Millimeter-wave Mobile Broadband: Unleashing 3-300GHz Spectrum

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Outline

- Introduction
  - Mobile broadband growth
  - The myth of traffic and revenue gap
  - The national broadband plan
- mmW spectrum
  - History of millimeter wave communications
  - Unleashing 3-300GHz spectrum
  - LMDS and 70/80/90 GHz bands
- mmW Propagation characteristics
  - Free Space Propagation
  - Material penetration loss
  - Oxygen and water absorption
  - Foliage absorption
  - Rain absorption
  - Diffraction
  - Ground reflection
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  - Stand-alone MMB system
  - MMB base station grid
  - Hybrid MMB + 4G systems
  - Deployment and antenna configuration
- MMB air-interface design
  - Duplex and multiple access schemes
  - Frame Structure
  - Channel coding and modulation
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- Radio frequency components design and challenges
  - RF transceiver architecture
  - MMB RF transceiver requirement
  - mmWave Power amplifier
  - mmWave LNA
- MMB system performance
  - Link budget analysis
  - Link Level performance
  - Geometry distribution
  - System throughput analysis
- Summary

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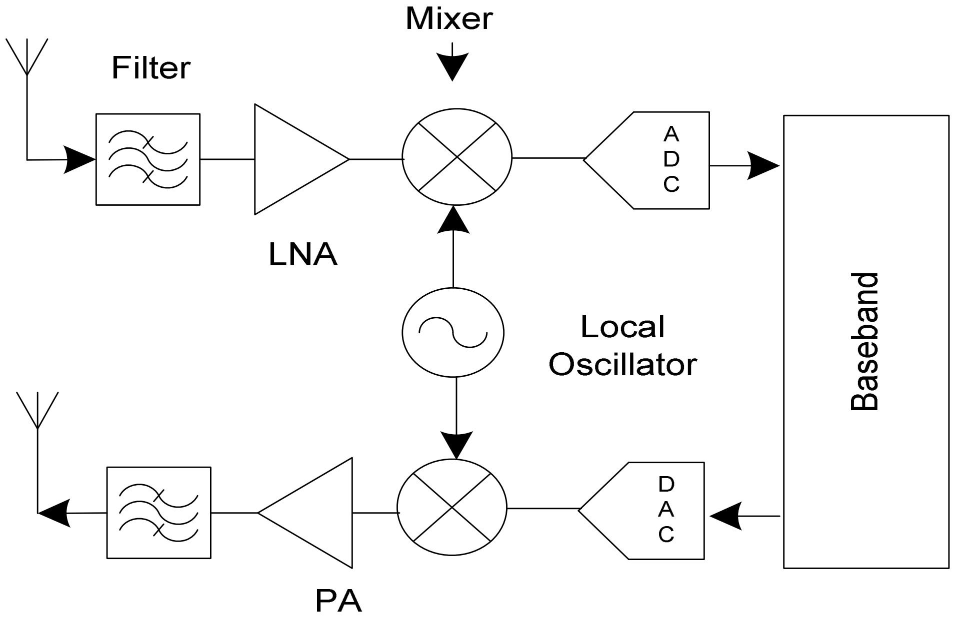
RF Transceiver



Transceiver key RF components

- Antenna, Filters, Power Amplifier (PA), Low-Noise Amplifier (LNA), Oscillator (VCO), Mixer and Data converters (DAC/ADC)

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Nonlinear Device

In the most general sense, the output response of a nonlinear circuit can be modeled as a Taylor series in terms of the input signal voltage

$$v_0 = a_0 + a_1 v_i + a_2 v_i^2 + a_3 v_i^3 + \cdots$$

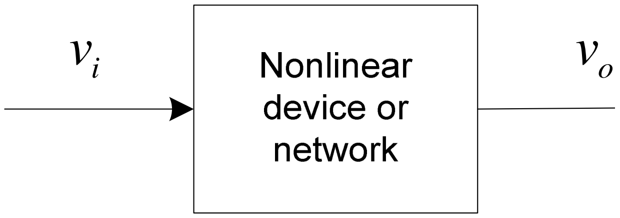
where the Taylor Coefficients are defined as

$$a_0 = v_0(0)$$

$$a_1 = \left.\frac{dv_0}{dv_i}\right| v_i = 0$$

$$a_2 = \left.\frac{d^2 v_0}{dv_i^2}\right| v_i = 0$$

and higher order terms



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Gain Compression

Consider the case where a single frequency sinusoid is applied to the input of a nonlinear device such as a power amplifier

$$v_i = V_0 \cos \omega_0 t$$

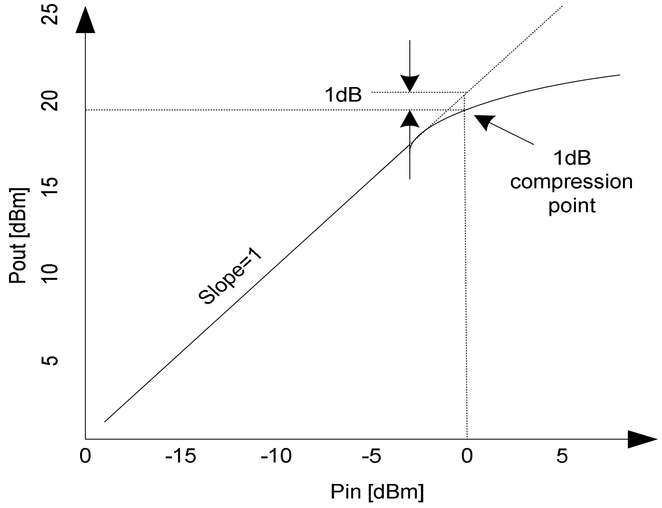
$$v_0 = a_0 + a_1 V_0 \cos \omega_0 t + a_2 V_0^2 \cos^2 \omega_0 t + a_3 V_0^3 \cos^3 \omega_0 t + \cdots$$

$$v_0 = a_0 + a_1 V_0 \cos \omega_0 t + a_2 V_0^2 \left(\frac{1+\cos 2\omega_0 t}{2} \right) +$$

$$a_3 V_0^3 \frac{1}{2} \cos \omega_0 t + a_3 V_0^3 \left(\frac{\cos \omega_0 t + \cos 2\omega_0 t}{4} \right) + \cdots$$

$$v_0 = \left(a_0 + \frac{1}{2} a_2 V_0^2 \right) + \left(a_1 V_0 + \frac{3}{4} a_3 V_0^3 \right) \cos \omega_0 t +$$

$$\frac{1}{2} a_2 V_0^2 \cos 2\omega_0 t + \frac{1}{4} a_3 V_0^3 \cos 3\omega_0 t + \cdots$$



Voltage gain at frequency ω_0

$$G = \frac{a_1 V_0 + \frac{3}{4} a_3 V_0^3}{V_0} = a_1 + \frac{3}{4} a_3 V_0^2$$

a_3 is negative in most practical amplifiers

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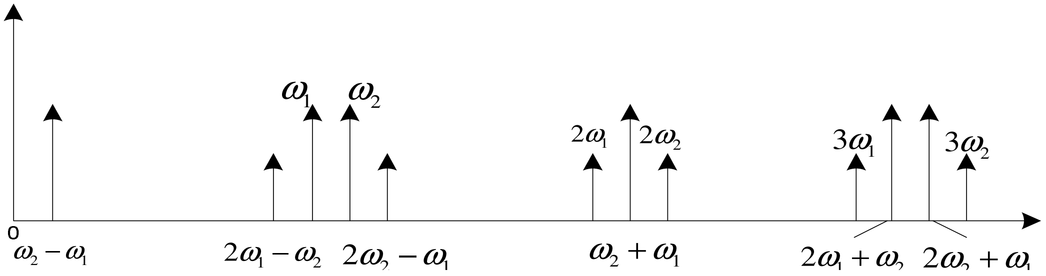
Intermodulation Distortion

Consider two-tone input voltage consisting of two closely spaced frequencies



$$v_i = V_0\left(\cos\omega_1 t + \cos\omega_2 t\right)$$

$$v_o = a_0 + a_1V_0\left(\cos\omega_1 t + \cos\omega_2 t\right) + a_2V_0^2\left(\cos\omega_1 t + \cos\omega_2 t\right)^2 + a_3V_0^3\left(\cos\omega_1 t + \cos\omega_2 t\right)^3 + \cdots$$

$$v_o = a_0 + a_1V_0\cos\omega_1 t + a_1V_0\cos\omega_2 t + \frac{1}{2}a_2V_0^2\left(1+\cos 2\omega_1 t\right) + \frac{1}{2}a_2V_0^2\left(1+\cos 2\omega_2 t\right)$$

$$+a_2V_0^2\cos(\omega_1-\omega_2)t + a_2V_0^2\cos(\omega_1+\omega_2)t + a_3V_0^3\left(\frac{3}{4}\cos\omega_1 t + \frac{1}{4}\cos 3\omega_1 t\right) + a_3V_0^3\left(\frac{3}{4}\cos\omega_2 t + \frac{1}{4}\cos 3\omega_2 t\right)$$

$$+a_3V_0^3\left[\frac{3}{2}\cos\omega_2 t + \frac{3}{4}\cos(2\omega_1-\omega_2)t + \frac{3}{4}\cos(2\omega_1+\omega_2)t\right] + a_3V_0^3\left[\frac{3}{2}\cos\omega_1 t + \frac{3}{4}\cos(2\omega_2-\omega_1)t + \frac{3}{4}\cos(2\omega_2+\omega_1)t\right] + \cdots$$

Output spectrum consists of harmonics of the form

$m\omega_1 + n\omega_2 \quad m,n = 0,\pm1,\pm2\pm3,\cdots$

These combinations of the two input frequencies are called intermodulation products

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Third-Order Intercept Point (IP3)

$$P_{\omega_1} = \frac{1}{2} a_1^2 V_0^2$$

$$P_{2\omega_1 - \omega_2} = \frac{1}{2}\left(\frac{3}{4} a_3 V_0^2\right)^2 = \frac{9}{32} a_3^2 V_0^6$$

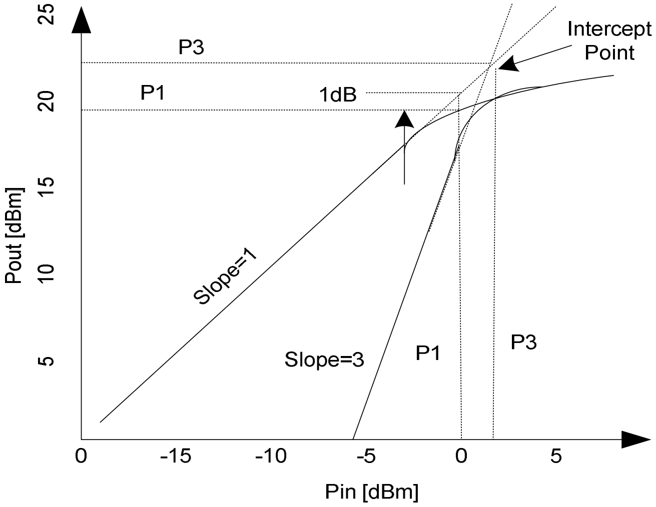
These two powers are equal at the third-order IP

