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**Ultrafast Optoelectronics and Electronics:**

**What?**
Generation, propagation, and detection of picosecond (and femtosecond) electrical & optical transients

**Why?**

- **Basic Technology/Science:**
  Generation and detection of pulse & CW signals in the 100 GHz-10 THz region (the spectral "black hole")

- **Characterization of mm-wave circuits and devices.**
  Network analysis, waveform measurements, internal probing

- **Measurement of semiconductor material and electron transport properties relevant to devices.**
  50 GHz circuit needs $\approx$100-200 GHz transistors.
  200 GHz transistor needs $\approx$1 ps carrier transit times.

- **Optoelectronic, Electronic Devices for High-Speed Applications**
  fiber and wireless communications at 10 GHz and beyond.
Future High-Frequency Applications:

• 10 / 40 / 100 / 160... GB/sec fiber-optic data transmission

• Wireless microwave & mm-wave digital radio links
  capacity for the data explosion
  bandwidth is cheap at mm-waves

• mm-wave radar & imaging, car collision avoidance
  runway imaging, fly-by-night, military
  small antennas, high resolution

• Radio astronomy / Earth remote sensing
  THz heterodyne radiometers on Satellites
  ozone depletion

• ... & instrumentation for these applications
The Electronic Bottleneck?

The issue:
- Can we make electronics at 100 GHz and above?

Today's solid-state devices:
- III/V HBTs: $f_{\text{max}} \approx 400$ GHz
- Si/Ge HBTs: $f_{\text{max}} \approx 100$ GHz planar, 180 GHz mesa
- HEMTs: $f_{\text{max}} \approx 500$ GHz
- Schottky Diodes: $f_c \approx 20$ THz

Assertions (research objectives):
- device $f_{\text{max}}$ can be improved
- at $f_{\text{max}} / 4$ many circuits can be implemented.
High-Speed Optolectronics: $$$ Realities

- devices, circuits feasible to $\geq 100$ GHz.

- expensive technology: small, high-value market
  - military, instruments, satellites, high-capacity fiber

- mm-waves: bandwidth is cheap, ICs are expensive

- goal: cheap mm-wave ICs for high-volume markets
  - ... lower-cost III-V devices???
  - ... mm-wave silicon ICs???

Feasibility of 100 + GHz electronics is mainly an economic issue
Bandwidth of Semiconductor Devices

Interaction of transit time and RC charging time sets bandwidth

Applies to most semiconductor devices:
- Schottky diodes, photodiodes, RTDs
- Bipolar transistors, field-effect transistors, …
How Do Semiconductor Devices Work?

Signal power generated by sweeping carrier through field.
Current independent of output voltage
Current controlled by input voltage

Alternative:
Negative-resistance devices

Diode

Bipolar

FET
Total effective time constant and scaling

Vertical Ohmic Contacts

Horizontal Ohmic Contact
Vertical Schottky Contact

- Note the differing scaling laws!
What Limits Semiconductor Device Bandwidth?

- Bulk resistances
- Ohmic contact resistances
- Lateral contact access resistances
- These are for **undepleted** semiconductor layers

\[ R_{\text{bulk}} = \frac{\rho D}{A} \]
\[ R_{\text{contact}} = \frac{\rho_{\text{contact}}}{A} \]
\[ R_{\text{cont,horiz}} \propto \frac{1}{L} \]
What Limits Semiconductor Device Bandwidth?

- Depletion layer capacitances
- Depletion layer transit times

\[ C_{depl} = \varepsilon A / D \]

\[ \tau_{transit} \propto D / v_{electron} \]
RC Charging Times And Scaling

Vertical Ohmic Contacts

Vertical Schottky Contact

- note the different scaling law
Super-Scaled, THz-Bandwidth Devices

\[ C \propto \frac{W L}{D} \quad \tau_{\text{transit}} \propto \frac{D}{v_{\text{electron}}} \]

\[ R \propto \frac{1}{L} \]

\[ \rightarrow R C \propto \frac{W}{D} \]

*deep submicron junction stripe widths,*

*deep submicron epitaxial layer thicknesses*

--> THz device bandwidths
2-Terminal Devices

- Photodiodes & Photconductors
- Schottky Diodes
- Resonant Tunnel Diodes
Photodiodes

Photoconductors: light-controlled resistors

Photodiodes: light-controlled current sources:
Photodiode Structures:

