as explained in the above section and in the Appendix. This means that the six-port radio architecture by its nature should be able to provide higher data rate at high frequencies, compared to conventional radio technologies based on RF signal processing using lumped elements integrated in either silicon or III-V semiconductors. Moreover, the problem associated with the extreme requirement on DACs and digital signal processing for extremely high speed data is in principle avoided. For instance, to reduce inter-symbol interference for the modulator shown in Figure 1, digital filters such as raised-cosine filters are used in DACs in order to reduce inter-symbol interference in the mixers [7]. However, in the six-port modulator shown in Figure 2, no filter is needed for the binary baseband data, since the digital modulation is not done by mixers, instead it is done by TEM wave reflection from impedance loads and TEM wave interferometry in the six-port correlator. Sharp binary pulses in the baseband result in least inter-symbol interference, since they change the load impedances from low to high and vice versa, generating the required reflection coefficients from $\Gamma_3$ to $\Gamma_6$.

B. OPTICAL DATA TO RF DIRECT CONVERSION

In [12] we have also pointed out that lights in optical fibers may be utilized directly to modulate the impedance loads of a six-port modulator. This is illustrated in Figure 3 in which $Z_1$ to $Z_3$ are load impedances in a six-port modulator shown in Figure 2. If $Z_1$ to $Z_3$ are photon-sensitive material or devices, the reflection coefficient $\Gamma_R$ will be modulated by the lights in the optical fibers. This means that binary data in optical fibers can be converted directly to the high order modulated RF signal by the six-port modulator shown in Figure 2. The following key characteristics can be observed from Figure 3:

- The only noise sources are thermal noise in the power combiners and possible Shot noise from $Z_1$ to $Z_3$.
- The reflection coefficient $\Gamma_R$ is used as $\Gamma_3$, $\Gamma_4$, $\Gamma_5$ or $\Gamma_6$ in the six-port modulator shown in Figure 2.
- If the photon-sensitive components $Z_1$ to $Z_3$ are fast enough, the data embedded in $\Gamma_R$ may reach the data rate in the optical fibers. Many very low cost photoconductive devices and photodiodes are commercially available [14].

C. PARALLEL PROCESSING

As listed in Table 1, another method to increase the data rate is to do parallel RF processing. Figure 4 shows our previous studies [6], [15], [16] that parallel and full wave processing can be realized from the broadband antenna to six-port modulators and demodulators, via a triplexer with three bandpass filters. Figure 4a is a photo taken from a circular dipole antenna with a balun and an SMA connector [15]. Figure 4b is a photo taken from a triplexer with 3 bandpass filters [16], and Figure 4c is the measured result from the triplexer, showing three separated sub-bands [16]. Figure 4d shows the interface between six-port modulators/demodulators and baseband data, either electrical or optical [6]. In this way, a large RF bandwidth is divided into three sub-bands for parallel processing from the broadband RF signal to the digital baseband data, or vice versa. There are several advantages with this type of parallel processing. Firstly, except the sinusoidal LO generation block, other blocks including the antenna, triplexer and six-port modulators/demodulators are with passive devices, resulting in the minimum noise level. Secondly, the signal to noise ratio is enhanced because of reduced thermal noise in each subband due to smaller bandwidth. Thirdly, each sub-band has better linearity than the entire broadband, leading to higher modulation order. Accordingly, with the architecture shown in Figure 4, a high data rate can be achieved with a large bandwidth $B$, a large modulation order $M$ and a high signal-to-noise ratio SNR. This is a unique property to meet all the basic requirements as indicated by (1) and (2) and analyzed in Table 1.

IV. DISCUSSION

This article has focused on basic principles and methods for wireless communications to reach a data rate above 100 Gbps. We have used a typical single carrier homodyne QAM modulator shown in Figure 1 for analysis.
There are other types of modulators such as heterodyne QAM modulators and multi-carrier OFDM (orthogonal frequency division multiple access) utilizing IFFT (inverse fast Fourier transform). However, from a signal processing point of view, the single carrier homodyne QAM modulator represents the simplest one as compared to OFDM and other types of modulators for digital communications. As our analysis has shown that signal processing using lumped elements in CMOS/BiCMOS integrated circuits had better be avoided for digital modulation at extremely high speed, an OFDM modulator cannot be a much better solution. However, using a six-port homodyne QAM modulator shown in Figure 2 solves the problem, since only lights and guided TEM waves are involved. No digital processing is needed for modulation, resulting in low power consumption and low noise.

In Table 2, we have assumed that the circuit size is one centimeter. One may argue that to integrated the circuitry shown in Figure 1 does not require such a size. However, when considering on-chip interconnects, packaging and off-chip interconnects, the size can be larger than one centimeter. Moreover, the harmonics associated with digital signals are multiply higher than the fundamental clock frequency, causing more problems as compared to the single frequency from a sinusoidal wave.