Let input signal voltage at the IP be V_{IP}

$$\frac{1}{2} a_1^2 V_{IP}^2 = \frac{9}{32} a_3^2 V_{IP}^6$$

$$V_{IP} = \sqrt{\frac{4a_1}{3a_2}}$$

$$P_3 = P_{\omega_1}\Big|_{V_0 = V_{IP}} = \frac{1}{2} a_1^2 V_{IP}^2 = \frac{2a_1^3}{3a_3}$$



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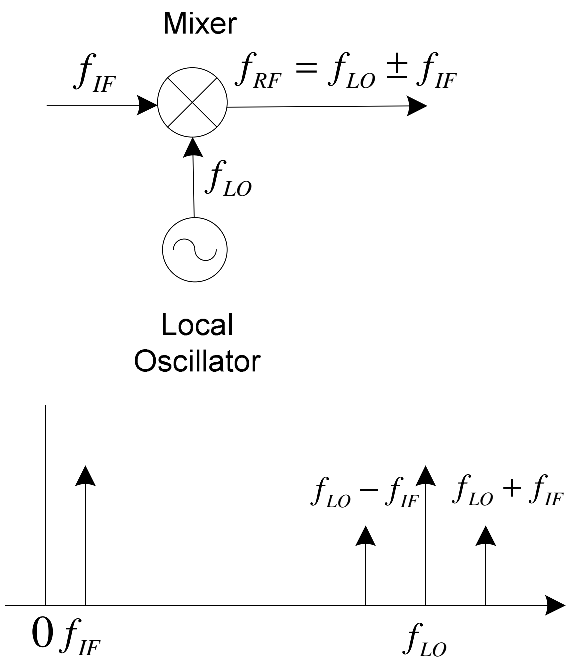
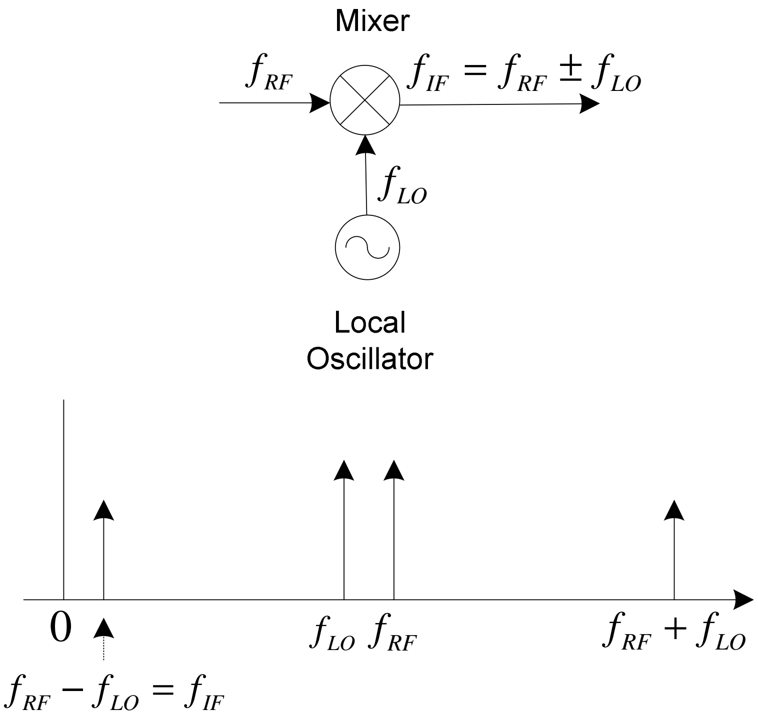
Mixer





$$v_{RF}(t) = K\left[v_{LO}(t)\,v_{IF}(t)\right] = K\cos 2\pi f_{LO}t\cos 2\pi f_{IF}t$$

$$v_{RF}(t) = \frac{K}{2}\left[\cos 2\pi\left(f_{LO} - f_{IF}\right)t + \cos 2\pi\left(f_{LO} + f_{IF}\right)t\right]$$

$$v_{IF}(t) = K\left[v_{LO}(t)\,v_{RF}(t)\right] = K\cos 2\pi f_{LO}t\cos 2\pi f_{RF}t$$

$$v_{IF}(t) = \frac{K}{2}\left[\cos 2\pi\left(f_{RF} - f_{LO}\right)t + \cos 2\pi\left(f_{RF} + f_{LO}\right)t\right]$$

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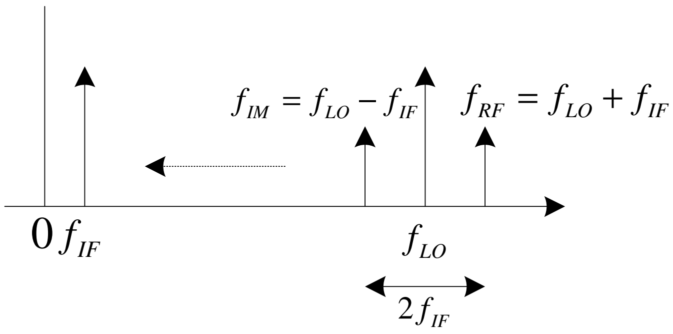
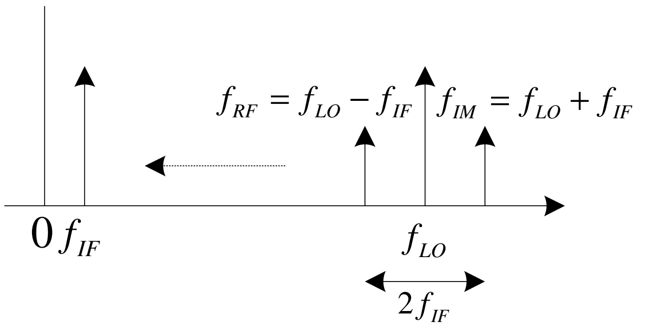
Image Frequency





$$f_{IF} = f_{RF} - f_{LO} = (f_{LO} + f_{IF}) - f_{LO} = f_{IF}$$
$$f_{IF} = f_{IM} - f_{LO} = (f_{LO} - f_{IF}) - f_{LO} = -f_{IF}$$

$$f_{IF} = f_{RF} - f_{LO} = (f_{LO} - f_{IF}) - f_{LO} = -f_{IF}$$
$$f_{IF} = f_{IM} - f_{LO} = (f_{LO} + f_{IF}) - f_{LO} = f_{IF}$$

-f_{IF} is mathematically identical to f_{IF} *because the frequency spectrum of any real signal is symmetric about zero frequency , and thus contains negative as well as positive frequencies*

A received RF signal at the image frequency f_{IM} is indistinguishable at the IF stage from the desired RF signal of frequency f_{RF}

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Homodyne (Zero-IF) Receiver

$$I(t) = \cos\omega_{RF}t\cos\omega_{RF}t = \cos^2\omega_{RF}t = \frac{1+\cos 2\omega_{RF}t}{2}$$

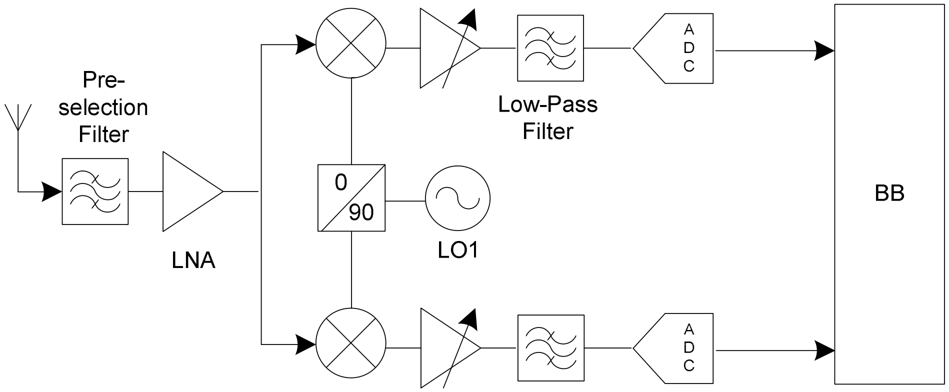
After low pass filtering

$$I(t) = \frac{1}{2}$$

$$Q(t) = \sin\omega_{RF}t\sin\omega_{RF}t = \sin^2\omega_{RF}t = \frac{1-\cos 2\omega_{RF}t}{2}$$

After low pass filtering

$$Q(t) = \frac{1}{2}$$



Benefits	Drawbacks
Less hardware	LO Leakage
Low power consumption	DC offset errors
No IF stage and hence no image filter	I/Q mis-match
	Flicker (or 1/*f*) noise

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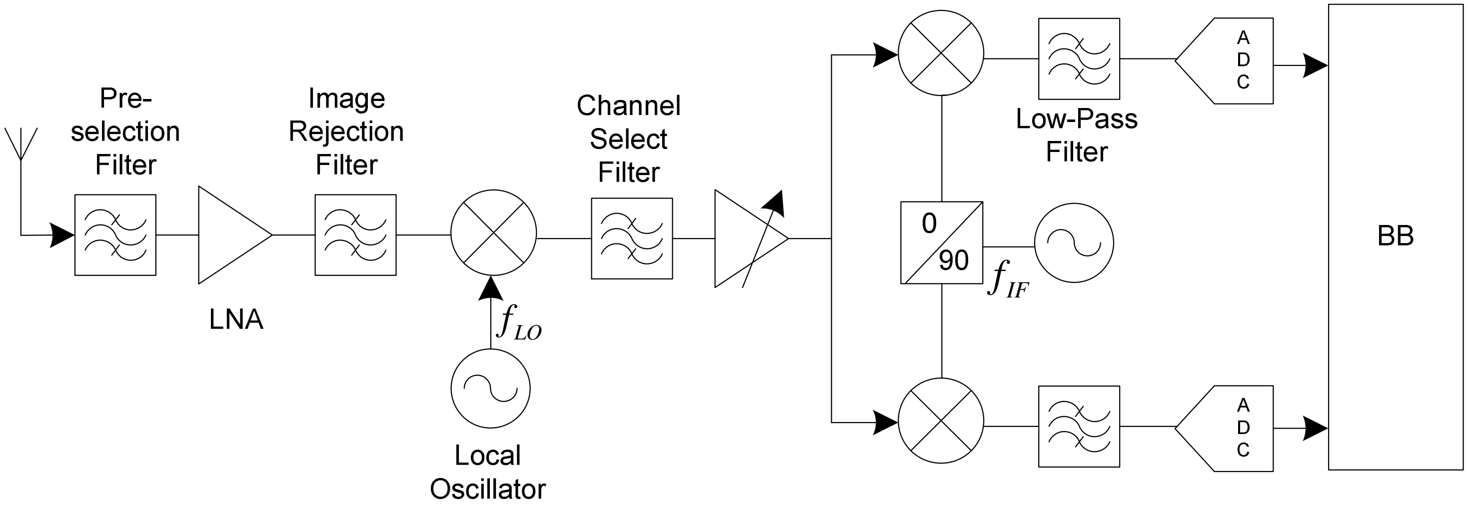
Super-heterodyne Receiver



