60 GHz radio: prospects and future directions

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Abstract—This paper addresses the basic issues regarding the design and development of wireless systems that will operate in the 60 GHz band. The 60 GHz band is of much interest since this is the band in which a massive amount of spectral space (5 to 7 GHz) has been allocated for dense wireless local communications.

Keywords— 60 GHz, millimetre wave band

1. INTRODUCTION

There exists an ever increasing supply of, and demand for, broadband multimedia applications calling for an ever increasing capacity of wireless networks. Finally, this will cause a demand for wireless transfer capacity far in excess of what can be accommodated in the currently used bands at 2.4-2.5 and 5.2-5.8 GHz [1]. An obvious solution to this problem is to resort to the 60 GHz band, where bandwidth is abundantly available. In particular, for dense local communications, the 60 GHz band is of special interest because of the specific attenuation characteristic due to atmospheric oxygen of 10 to 15 dB/km. This regime makes the 60 GHz band unsuitable for long-range (> 2 km) communications so that it can be dedicated entirely to short-range (< 1 km) communications. For the small distances to be bridged in an indoor environment (<50 m) the 10 to 15 dB/km attenuation has no significant impact. The specific attenuation in excess of 10 dB/km occurs in a band-width of about 8 GHz centred around 60 GHz. Thus, from physical point of view, there is about 8 GHz bandwidth available for dense wireless local communications. This makes the 60 GHz band of utmost interest for all kinds of short-range wireless communications.

II. REGULATION

In the United States, the Federal Communications Commission (FCC) set aside the 59-64 GHz frequency band for general unlicensed applications [2]. This was the largest contiguous block of radio spectrum ever allocated. FCC rules allow 10 Watts of equivalent isotropic radiated power in this band, which complies with a maximum power density of 9 μW/cm² at 3 meters distance. This means that 20 dBm transmit power would be the legal power limit with an antenna having 20 dBi gain. Commercial power amplifier GaAs MMICs are now available that can produce 16 dBm of transmit power with good linearity. In Japan, there was a new regulation in August 2000 for high speed data communication. The frequency range is 54.25-59 GHz for licensed use with a maximum output power of 100 mW and a minimum antenna gain of 20 dBi and 59 – 66 GHz for unlicensed use with a maximum output power of 10 mW and a maximum antenna gain of 47 dBi. In Europe, frequency is allocated for mobile in general in the 59-66 GHz band. No specific recommendation or decision has been issued yet in this mobile frequency band. However, the 54 – 66 GHz band is considered by CEPT as a main priority issue; in a recommendation document [3] CEPT considers: “the high-frequency re-use achievable in the oxygen-absorption band reduces the requirement for sophisticated frequency planning techniques and offers the possibility of a pan-European deregulated telecommunications environment for various low-power, low cost, short-range applications” and “there is an urgent need to identify and harmonize civil requirements in the frequency range 54 – 66 GHz”. An important statement that has to be made here is that Europe should follow the US and Japan by opening a significant part of this band for unlicensed use, because license free is an important condition for promoting 60 GHz systems towards the market!

III. STANDARDIZATION

Currently, there is only one standard addressing the 60 GHz band and that is the IEEE 802.16 standard for Wireless MAN which covers 10 to 66 GHz [4]. It concerns a last-mile broadband wireless connectivity alternative to fiber-based DSL. In the design of the physical layer, line-of-sight (LOS) propagation was deemed a practical necessity. Therefore, the standard specifies a single carrier interface which is designated “WirelessMan-SC”. This opens the door for the creation of fixed Broadband Wireless Access, which could provide license-free network access support to buildings with speeds that approach those offered by high-speed fiber optic networks, which saves
tremendous initial investments in the deployment of last-mile networking technology.