Regulations for frequency spectrum have not been considered. However, by utilizing multi-band parallel RF processing similar to the one shown in Figure 4, flexible frequency band allocations can be done. Especially, a cable illustrated in Figure 3 can have many optical fibers which can be linked to parallel RF processing in different frequency bands. For example, massive MIMO [11] can be realized when multiple six-port modulators are used together with an antenna array.

Six-port demodulators or receivers are not shown in this article, but further reading can be done in [13] and [17]–[23]. Other work on six-port modulators can be read in [24]–[26].

V. CONCLUSION

The key variables to consider for high speed digital communications are bandwidth, modulation order and signal-to-noise ratio. In principle, it should be possible to achieve a wireless data rate at 100 Gbps within the frequency spectrum below 20 GHz. To realize that, one should use Maxwell’s equations instead of Kirchhoff’s electrical current and voltage laws for designing communication systems. That is, we should use approaches treating electromagnetic waves, instead of electrical current and voltage signals. In this respect, the six-port modulator can be a good candidate for disruptive innovation, especially when baseband data in optical fibers are used directly to do modulation. This is a simple architecture, but it fulfills the requirements for a large bandwidth, high modulator order and high signal-to-noise ratio. Therefore, it has the potential to reach the maximum data rate that physical principles permit.

APPENDIX

The four Maxwell’s equations in differential form [27] under the condition of no magnetic current source are described below.

\[\nabla \times \hat{H} = \hat{J} + \varepsilon \frac{\partial \hat{E}}{\partial t}\]

\[\nabla \cdot \hat{E} = -\mu \frac{\partial \hat{H}}{\partial t}\]

\[\nabla \cdot \hat{E} = \rho\]

\[\nabla \cdot \mu \hat{H} = 0\]

where \(\hat{H}\) is the magnetic field intensity, \(\hat{E}\) the electric field intensity, \(\rho\) the electric charge density, \(\varepsilon\) permittivity of medium, \(\mu\) permeability of medium and \(t\) the time variable.

If \(\varepsilon = \varepsilon_r \varepsilon_0 = 0\) or \(\frac{\partial \hat{E}}{\partial t} = 0\), \(\nabla \times \hat{H} = \hat{J}\) \hspace{1cm} (9)

If \(\mu = \mu_r \mu_0 = 0\) or \(\frac{\partial \hat{H}}{\partial t} = 0\), \(\nabla \cdot \hat{H} = 0\) \hspace{1cm} (10)

where \(\varepsilon_r\) is the relative permittivity (dielectric constant) and \(\varepsilon_0 = 8.85 \times 10^{-12} \text{ F/m}\) the permittivity of free space, whereas \(\mu_r\) is the relative permeability and \(\mu_0 = 4\pi \times 10^{-7} \text{ H/m}\) the permeability of free space. Using (9),

\[\nabla \cdot \hat{J} = \nabla \cdot (\nabla \times \hat{H}) = 0\]

which is equivalent to the following equation in the integral form according to the divergence theorem [27].

\[\int_v \nabla \cdot \hat{J} \cdot dv = \int_s \hat{J} \cdot d\hat{s}\]

\[= \sum_k I_k\]

\[= 0 \text{ at a node} \hspace{1cm} -\text{KCL (12)}\]

where \(I\) is the electrical current.

Similarly, from (10) the following form can be obtained according to the Stokes’ theorem [27]:

\[\int_s (\nabla \times \hat{E}) \cdot d\hat{s} = \int_c \hat{E} \cdot d\hat{h}\]

\[= \sum_k V_k\]

\[= 0 \text{ along a closed loop} \hspace{1cm} -\text{KVL (13)}\]
where $V = \int \mathbf{E} \cdot d\mathbf{i}$ is the electrical voltage between two points.

It is apparent that Kirchhoff’s Current Law (KCL), i.e., equation (12), is a simplification from Maxwell’s equations under the assumption either $\varepsilon = \varepsilon_0 \varepsilon_\infty = 0$ or $\mathbf{E}_0 \cdot d\mathbf{i} = 0$. Similarly, Kirchhoff’s Voltage Law (KVL), i.e., equation (13), is a simplification under the assumption either $\mu = \mu_0 \mu_\infty = 0$ or $\mathbf{H}_0 \cdot d\mathbf{i} = 0$. Note that firstly neither $\varepsilon$ nor $\mu$ is equal to zero in reality, except artificial epsilon-near-zero material [28]. Secondly, $\mathbf{E}_0 \cdot d\mathbf{i} = 0$ or $\mathbf{H}_0 \cdot d\mathbf{i} = 0$ means time-invariant electrostatic or magnetostatic field condition, which is equivalent to the direct current (dc) condition in the circuit theory.

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**REFERENCES**


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**MAGNUS KARLSSON** received the M.Sc., Licentiate of Engineering and Ph.D. degrees from Linköping University, Sweden, in 2002, 2005, and 2008, respectively. In 2003, he was with the Communication Electronics Research Group, Linköping University, where he is currently a Senior Researcher and a Senior Lecturer. His main work involves wideband antenna techniques, wideband transceiver front-ends, and wireless communications.