Benefits	Drawbacks
Good sensitivity	High Q filter
Good selectivity	High performance oscillator
	LNA output impedance matched to 50 ohm is difficult
	Integration of HF image reject filter is a major problem

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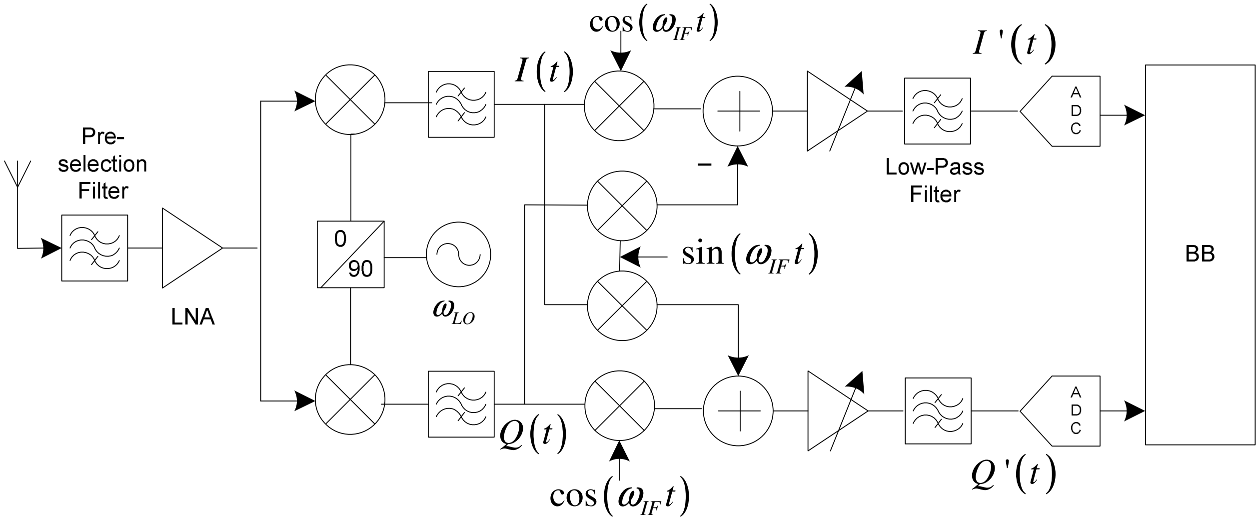
Wideband-IF Receiver



Benefits	Drawbacks
Image cancellation by IR mixer	IR Mixer
Image rejection from the RF front-end pre-selection filter	
Good phase noise performance	

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Wideband-IF Receiver (Image Rejection)

Low-side injection $\omega_{RF}-\omega_{LO}=\omega_{LO}-\omega_{IM}=\omega_{IF}$

Signal of interest $x_{RF}\cos\left(\omega_{RF}t-\alpha\right)=x_{RF}\cos\alpha\cos\omega_{RF}t+x_{RF}\sin\alpha\sin\omega_{RF}t$

Image $x_{IM}\cos\left(\omega_{IM}t-\beta\right)=x_{IM}\cos\beta\cos\omega_{IM}t+x_{IM}\sin\beta\sin\omega_{IM}t$

$$I\left(t\right)=\left[x_{RF}\cos\left(\omega_{RF}t-\alpha\right)+x_{IM}\cos\left(\omega_{IM}t-\beta\right)\right]\cos\omega_{LO}t$$

$$=\frac{1}{2}x_{RF}\left\{\cos\left(\omega_{IF}t-\alpha\right)+\cos\left[\left(\omega_{RF}+\omega_{LO}\right)t-\alpha\right]\right\}+\frac{1}{2}x_{IM}\left\{\cos\left(\omega_{IF}t+\beta\right)+\cos\left[\left(\omega_{IM}+\omega_{LO}\right)t-\beta\right]\right\}$$

$$Q\left(t\right)=\left[x_{RF}\cos\left(\omega_{RF}t-\alpha\right)+x_{IM}\cos\left(\omega_{IM}t-\beta\right)\right]\sin\omega_{LO}t$$

$$=\frac{1}{2}x_{RF}\left\{-\sin\left(\omega_{IF}t-\alpha\right)+\sin\left[\left(\omega_{RF}+\omega_{LO}\right)t-\alpha\right]\right\}+\frac{1}{2}x_{IM}\left\{\sin\left(\omega_{IF}t+\beta\right)+\sin\left[\left(\omega_{IM}+\omega_{LO}\right)t-\beta\right]\right\}$$

After low-pass filtering

$$I\left(t\right)=\frac{1}{2}\left[x_{RF}\cos\left(\omega_{IF}t-\alpha\right)+x_{IM}\cos\left(\omega_{IF}t+\beta\right)\right],\ \ Q\left(t\right)=\frac{1}{2}\left[-x_{RF}\sin\left(\omega_{IF}t-\alpha\right)+x_{IM}\sin\left(\omega_{IF}t+\beta\right)\right]$$

$$I'\left(t\right)=I\left(t\right)\cos\omega_{IF}t-Q\left(t\right)\sin\omega_{IF}t=\frac{1}{2}x_{RF}\cos\alpha+\frac{1}{2}x_{IM}\cos\left(2\omega_{IF}t+\beta\right)$$

$$Q'\left(t\right)=I\left(t\right)\sin\omega_{IF}t+Q\left(t\right)\cos\omega_{IF}t=\frac{1}{2}x_{RF}\sin\alpha+\frac{1}{2}x_{IM}\sin\left(2\omega_{IF}t+\beta\right)$$

After low-pass filtering

$$I'\left(t\right)=\frac{1}{2}x_{RF}\cos\alpha,\ Q'\left(t\right)=\frac{1}{2}x_{RF}\sin\alpha$$

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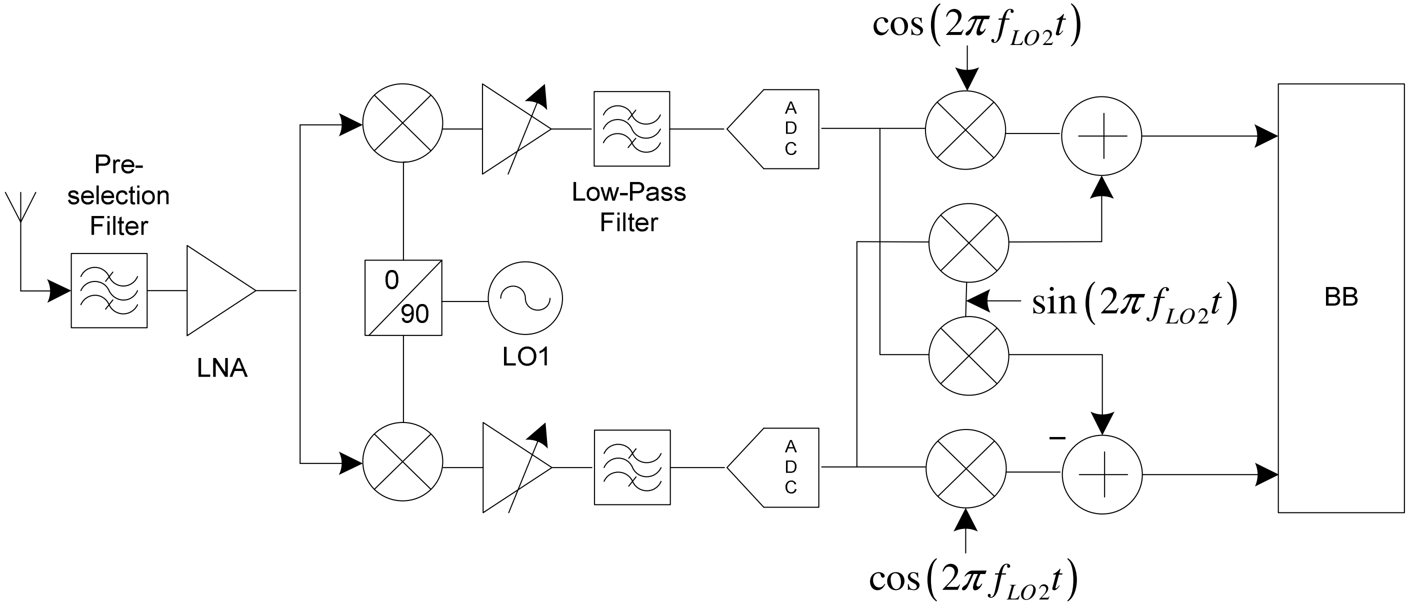
Low-IF Receiver



Benefits	Drawbacks
Potential advantages of both heterodyne and homodyne receivers.	ADC dynamic range
The IF frequency is just one or two channels bandwith away from DC, which is just enough to overcome DC offset problems.	
Image reject mixer which is implemented in digital baseband	

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Downlink RF Transceiver Requirement

Base station transmitter

- Transmit antennas / antenna arrays
  - 20 – 30 dB antenna gain, horn antennas or phase antenna arrays (64 – 1024 elements)
- Power amplifier
  - 20 – 50 dBm, >20% efficiency, EVM < 5% for OFDM waveform
- Packaging
  - Integrated solution of antenna array / PA / MMIC / RFIC to minimize transmission loss

Mobile station receiver

- Receive antenna arrays
  - 6 – 18 dB antenna gain, phase antenna arrays (4 – 64 elements)
  - Receiver sensitivity < -80dBm
  - Total Rx chain Noise Figure < 7dB
- Similar solutions exist today!
  - 60GHz CMOS RFIC with phase antenna array (BWRC)
  - 60GHz Single-chip integrated antenna and RFIC (GEDC)
- Packaging
  - Integrated solution of antenna array / LNA / MMIC / RFIC to minimize transmission loss

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Uplink RF Transceiver Requirement

Mobile station transmitter

- Transmit antenna arrays
  - 6 – 18 dB antenna gain, phase antenna arrays (4 – 64 elements)
- Power amplifier
  - 20 – 23 dBm, >20% efficiency, EVM < 5% for 16QAM single-carrier waveform
- Packaging
  - Integrated solution of antenna array / PA / MMIC / RFIC to minimize transmission loss
- Power consumption on the order of 100mW ~ 1W

Base station receiver

- Receiving antennas / antenna arrays
  - 20 – 30 dB antenna gain, horn antennas or phase antenna arrays (64 – 1024 elements)
- Receiver sensitivity < -95 dBm
- Total Rx Noise Figure < 5dB
- Packaging
  - Integrated solution of antenna array / PA / MMIC / RFIC to minimize transmission loss