Recently a paradigm shift from “60 GHz is strictly for LOS operation” to “60 GHz may also be suitable for non-LOS situations under certain circumstances” has gained considerable support. As a first step in the creation of a standard, the IEEE has formed an interest group to explore the use of the 60 GHz band for wireless personal area networks (WPANs), which typically have to support non-LOS operation with a range of 10 meters. The IEEE 802.15.3(TM) Millimetre Wave Interest Group (mmWIG) was formed in July 2003 as part of an effort to develop a millimetre-wave-based alternative physical layer for the IEEE high-rate WPAN standard, IEEE 802.15.3(TM) 2003.

IV. CHOICE OF RF TECHNOLOGY

Cost efficient RF solutions for high data rate transmission at 60 GHz still have to be determined. In this respect, some important choices have to be made which might be crucial for commercial success:

- choice of the 60 GH radio front-end architecture,
- choice of technology in which the radio front-end should be implemented: GaAs, InP, Si, or SiGe.

A. Front-end architecture

With respect to the choice of the architecture of the 60 GHz front-end radio there are, in principle, three options:

- employing subsampling,
- employing direct conversion (i.e., “zero RF”),
- employing superheterodyning.

Employing analog-to-digital conversion (ADC) and digital-to-analog conversion (DAC) directly at the antennas would make the complete RF and IF part obsolete. However, this option can be ruled out immediately because this would require ADC and DAC devices having 60 GHz bandwidth. Low-cost implementation of this in the medium term will be unfeasible. Apart from this approach, the subsampling receiver represents the “ultimate” solution for simple low power down conversion, which essentially consists of a sampling switch, clocked at a much lower frequency, and an A/D converter. The limitations of the subsampling approach, however, demonstrate some of the inherent problems in low-power receiver implementations. In a subsampling receiver, image frequencies exist at integral multiples of the sampling rate and can alias into the band of interest. As a result, careful filtering prior to down-conversion is required. For example, downconversion of an RF signal having a bandwidth of 500 MHz, would require a sample rate of at least 1 GHz, assuming a “brick wall filter”. In practice, the sample rate will have to be much higher –at least 2 GHz- in order to minimise the finite bandwidth effects of the filter. It is questionable whether 2 GHz A/D conversion, with let-us-say 10 bit quantization, will become feasible in the medium term. This might be examined, in addition to the problem that the resulting signal-to-noise ratio of the down-sampled signal will inevitably be poorer than that of an equivalent system employing a mixer for downconversion, due to the noise aliased from the bands between DC and the passband [5].

The advantages of direct conversion are that it is uniquely well suited to monolithic integration, due to the lack of image filtering, and its intrinsically simple architecture, see [6,7]. FSK modulated signals are especially well-suited to direct conversion, due to their low-signal energy at DC. However, the direct conversion receiver has not gained widespread acceptance to date, especially in high performance wireless transceivers, due to its intrinsic sensitivity to DC offset problems, even harmonics of the input signal, and local oscillator leakage problems back to the antenna. The latter problem may be considered as the most serious problem. Offset arises from three sources [8]:

- transistor mismatch in the signal path,
- LO signal leaking to the antenna because of poor reverse isolation through the mixer and RF amplifier, then reflecting off the antenna and self-downconverting to DC through the mixer,
- a large near-channel interferer leaking into the LO part of the mixer, then self-down-converting to DC.

Good circuit design may reduce these effects to a certain extend, but cannot be eliminated completely, particularly not if quadrature phase shift keying (QPSK) or gaussian minimum shift keying is used since the spectra of these schemes exhibit a peak at DC. But when Orthogonal Frequency Division Multiplex (OFDM) is applied there may be a solution and that is to avoid the use of those subcarriers that, after conversion, correspond with, or will be close to, the DC component. This is just an example of a possible solution but there might also be other solutions that exploit the particularities of the 60 GHz physical layer.

As regards the superheterodyning option, let us consider the simple architecture as depicted in Fig. 1. This figure shows a basic 60 GHz RF front-end architecture for application at the portable station (PS) end. Ideally it should be an integrated on-chip solution consisting of a receive branch, a transmit
branch and a frequency generation function. The receive branch consists of the receive antenna, a low noise amplifier (LNA) and a mixer which downconverts to IF. The transmit branch consists of a mixer, a power amplifier (PA) and the transmit antenna. The antennas are (integrated) patch antennas. The mixers are image rejecting mixers. They need not to be IQ mixers. The IF is taken at 5 GHz with the idea that, with appropriate modifications, IEEE 802.11a RF can serve as IF here to allow dual mode operation and interoperability.