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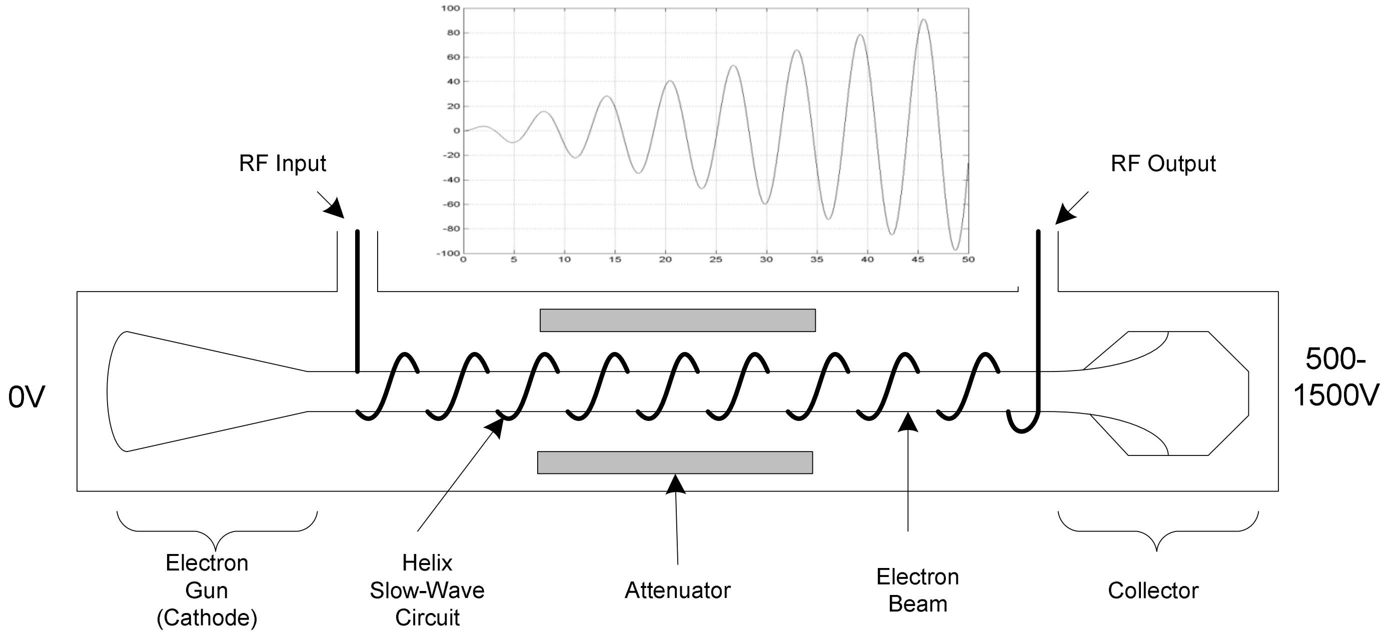
Travelling Wave Tube (TWT) Power Amplifier



TWT amplifiers have been extensively used for high power applications at millimeter wave frequencies

- Provides KWs to MWs power for satellite and radar
- Cost in 10K's of US$ (too expensive for cellular)

Need to consider solid-state amplifier design for MMB

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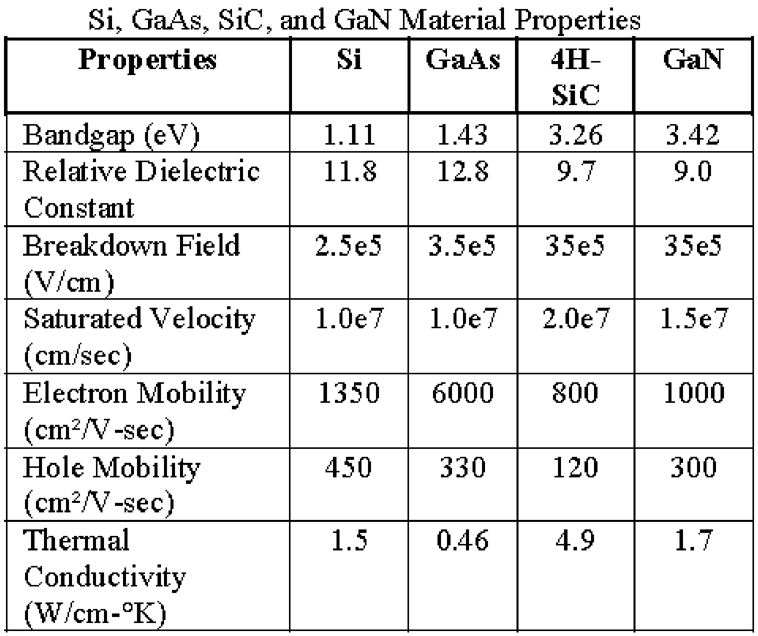
Solid-state Power Amplifier

Gallium-Nitride based power amplifier

- Wide bandgap materials such as gallium nitride (GaN) or silicon carbide (SiC) have much larger bandgaps than conventional semiconductors
- Gallium-nitride High Electron Mobility Transistor (GaN HEMT) devices have breakdown voltages 10 times higher than GaAs HEMT devices, allowing GaN HEMT devices to operate with much higher voltages



Si, GaAs, SiC, and GaN Material Properties

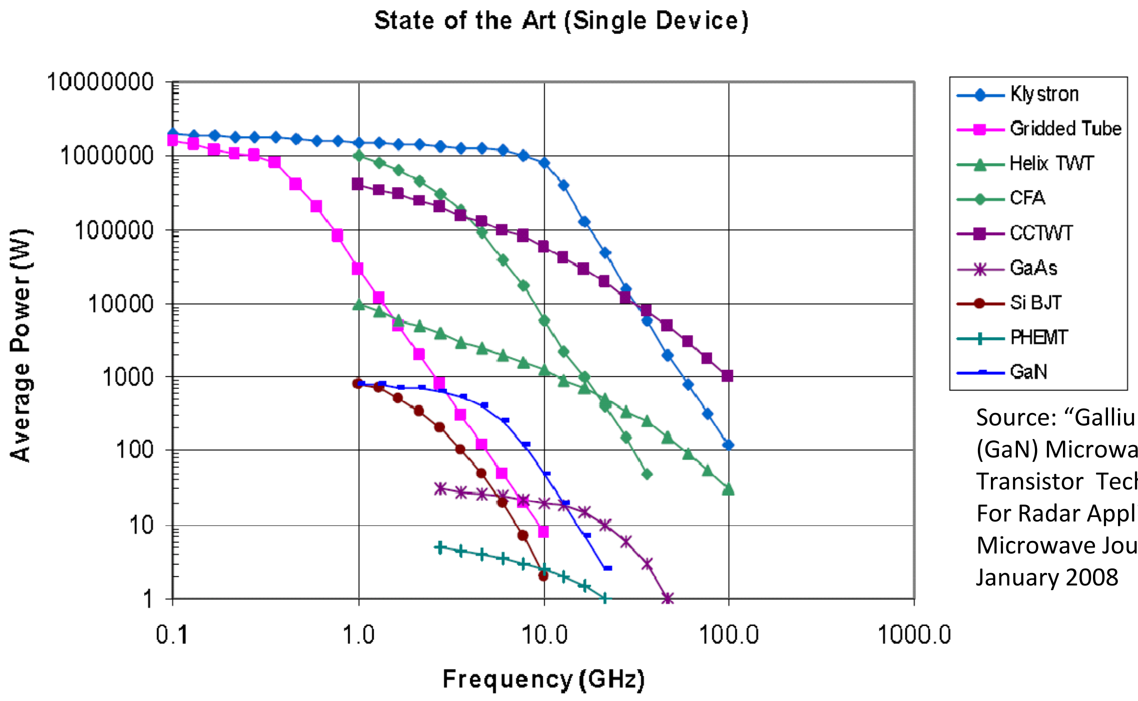
Properties	**Si**	**GaAs**	**4H-SiC**	**GaN**
Bandgap (eV)	1.11	1.43	3.26	3.42
Relative Dielectric Constant	11.8	12.8	9.7	9.0
Breakdown Field (V/cm)	2.5e5	3.5e5	35e5	35e5
Saturated Velocity (cm/sec)	1.0e7	1.0e7	2.0e7	1.5e7
Electron Mobility (cm^2/V-sec)	1350	6000	800	1000
Hole Mobility (cm^2/V-sec)	450	330	120	300
Thermal Conductivity (W/cm-°K)	1.5	0.46	4.9	1.7

Source: “Gallium Nitride (GaN) Microwave Transistor Technology For Radar Applications”, Microwave Journal, January 2008

Solid-state Power Amplifier

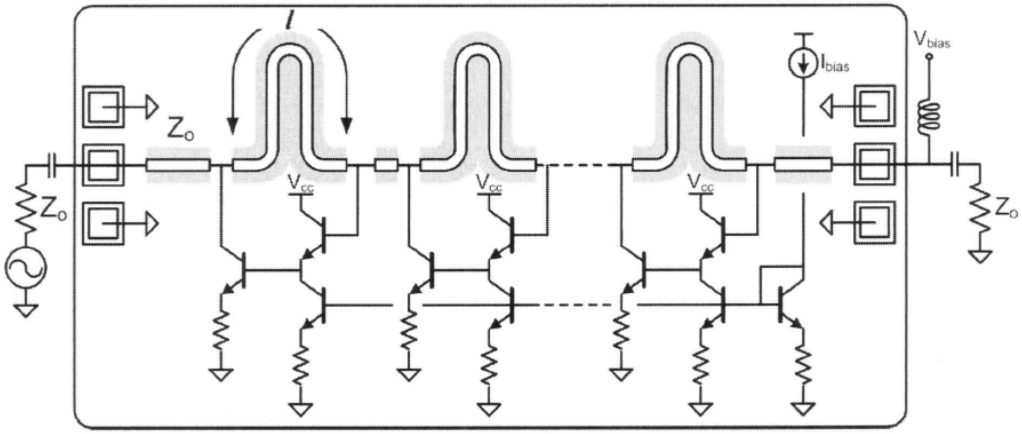
State-of-the-art for solid-state mmWave PAs

- 11 Watts at 34 GHz (D. C. Streit, et. al., “The future of compound semiconductors for aerospace and defense applications”, CSIC 2005)
- 842 mW at 88 GHz (M. Micovic, et. al., “W-Band GaN MMIC with 842mW output power at 88 GHz”, IMS 2010)
- 5.2 Watts at 95 GHz with a 12-way radial-line combiner (James Schellenberg, et. al., “W-Band, 5W solid-state power amplifier/combiner”, IMS 2010)



Source: “Gallium Nitride (GaN) Microwave Transistor Technology For Radar Applications”, Microwave Journal, January 2008

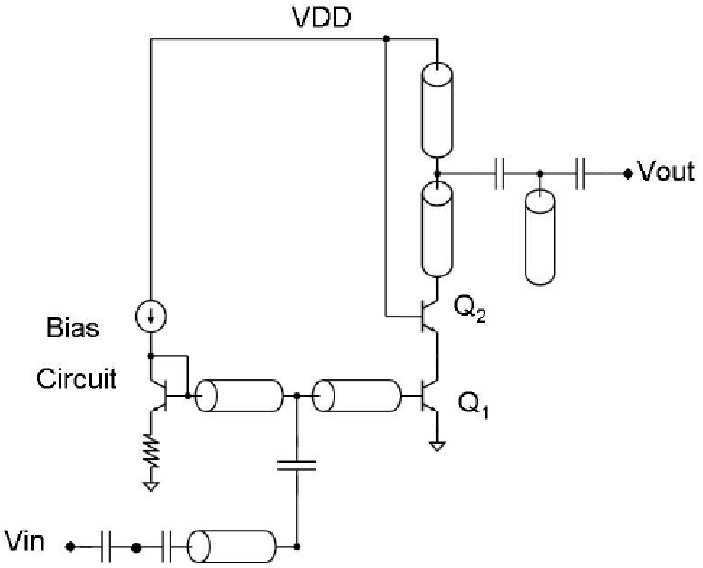
Cascaded Constructive Wave Amplifier

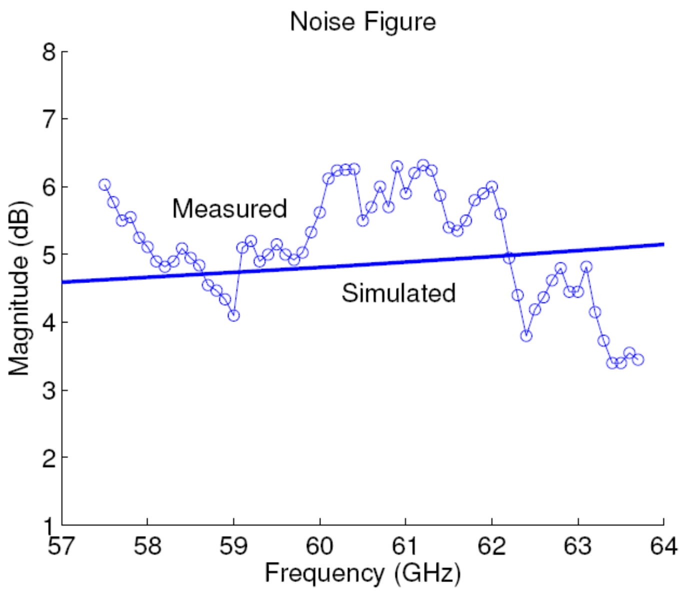


Source: J. Buckwalter and J. Kim, ISSCC 2009

- Forward wave is amplified as it propagates along the transmission line
- Backward wave is attenuated as it propagates
- Distribution of N cascaded traveling wave stages
- Active devices along the transmission line provide feedback
- Relative phase of transmission line and active device determines amplification/ attenuation.

Low-Noise Amplifier [1/2]



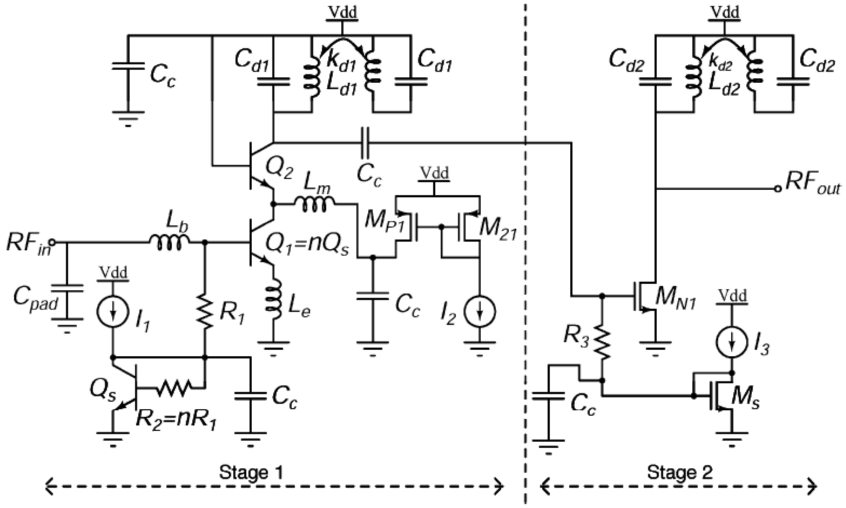


Single Stage 60 GHz LNA

Source: Javier Alvarado, PhD thesis, May 2008

Gain	12 dB
Noise Figure	5 dB over 57 – 64 GHz
Power Consumption	4.5*mA* from a 1.8 V source
1-dB compression point	+1.5dBm
Efficiency	17.4%
Process	IBM0.12 µm, 200 GHz f_T, SiGe technology.

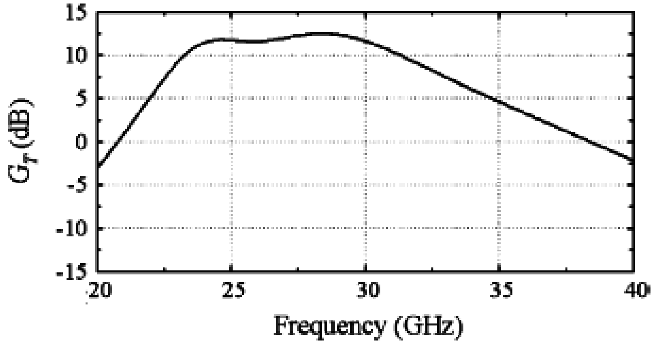
Low-Noise Amplifier [2/2]

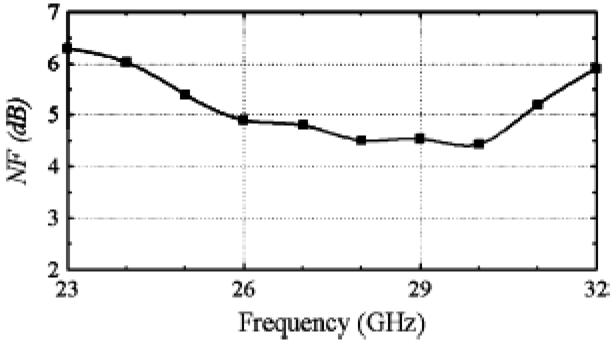


Two Stage 23–32GHz LNA

$$\frac{1}{IP_{3,tot}} = \frac{1}{IP_{3,1}} + \frac{G_1}{IP_{3,2}}$$

Source: El-Nozahi et al,
IEEE JOURNAL OF SOLID-STATE CIRCUITS, FEB 2010





Gain	12 dB
Noise Figure	4.5–6.3dB over 23–32 GHz
Power Consumption	13mW from a 1.5 V source
IP3	-4.5dBm to -6.3dBm [stage1=-2dBm, stage2=7dBm]
Efficiency	NA
Process	Jazz Semiconductor 0.18 m BiCMOS

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MMB downlink budget

Key system configuration parameters

- Base station Tx power: 35dBm – 40dBm
- Base station Tx antenna gain: 17 dB – 23 dB
- Mobile station Rx antenna gain: 3 dB – 10 dB

MMB link downlink budget analysis	Case 1	Case 2	Case 3	Case 4	Case 5	Case 6	Case 7	Case 8
Transmit Power (dBm)	35.00	40.00	35.00	40.00	35.00	40.00	35.00	40.00
Transmit Antenna Gain (dBi)	17.00	17.00	23.00	23.00	17.00	17.00	23.00	23.00
Carrier Frequency (GHz)	28.00	28.00	28.00	28.00	28.00	28.00	28.00	28.00
Distance (km)	0.50	0.50	0.50	0.50	0.50	0.50	0.50	0.50
Propagation Loss (dB)	115.32	115.32	115.32	115.32	115.32	115.32	115.32	115.32
Other losses	20.00	20.00	20.00	20.00	20.00	20.00	20.00	20.00
Receive Antenna Gain (dB)	3.00	3.00	3.00	3.00	10.00	10.00	10.00	10.00
Received Power (dBm)	-80.32	-75.32	-74.32	-69.32	-73.32	-68.32	-67.32	-62.32
Bandwidth (MHz)	500	500	500	500	500	500	500	500
Thermal Noise PSD (dBm/Hz)	-174.00	-174.00	-174.00	-174.00	-174.00	-174.00	-174.00	-174.00
Noise Figure	7.00	7.00	7.00	7.00	7.00	7.00	7.00	7.00
Thermal Noise (dBm)	-80.01	-80.01	-80.01	-80.01	-80.01	-80.01	-80.01	-80.01
SNR (dB)	**-0.31**	**4.69**	**5.69**	**10.69**	**6.69**	**11.69**	**12.69**	**17.69**
Implementation loss (dB)	5.00	5.00	5.00	5.00	5.00	5.00	5.00	5.00
Spectram Efficiency	0.37	0.95	1.12	2.23	1.31	2.50	2.78	4.29
Data rate (Mbps)	**186.08**	**474.53**	**559.37**	**1117.08**	**653.70**	**1250.93**	**1390.35**	**2145.23**

Path loss formula: PL = 141.3 + $20\log_{10}d$ with d in km (free-space loss + 20dB)

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MMB uplink budget

Key system configuration parameters

- Mobile station Tx power: 20dBm – 23dBm
- Mobile station Tx antenna gain: 3 dB – 10 dB
- Base station Rx antenna gain: 17 dB – 23 dB

MMB uplink budget analysis	Case 1	Case 2	Case 3	Case 4	Case 5	Case 6	Case 7	Case 8
Transmit Power (dBm)	20.00	23.00	20.00	23.00	20.00	23.00	20.00	23.00
Transmit Antenna Gain (dBi)	3.00	3.00	3.00	3.00	10.00	10.00	10.00	10.00
Carrier Frequency (GHz)	28.00	28.00	28.00	28.00	28.00	28.00	28.00	28.00
Distance (km)	0.50	0.50	0.50	0.50	0.50	0.50	0.50	0.50
Propagation Loss (dB)	115.32	115.32	115.32	115.32	115.32	115.32	115.32	115.32
Other losses	20.00	20.00	20.00	20.00	20.00	20.00	20.00	20.00
Receive Antenna Gain (dB)	17.00	17.00	23.00	23.00	17.00	17.00	23.00	23.00
Received Power (dBm)	**-95.32**	**-92.32**	**-89.32**	**-86.32**	**-88.32**	**-85.32**	**-82.32**	**-79.32**
Bandwidth (MHz)	50	50	50	50	50	50	50	50
Thermal Noise PSD (dBm/Hz)	-174.00	-174.00	-174.00	-174.00	-174.00	-174.00	-174.00	-174.00
Noise Figure	5.00	5.00	5.00	5.00	5.00	5.00	5.00	5.00
Thermal Noise (dBm)	-92.01	-92.01	-92.01	-92.01	-92.01	-92.01	-92.01	-92.01
SNR (dB)	**-3.31**	**-0.31**	**2.69**	**5.69**	**3.69**	**6.69**	**9.69**	**12.69**
Implementation loss (dB)	5.00	5.00	5.00	5.00	5.00	5.00	5.00	5.00
Spectram Efficiency	0.20	0.37	0.67	1.12	0.80	1.31	1.98	2.78
Data rate (Mbps)	**9.92**	**18.61**	**33.32**	**55.94**	**39.92**	**65.37**	**98.96**	**139.03**