As regards the choice of semiconductor technology, the most important performance parameters are the transit frequency $f_T$, i.e. the frequency at which the current gain is one, and the maximum frequency of oscillation $f_{\text{max}}$. It is commonly agreed that for 60 GHz RF and $f_T$ and $f_{\text{max}}$ of at least 120 GHz are required. The cheapest semiconductor technology is based on silicon CMOS. Fig. 2 shows the performance increase of CMOS. It shows that for CMOS the minimum $f_T$ requirement has already been reached but not much more than that. Silicon will always have a natural disadvantage, when compared to its competitors, as regards electron mobility, which frustrates the further increase of $f_T$ and $f_{\text{max}}$.

**B. Front-end technology**

Clearly, the traditional discrete front-end technology based on waveguide cannot be used for our purposes because of weight, volume and cost. The only viable alternative is the use of Microwave Monolithically Integrated Circuits (MMICs). From cost point of view it is mandatory to reduce the number of components on the PCB board as much as possible. This reduction of the number of RF chips to a minimum is also important for minimising losses in chip interconnection which can become easily significant at high frequencies. So, the level of integration should be as high as possible.

The oscillator circuit could be a voltage controlled oscillator (VCO) controlled by an (off-chip) frequency synthesizer. In traditional designs the VCO is mostly implemented off-chip because it takes too much space on the chip for providing sufficient performance. At frequencies as high as 60 GHz it may become, however, feasible to implement the VCO directly on chip because the minimum dimensions to achieve a certain performance become much smaller. The advantage of this approach is a reduction of components that have to be mounted on the PCB and the avoidance of on-chip frequency multiplication circuits, saving space on the chip and saving VCO performance degradation due to phase noise and frequency offset. Important: An on-chip VCO that directly generates a reference frequency close to 60 GHz may have a relatively relaxed performance when compared with the requirements of a VCO that operates on a much lower frequency in combination with a couple of frequency multipliers.

**Figure 1: Simple 60 GHz RF architecture**

![Simple 60 GHz RF architecture](image1)

A strong competitor of silicon is Gallium Arsenide (GaAs) which is known to have the following merits:

- high $f_T$ and $f_{\text{max}}$ (>120 GHz), driven to still higher values by future gigabit applications,
- low noise factor (4 dB),
- excellent power added efficiency (> 60%),
- good linearity,
- commercial 60 GHz front-end MMIC’s available (thus proven technology for 60 GHz radio).

A good impression of the state-of-the-art performance of 60 GHz GaAs-based MMIC chips that are currently commercially available is given by [9] – [11]. From these data we learn that, in principle, an RF front-end can be composed having considerable performance: 16 dBm transmit power, 5 dB noise factor and –100 dBc/Hz @ 100 kHz phase noise. GaAs-based 60 GHz devices such as low-noise amplifiers, high power amplifiers, multipliers and switches can nowadays be ordered in large quantities in die form at prices in the order of 15 € a piece [12].

According to IBM [13] silicon germanium (SiGe) technology can compete with III-V semiconductor performance, while simultaneously maintaining the multitude of advantage of silicon materials:

- applying wafers of defect-free silicon material with large diameter (200 mm),
- low substrate and processing costs,
• high process uniformity,
• high yield (low defect densities),
• high reliability,
• high design robustness,
• high integration level.

To this list we can add a favourable property and that is a better heat removal due to a higher substrate conductivity which may become of crucial importance since we have to cope with considerable heat generation by the power amplifier at a very small spot.

The current generation SiGe technology features $f_T$ and $f_{max}$ values of ~120 GHz. It can be anticipated that $f_T$ and $f_{max}$ will remain to increase up to around 200 – 250 GHz. Increase of $f_T$ and $f_{max}$ is coupled with a decrease in break-down voltage which may limit the peak power output.

Table I lists GaAs and SiGe cost estimates according to [14]. For heterojunction bipolar transistor technology we can conclude that the cost of GaAs is about 4 times the cost of SiGe on a $/mm²$ basis. This conclusion is confirmed by the cost estimation made in [15], which reads: 0.12 $/mm²$ for 6” SiGe, 0.5 $/mm²$ for 4” GaAs and 1.2 $/mm²$ for 3” Indium phosphide (InP).

<table>
<thead>
<tr>
<th>Item</th>
<th>GaAs</th>
<th>SiGe</th>
<th>Units</th>
</tr>
</thead>
<tbody>
<tr>
<td>Feature Size (µm)</td>
<td>0.5</td>
<td>2.0</td>
<td>HBT</td>
</tr>
<tr>
<td>Starting Material</td>
<td>200</td>
<td>600</td>
<td>BiCMOS</td>
</tr>
<tr>
<td>Mask steps</td>
<td>12</td>
<td>14</td>
<td></td>
</tr>
<tr>
<td>Photo cost</td>
<td>1200</td>
<td>1400</td>
<td>$</td>
</tr>
<tr>
<td>Raw cost</td>
<td>1400</td>
<td>2000</td>
<td>$</td>
</tr>
<tr>
<td>Wafer Diameter (mm)</td>
<td>100</td>
<td>100</td>
<td></td>
</tr>
<tr>
<td>Yield</td>
<td>80</td>
<td>70</td>
<td>%</td>
</tr>
<tr>
<td>Cost/mm²</td>
<td>0.22</td>
<td>0.36</td>
<td>$/mm²</td>
</tr>
</tbody>
</table>

In the short term GaAs as well as InP technology will be applied since these are already mature technologies providing excellent performance. Because of the cost advantage, however, it is likely that SiGe technology will become the ultimate solution for low-cost high-volume 60 GHz front-end MMIC chips.

### V. ANTENNAS

60 GHz antennas should feature the following properties:

• low fabrication cost, readily amenable to mass production,
• light weight, low volume,
• high efficiency which implies low-loss feed,
• easily integratable with MMIC RF front-end circuitry,
• covering the 59 – 66 GHz frequency band,
• eventually, circular polarisation.

The requirements of low cost, low weight and low volume rule out many antenna structures: Obviously, the classical microwave aperture antennas of the type “heavy metal” are unsuitable. Also lens antennas cannot be used because a lens is typically an expensive part and is not readily amenable to mass production. The well-known whip antenna with coaxial feed may be considered as too lossy and too expensive, in case it is applied at 60 GHz. A relatively new development is the application of micro-electromechanical systems (MEMS) in antenna structures, i.e., the use of small (chip-level) electro-mechanical parts that can be actuated by supplying a certain actuator voltage. Such antenna structures have the potential to become cheap and small. The current state-of-the-art is, however, that MEMS antennas cannot be used for our purposes, because of the high (25-100 volts) actuator voltage that is required to establish a significant movement. A related problem is that very thin hinges are required to achieve some flexibility of the moving part, which gives rise to ageing. Nevertheless, for the longer term when these problems are solved, MEMS may become of significance. In this respect, it should be noted that MEMS may also become significant for voltage controlled phase shifters to steer adaptive arrays.

For the shorter term, the only viable solution that remains is the use of microstrip antennas. Microstrip patch antennas feature all of the properties listed in the aforementioned list of required properties. Linear polarisations are possible with a straight forward feed structure. Patch antennas can also have circular polarisation. The application of circular polarisation is considered because there are strong indications that channel delay spread is substantially lower in case circular polarisation is used instead of linear polarisation, see [16]. Feed lines and matching networks can be fabricated simultaneously with the antenna structure. Finally, dual-frequency and dual-polarisation antennas can be easily made. However, in general it can be said that microstrip antennas also have some limitations:

a) most microstrip antennas radiate into half-space,
b) low power handling capability (~100 W),
c) narrow bandwidth and associated tolerance problems,
d) radiation from feeds and junctions,