Path loss formula: PL = 141.3 + 20log$_{10}$d with d in km (free-space loss + 20dB)

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Link Budget Analysis Summary

MMB downlink budget

- Low end: 35 dBm Tx power, 17 dB Tx antenna gain, 3 dB Rx antenna gain, 5 dB implementation loss ➔ 180 Mbps on 500 MHz bandwidth at 500 meters
- High end: 40 dBm Tx power, 23 dB Tx antenna gain, 10 dB Rx antenna gain, 5 dB implementation loss)➔ 2145 Mbps on 500 MHz bandwidth at 500 meters

MMB uplink budget

- Low end: 20 dBm Tx power, 3 dB Tx antenna gain, 17 dB Rx antenna gain, 5 dB implementation loss➔ 9.92 Mbps on 50 MHz bandwidth at 500 meters
- High end: 23 dBm Tx power, 10 dB Tx antenna gain, 23 dB Rx antenna gain, 5 dB implementation loss➔ 139 Mbps on 50 MHz bandwidth at 500 meters

Conclusion

- Assuming free-space plus 20dB path loss, MMB can provide 100 Mbps ~ 2 Gbps ***cell-edge*** throughput on the downlink and 10 Mbps ~ 100 Mbps ***cell-edge*** throughput on the uplink at 28 GHz for cell radius of 500 meters.

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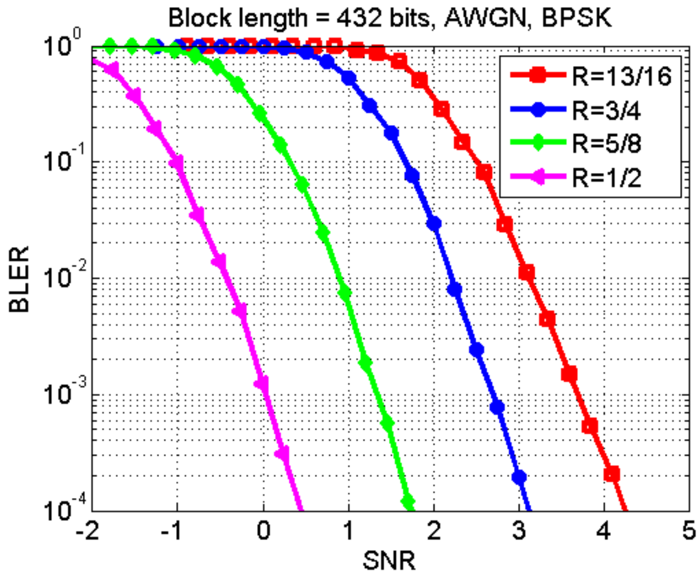
Link Level Performance

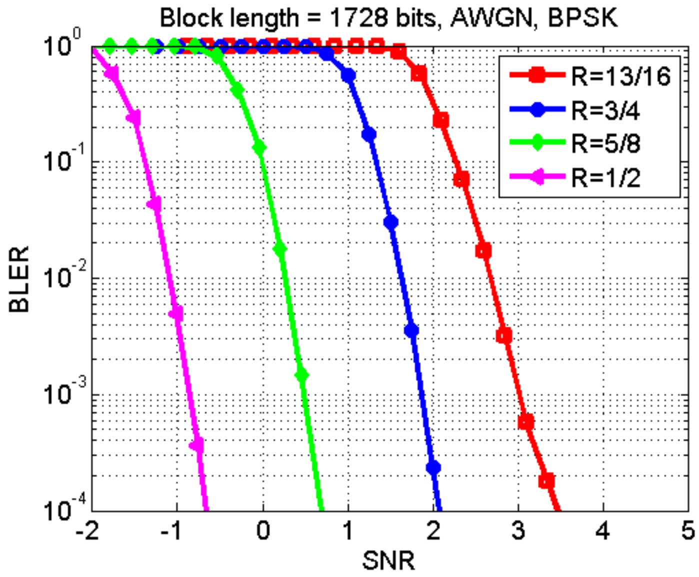
Length-432 and Length-1728 LDPC

Code rate 1/2, 5/8, 3/4, 13/16

Layered decoding

Maximum number of iterations


Block length = 432 bits, AWGN, BPSK


Block length = 1728 bits, AWGN, BPSK

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System Level Performance

19 cells wrap-around

12 sectors per cell

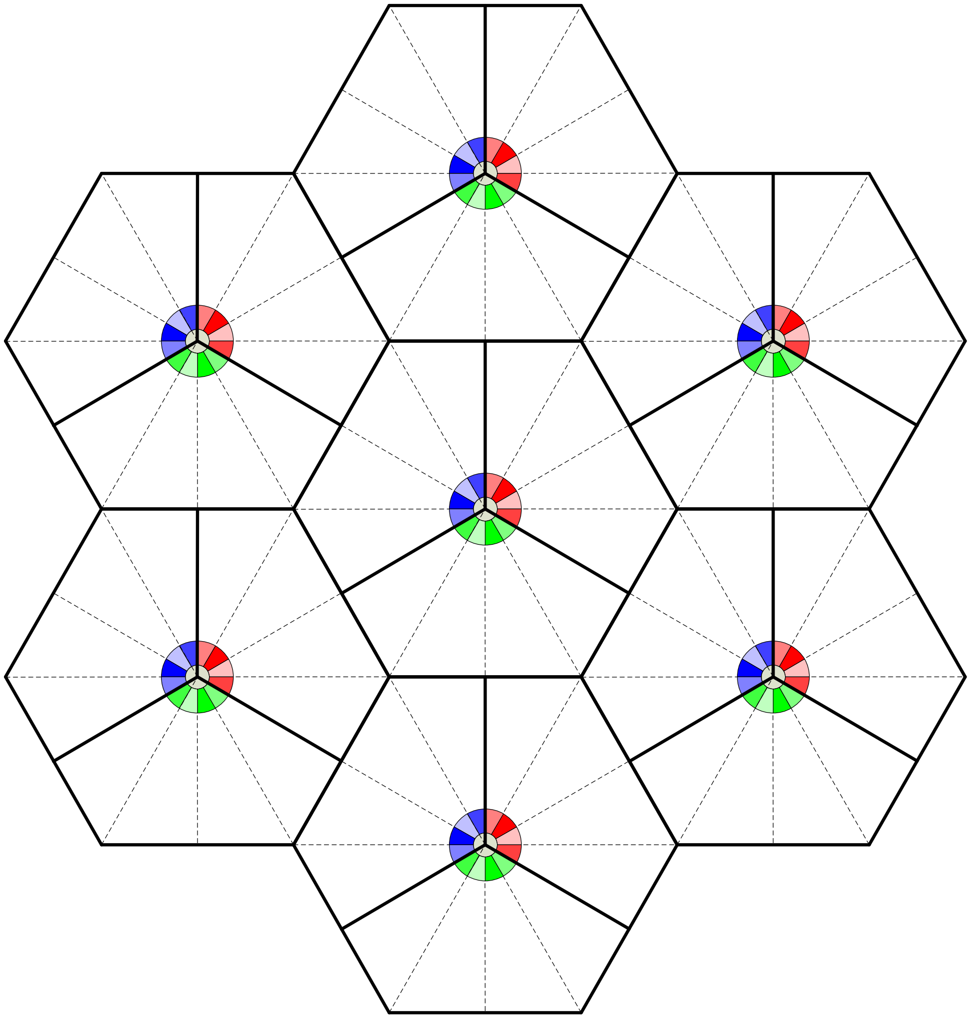
1 horn antenna per sector

20 dB antenna gain

- 17.5^{o} 3-dB beamwidth in azimuth domain
- 10^{o} 3-dB beamwidth in elevation domain
- 30 dB front-to-back ratio

Base station Tx power = 13, 16, 19, 22 dBm/MHz

Mobile station uniformly dropped in the coverage area



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Geometry with Single Rx Antenna

Site-to-site distance = 500 meters

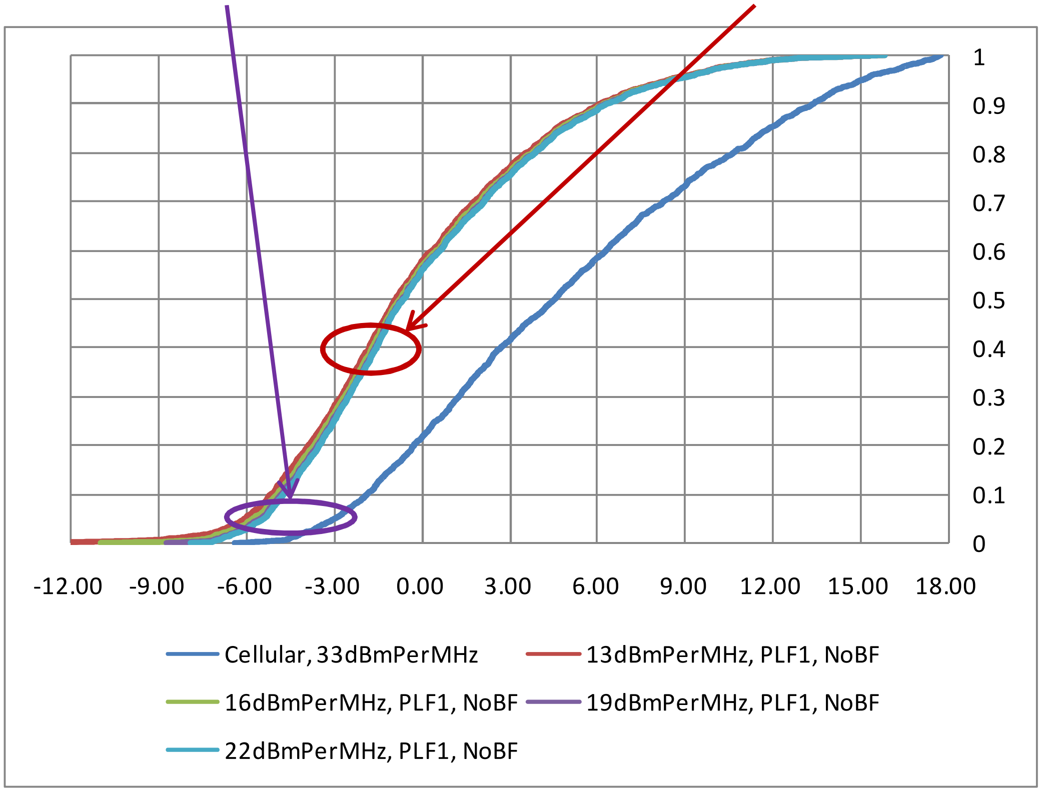
Single Rx antenna with -1 dB antenna gain