For our application, a) has little significance; An access point antenna should have high gain, whereas a portable station antenna should radiate in horizontal direction or slightly upwards, but not downwards.
Limitation b) is of no significance at all because radiated power will never exceed 100 mW or so. The other limitations can be circumvented by taking suitable choices for, in particular, the dielectric constant $\varepsilon_r$ and the thickness of the substrate. For 60 GHz, a substrate thickness of 100 µm is quite acceptable which is the typical thickness of GaAs and SiGe chip substrate. $\varepsilon_r$ for GaAs and SiGe is about 12 which yield a bandwidth of 2% and a radiation efficiency of 80% [17]. Hence, a 60 GHz patch antenna integrated with the 60 GHz RF front-end might be a good option.

### VI. CHANNEL PROPERTIES

At 60 GHz there is much more free space loss than at 2 or 5 GHz since free space loss increases quadratically with frequency. In principle this higher free space loss can be compensated by the use of antennas with more pattern directivity while maintaining small antenna dimensions. When such antennas are used, however, antenna obstruction, e.g. by a human body, and mispointing may easily cause a substantial drop of received power which may nullify the gain provided by the antennas. This effect is typical for millimetre waves because the diffraction of millimetre waves (i.e., the ability to bend around edges of obstacles) is only weak. As regards blocking effects omnidirectional antennas have an advantage in a reflective (e.g. indoor) environment since there they have the ability to still collect contributions of reflected power at the event of line-of-sight (LOS) obstruction.

Walls may considerably attenuate millimetre waves. The transmissivity strongly depends on material properties and thickness. At 60 GHz transmissivity of glass may range from 3 to 7 dB whereas transmission through a 15 cm thick concrete wall can be as high as 36 dB [18]. We may therefore expect that concrete floors between stocks of a building act as reliable cell boundaries. This helps to create small indoor cells for hot spot communications.

Table II

<table>
<thead>
<tr>
<th>Thickness (cm)</th>
<th>60 GHz</th>
<th>2.5 GHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>drywall</td>
<td>2.5</td>
<td>6.0</td>
</tr>
<tr>
<td>Office whiteboard</td>
<td>1.9</td>
<td>9.6</td>
</tr>
<tr>
<td>Clear glass</td>
<td>0.3</td>
<td>3.6</td>
</tr>
<tr>
<td>Mesh glass</td>
<td>0.3</td>
<td>10.2</td>
</tr>
<tr>
<td>Clutter</td>
<td>-</td>
<td>1.2</td>
</tr>
</tbody>
</table>

A consequence of the confinement to smaller cells is that channel dispersion is smaller when compared with values encountered at lower frequencies because echo paths are shorter on average. Rms delay spread may range from a few to 100 ns. It is expected to be highest if omnidirectional antennas are used in large reflective indoor environments [18,20]. When, instead, high gain antennas are used, rms delay spread may be limited to a few ns only [18,21].

Movements of the portable station as well as movements of objects in the environment cause Doppler effects as frequency shift and spectrum broadening of the received signal. These Doppler effects are relatively severe at 60 GHz because they are proportional with frequency. If persons move at a speed of 1.5 m/s (walking speed) then the Doppler spread that results at 60 GHz is 1200 Hz [18].

### VII. FEASIBLE LINK PERFORMANCE

A consequence of the low wall penetration of millimetre waves is that, in many cases, at least one access point per indoor environment is required. From a coverage point of view the best place for the access point antenna would be somewhere near the centre of the room at a high position near the ceiling. From a network deployment viewpoint, however, the need to mount antennas in the ceiling in each and every room is tiresome and cables would probably have to run over the ceiling, which would be unaesthetical.

An attractive alternative option would be the possibility of placing a small sized access point in each room, with its small sized antenna(s) mounted on a wall where it can be readily connected to the existing LAN cabling that is already installed – just as is the case with today’s WLAN access points.