No Rx beamforming

PLF1: PL = 141.3 + $20\log_{10}d$ with d in km

12dB Lognormal shadowing

- 8dB Lognormal shadowing for cellular


3dB worse 5%-tile geometry than cellular

Interference limited

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Geometry with Single Rx Antenna

Site-to-site distance = 500 meters

Single Rx antenna with -1 dB antenna gain

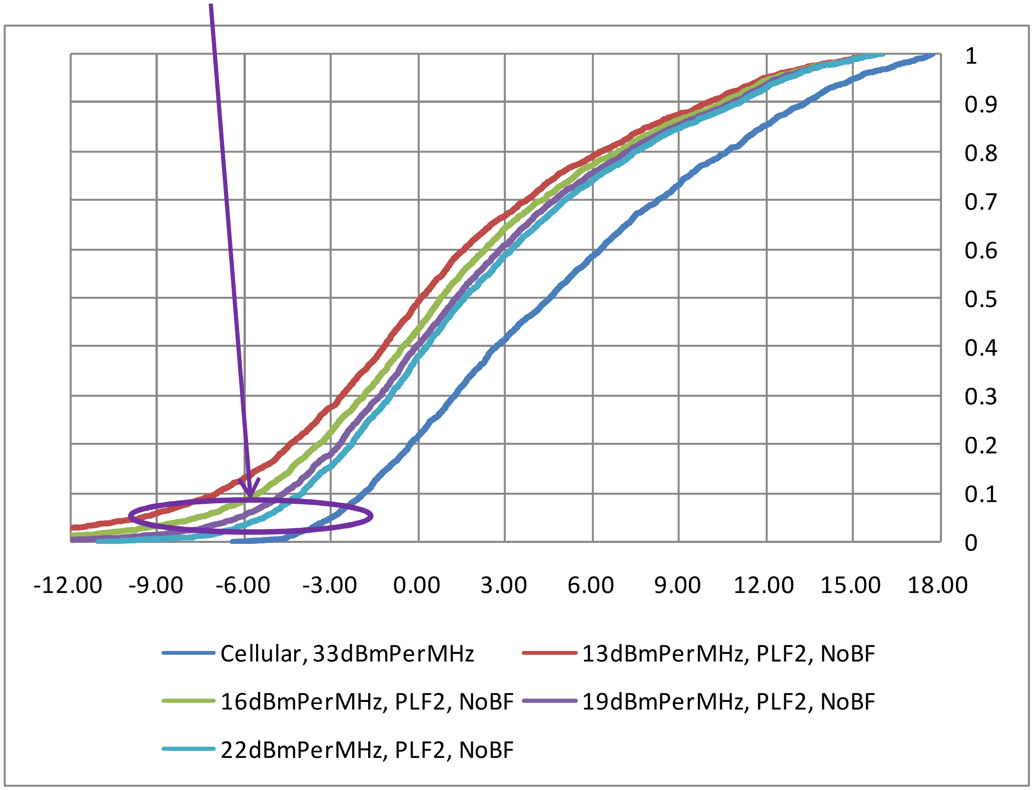
No Rx beamforming

PLF2: PL = 157.4 + $32\log_{10}d$ with d in km

12dB Lognormal shadowing

- 8dB Lognormal shadowing for cellular

2 – 6 dB worse 5%-tile geometry than cellular



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Mobile station Rx beamforming

$$\varphi = \phi + kd\cos\theta$$

$$AF = \frac{\sin\left(\frac{N\varphi}{2}\right)}{\sin\left(\frac{\varphi}{2}\right)}$$

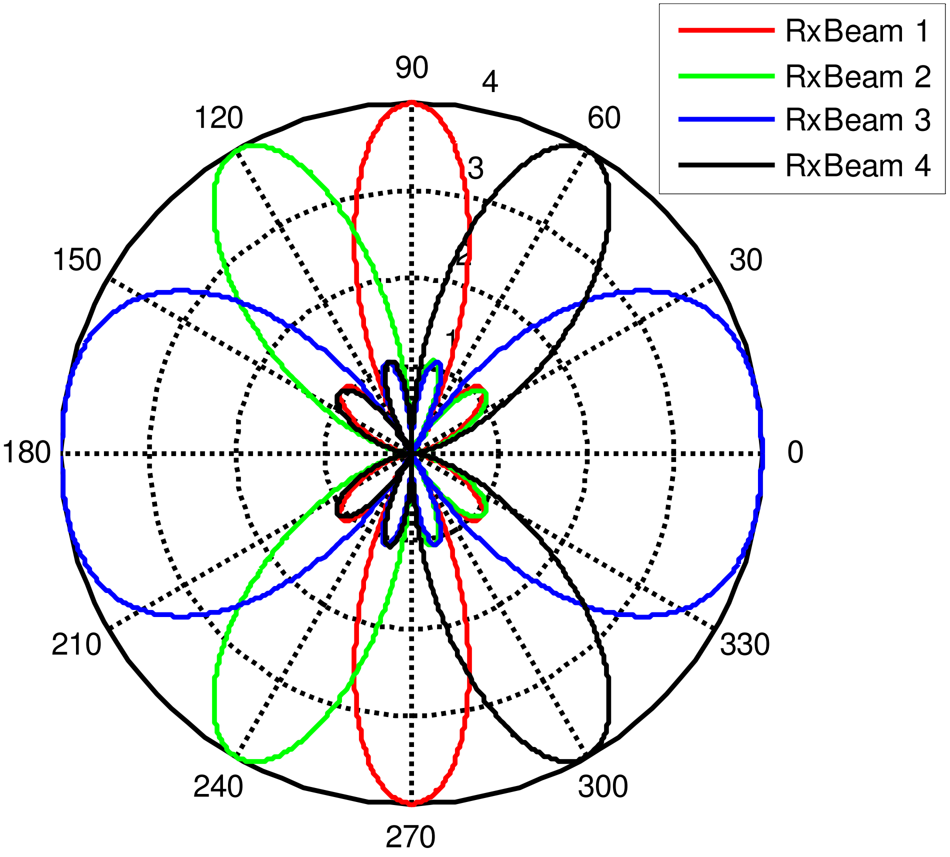
4-element uniform linear array

N=4, d=λ/2, k=2π/λ

4 fixed beams (ϕ=0, π/2, π, 3π/2)

Mobile station orientation is random ~ U[0, 2π)

Mobile station selects the beam that maximizes geometry



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Geometry with Rx Beamforming

Site-to-site distance = 500 meters

4-element antenna array with -3 dB antenna gain per element

Rx beamforming

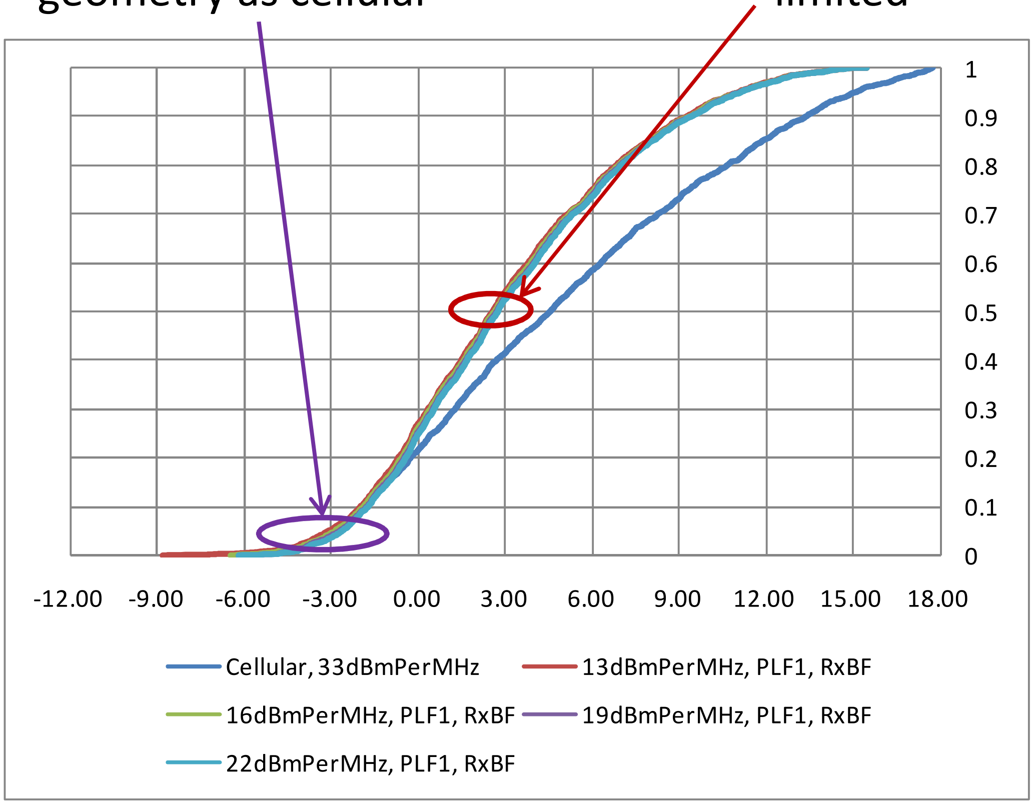
PLF1: PL = 141.3 + $20\log_{10}d$ with d in km

12dB Lognormal shadowing

- 8dB Lognormal shadowing for cellular

The same 5%-tile geometry as cellular

Interference limited



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Geometry with Rx Beamforming

Site-to-site distance = 500 meters

4-element antenna array with -3 dB antenna gain per element

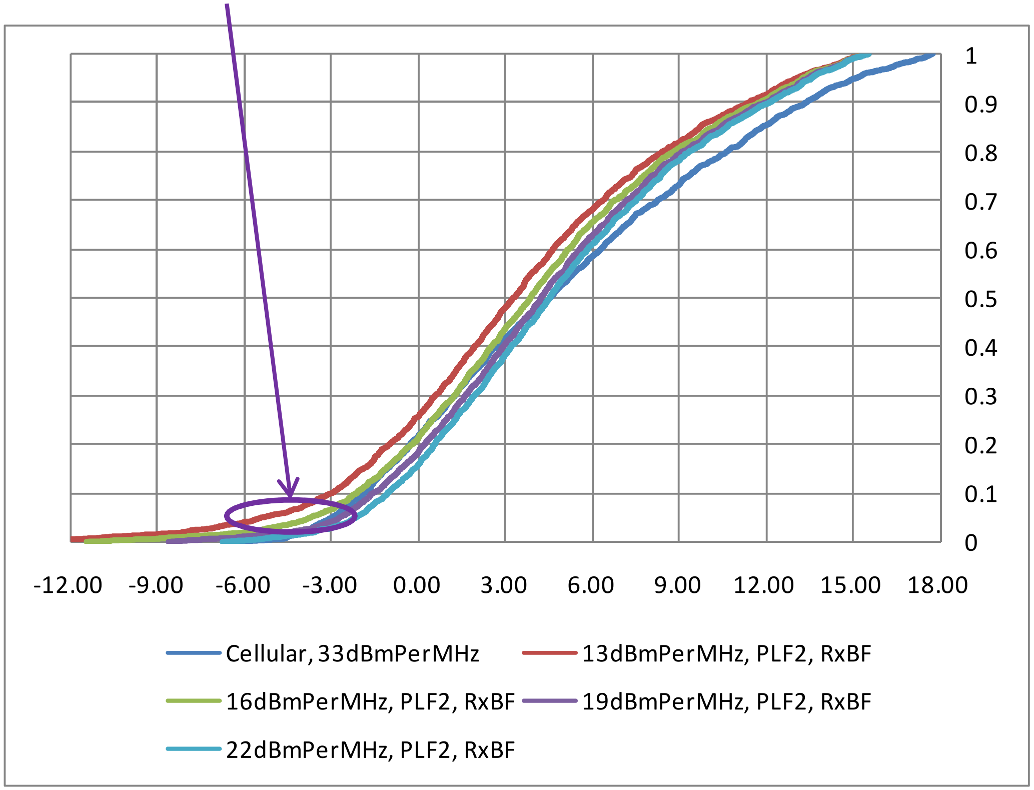
Rx beamforming

PLF1: PL = 157.4 + $32\log_{10}d$ with d in km

12dB Lognormal shadowing

- 8dB Lognormal shadowing for cellular

0-3dB worse 5%-tile geometry than cellular



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System Throughput Analysis

MMB system performance analysis assumptions	
Number of cells	19
Number of sectors per cell	12
Site to site distance	500 meters
Carrier frequency	28 GHz
System bandwidth	500 MHz
Path loss model	$141.3 + 20\log_{10}d$, or $157.4 + 20\log_{10}d$
Base station Tx power	40, 43, 46, or 49 dBm
Base station Tx antenna configuration	Single horn antenna
Base station Tx antenna gain	20 dB
Log-normal shadow fading STD	12 dB
Mobile station Rx noise figure	7 dB
Mobile station Rx antenna configuration	Single antenna, or Rx beamforming with 4-element ULA
Mobile station Rx antenna gain	-1 dB for single antenna case, -3 dB for ULA case
System overhead (cyclic prefix, control channels, etc.)	40%
Transceiver implementation loss	3 dB

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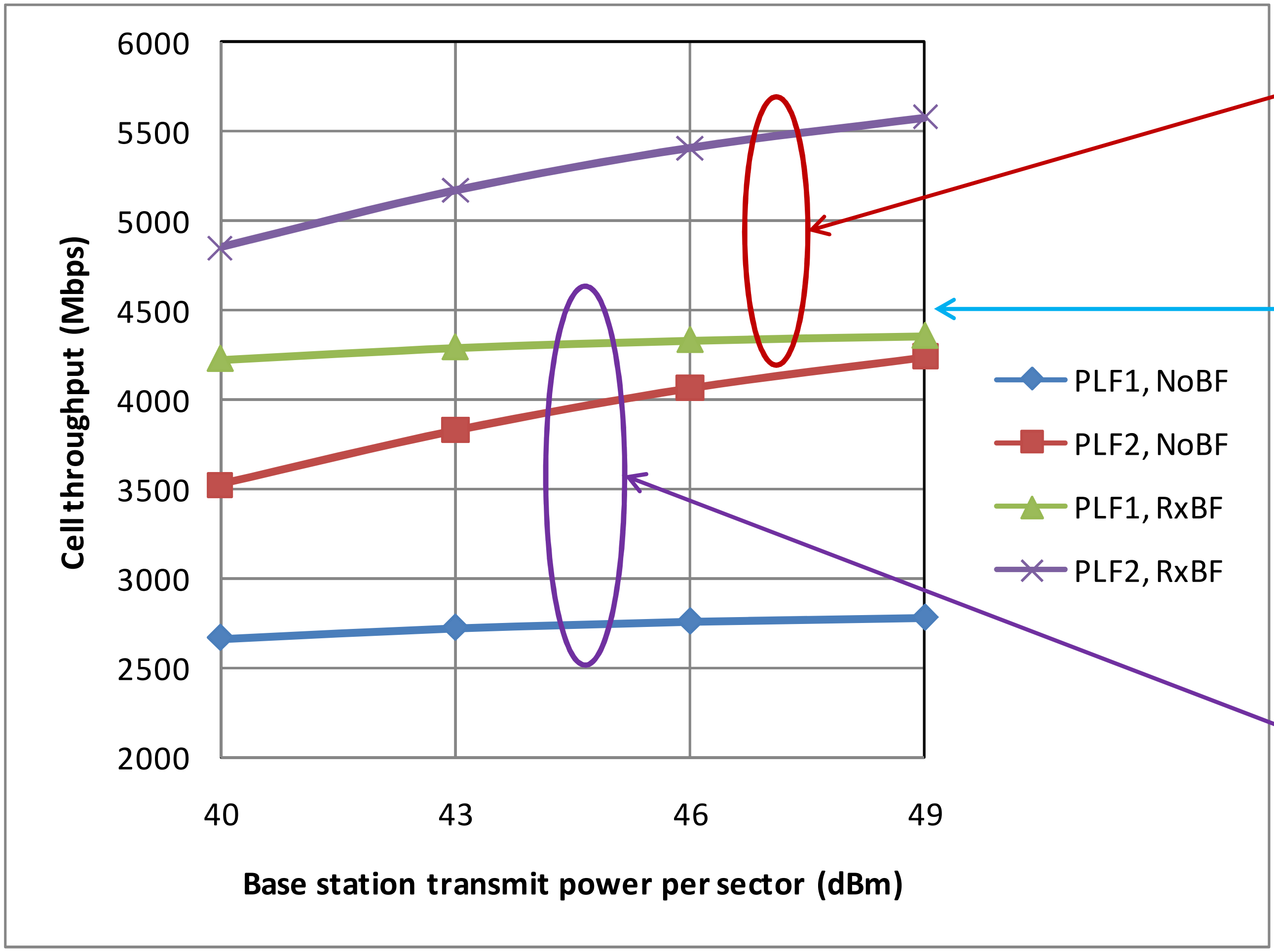
System Throughput



PLF2 provides higher system throughput than PLF1

30x LTE cell throughput (4Tx MIMO per sector, 3-sector cell, 20MHz system bandwidth)

RxBF significantly improves system throughput

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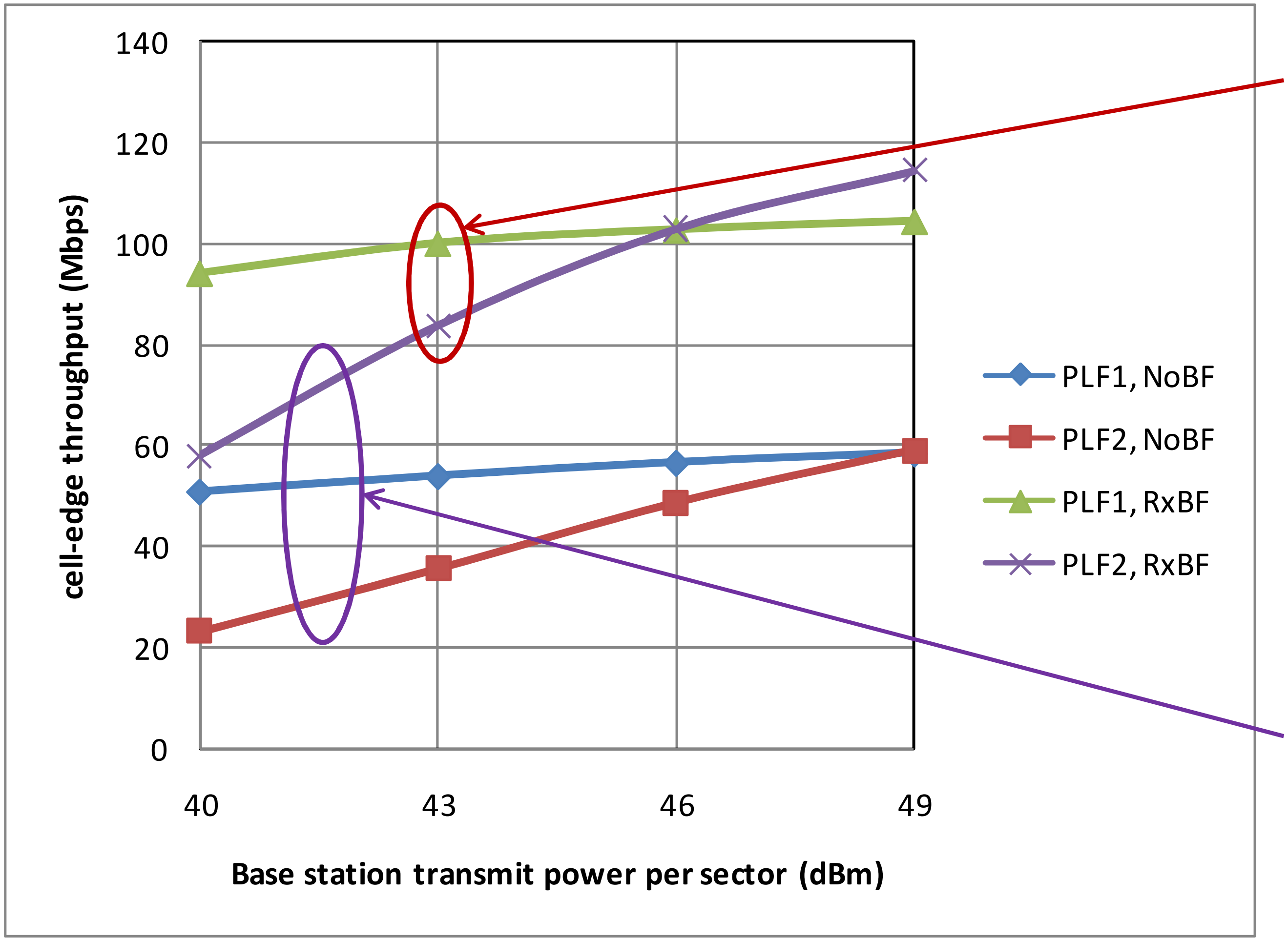
Cell-edge performance



PLF1 allows better cell-edge throughput than PLF2

RxBF significantly improves system throughput

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Summary

Millimeter wave spectrum (3-300GHz) can potentially provide the bandwidth required for mobile broadband applications for the next few decades and beyond.

- Opportunity to open 200 times the spectrum currently allocated for cellular below 3GHz.

Propagation and other losses due to rain, foliage and penetration through building materials needs better understanding

Millimeter waves are also attractive for mobile application due to small component sizes such as antennas.

- Further research is needed towards components and devices that meets mobile application demand of higher power and efficiency

Wireless community should take on the growing data demand by exploiting millimeter wave spectrum paving the way for multi-Gbps mobile broadband.

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