In order to allow flexible terminal use, the low position of the access point (antenna) necessitates measures to cope with the drop in received power due to line-of-sight (LOS) obstruction by a person or object. One measure is to apply macro diversity by...
switching to another access point as soon as the received signal drops below a certain threshold. However, this requires the use of more than one access point per room which may increase the costs significantly, particularly when many small rooms have to be covered. A more attractive solution may be found in other direction namely that of applying particular antenna patterns, which may be adaptive to some extent, e.g., by applying beam switching. Low- or medium-gain antennas may be preferred to high-gain antennas in order to avoid stringent antenna pointing and tracking requirements. Experimental work on 60 GHz antenna pattern optimisation has been carried out by the Radiocommunication Group at the Eindhoven University of Technology. Measurements have been conducted in many indoor environments. Fig. 3 shows the received power normalised on the transmitted power (NRP) in dB measured in the 58-59 GHz band as function of the separation distance between transmitter and receiver. These measurements have been performed in a room with dimensions 7.2*6*3.1 m³. The sides of the room consist of glass window and smoothly plastered concrete walls whereas the floor is linoleum on concrete. The ceiling consists of aluminium plates and light holders. The transmitting antenna was located in a corner of the room at a height of 2.5 m. This antenna has an antenna gain of 16.5 dBi and produces a fan-beam that is wide in azimuth and narrow in elevation. Its beam was aiming towards the middle of the room. A similar fan-beam antenna was applied at the receiving station which was positioned at various places in the room at 1.4 m above the ground.

The upper solid curve in Fig. 3 shows the NRP in case the beam of the receiving antenna is pointing exactly towards the transmitting antenna. The dotted curve represents the situation in which the fan beam at the receiver has an azimuth pointing deviation of 35°. The lower solid curve represents the situation in which the fan-beam antenna at the receiver is replaced by an antenna that has an antenna gain of 6.5 dBi and that radiates omnidirectional in the horizontal plane. As a reference, the dashed curves are added which represent the respective theoretical results according to the free-space law of Friis, i.e., a 6 dB decrease per doubling of distance. The curvature of both solid NRP curves is typical for indoor situations in which antenna patterns are not well pointed towards each other at short distances. In that area, the NRP increases with distance. This is because the increased free space loss is more than compensated by antenna gain since the antennas are better directed towards each other. If the separation distance is increased further these curves tend to become higher than the free-space curves because the reflections from walls etc. contribute effectively to the received power. The dotted curve remains lower because of the fixed 35° antenna mispointing at all distances.

All curves in Fig. 3 refer to the situation of a LOS path between transmitter and receiver. Fig. 4 shows the curves for non-LOS (NLOS) conditions. On applying the fan-beam antenna the average drop of NRP due to LOS path obstruction is about 11 dB for 0° as well as 35° pointing deviation. With the omnidirectional antenna this drop is about 4 dB. The results in Fig. 3 are representative for other indoor environments in the sense that the free-space law can be considered as a reliable lower bound of NRP at relatively large distances. Hence, we can estimate the feasible link performance on the basis of the free space loss. Let us, for instance, consider the antenna setup as described but in a larger room with an antenna separation distance of 10 m.
the receiver (room temperature = 290 K), \( B \) is noise bandwidth and \( F \) is receiver noise figure. With a receiver noise figure of 10 dB and a noise bandwidth of 500 MHz the received noise power amounts to \(-77 \text{ dBm}\). This yields a signal-to-noise ratio (SNR) of 22 dB. Within 500 MHz bandwidth, a data rate of 500 Mbit/s can be accommodated by using OFDM in combination with QPSK and ¾ rate convolutional coding. For sufficient performance in terms of bit error ratio (\(< 10^{-6}\)) a SNR of about 10 dB is required. This implies that 12 dB margin is left to cope with shadowing and performance degrading factors occurring in the transceiver such as phase noise and frequency shift. As shown by Fig. 3 and Fig. 4, this margin can be improved 16.5-6.5-11+4=3 dB by applying fan beam antennas instead of omnidirectional antennas. In order to avoid cumbersome pointing, a fan beam antenna can be used in the form of a sector antenna as presented in [22].

Now let us compare the SNR performance at 60 GHz with what we may expect at a much lower frequency, say 5 GHz. Since, according to the Friis formula, the free-space path loss is proportional to the square of the frequency, the link budget at 60 GHz is 21 dB less when compared with the linkbudget at 5 GHz under equal conditions (same antenna patterns, separation distances etc.). So, at first sight, there seems to be a substantial disadvantage of 60 GHz transmission. It should be realised, however, that any successful commercial system is essentially limited by co-channel interference, which will also be 21 dB lower at 60 GHz, as far as it concerns free-space loss. As a matter of fact, the signal-to-interference ratio of an interference-limited system is commonly modelled as being independent of the operating frequency, see e.g. [23]. If we also take into account the attenuation due to oxygen absorption as well as the extra severe wall attenuation at 60 GHz, we may even expect better signal-to-interference figures.

### IX. FEASIBLE NETWORK PERFORMANCE

A good figure of merit for network capacity is the obtainable information transfer density per square meter. As already stated in the previous section it should be readily possible to obtain a spectral efficiency of 1 bps/Hz with OFDM as transmission scheme. In that case the availability of 5 GHz spectral space implies an aggregate capacity of 5 Gbps per cell. With a cell radius of 10 m the information density is then 16 Mbps/m². Table III provides a comparison with corresponding figures obtainable with 802.11b, Bluetooth, 802.11a/g, and Ultra Wide Band (UWB) based on similar calculations. It reveals that 60 GHz radio has the potential to provide an information transfer density that is not only far in excess of what can be obtained with current systems but is also more than one order of magnitude higher than what UWB can offer. Further capacity enhancements can be relatively easily obtained by application of Multiple-Input Multiple-Output techniques since, at 60 GHz, antenna dimensions and mutual distances can be relatively small.

<table>
<thead>
<tr>
<th>Capacity (Mbps/m²)</th>
<th>0.001</th>
<th>0.03</th>
<th>0.1</th>
<th>1</th>
<th>16</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of channels</td>
<td>3</td>
<td>10</td>
<td>12</td>
<td>6</td>
<td>10</td>
</tr>
<tr>
<td>Inform. rate per channel (Mbps)</td>
<td>11</td>
<td>1</td>
<td>54</td>
<td>50</td>
<td>500</td>
</tr>
<tr>
<td>Cell radius (m)</td>
<td>100</td>
<td>10</td>
<td>50</td>
<td>10</td>
<td>10</td>
</tr>
<tr>
<td>802.11b</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
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<tr>
<td>Bluetooth</td>
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<tr>
<td>802.11a</td>
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<tr>
<td>UWB</td>
<td></td>
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<tr>
<td>60 GHz</td>
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</table>

### IX. CONCLUSIONS

The principal reason for focussing on the oxygen absorption band around 60 GHz band is the huge amount of allocated bandwidth, which can be used to accommodate all kind of short-range (<1 km) wireless communication. In addition, 60 GHz front-end technology is emerging rapidly. In principle, VCO’s as well as antennas can be implemented so small that they can be integrated with the RF MMIC yielding a considerable cost saving. When compared with the well-known 2.4-2.5 GHz band the channel dispersion appears to be relatively small. In addition, it is confirmed that the frequency reuse distance is relatively small.

In the United States as well as in Japan a large contiguous block of radio spectrum has been allocated for unlicensed use. Up to now Europe did not follow although it is recognized by CEPT that “the high-frequency re-use achievable in the oxygen-absorption band reduces the requirement for sophisticated frequency planning techniques and offers the possibility of a pan-European deregulated telecommunications environment for various low-power, low-cost, short-range applications”. The message herewith posed to CEPT is that it would be wise to follow the US and Japan by opening a large part of the 60 GHz band for unlicensed use. The resulting world-wide license-free band would be a clear incentive for the industry to develop standards and 60 GHz radio technology with the prospective of a new mass market.
REFERENCES