Schottky photodiode

- Schottky contact
- Wide band gap cap
- I-absorption layer
- N-ohmic contact

PIN photodiode

- P-ohmic contact
- I-absorption layer
- N-ohmic contact

Schottkys are simpler: smaller, faster but are difficult on many materials: need AllInAs cap on InGas Schottky

Narrow band-gap (absorbing) I-region (N+ and P+ regions are wider-bandgap, hence transparent)
Photodiode Carrier Transit-Time

- Short-circuit current response to a single electron-hole pair

GaAs Electron velocity: \( \approx 1-2 \times 10^7 \) cm/sec, higher for D<1000 Å
GaAs Hole velocity: \( \approx 6 \times 10^6 \) cm/sec (?), never ballistic
Photodiode Transit Time II

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<td><img src="image15" alt="time" /></td>
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\[
G(z, t) = \frac{(1 - R) E_0 \delta(t)}{h \nu} \frac{\alpha}{A} e^{-\alpha z} \quad \text{for} \quad P_{\text{incident}}(t) = E_o \delta(T)
\]
Photodiodes: Optimization of I-layer Thickness

Photodiodes: Optimization of Absorbtion Layer Thickness

Carrier Transit Time:
- proportional to $D$

R-C Charging Time:
- proportional to $A/D$

$$C_{depl} = \sqrt{\frac{A}{D}}$$

$$R_{diode} \propto A^{-1}$$

$$RMS\ duration = \frac{D}{\sqrt{12} V_{carrier}}$$

$$RMS\ duration = C_{depl}\left(R_{load} + R_{diode}\right)$$

Photodiode Impulse Response Duration

Photodiode Bandwidth

Analysis neglects diode series resistance (process-dependent, significant for smaller diodes). $f_{-3dB} = 0.83 f_{RMS}$ for Guassian response
High-Speed InGaAs Photodetector

[Diagram showing the structure of the photodetector with layers labeled as follows:

- \( p \) GBL
- InGaAs 0.6nm
- InP 1.2nm
- InGaAs 0.6nm
- InP 0.9nm
- InGaAs 1.2nm
- InP 0.6nm
- InGaAs 1.2nm
- InP 0.3nm
- InGaAs 0.6nm
- InP 1.2nm
- InGaAs 0.6nm

The diagram also includes layers labeled as:

- \( n \) GBL
- InGaAs 0.6nm
- InP 1.2nm
- InGaAs 0.6nm
- InP 0.9nm
- InGaAs 1.2nm
- InP 0.6nm
- InGaAs 1.2nm
- InP 0.3nm
- InGaAs 0.6nm
- n+ InP 373nm
- InGaAs 5nm
- InP 0.9nm
- InGaAs 0.6nm
- InP 6.6nm
- i InGaAs
- InP
- TiPt/Au/Ni
- Au
- Si InP Substrate

Additionally, there is a graph showing:

- PD impulse Response (AU) with a peak at 3.8 ps FWHM
- Magnitude (dB) with an inset showing -3 dB @ 108 GHz
- Frequency (GHz)
- Delay (ps)
Photodiode Efficiency-Bandwidth Limit

Quantum Efficiency:
\[ \eta = (1 - R)(1 - e^{-\alpha D}) \equiv \alpha D \]

Assuming:
- Antirelection coating: \( R = 0 \)
- Thin absorption layer: \( \alpha D << 1 \)

Bandwidth
\[ f_{3dB} = 0.445 V_{carrier} / D \]

Assuming:
- Transit time >> RC charging time
- Thin absorption layer: \( \alpha D << 1 \)

Efficiency-Bandwidth Product
\[ f_{3dB} \eta = 0.445 V_{carrier} \alpha \]

strongly wavelength-dependent
Beating the Efficiency Bandwidth Limit 1

PIN RCE photodiodes

- Mirrors result in multiple transits of absorbing region by incident light
- Optical impedance-matching problem (use Smith Chart!)
- Thin absorbing layers (fast device) and high efficiency both possible
Other RCE detector configurations

PIN RCE photodiode

Schottky RCE photodiodes

via-etched

flip-chip

---

Ohmic contacts

P+ layer

N+ layer

wide band-gap cap layer

undoped (i-layer, substrate)

Schottky/interconnect metals

Bragg reflector stack

Host (transistor IC) substrate
Waveguide Photodetector with 172 GHz Bandwidth and 76 GHz Bandwidth-Efficiency Product

1.47 ps FWHM

Waveguide (and traveling-wave) photodiodes.
- Optical Illumination perpendicular to electron transport - independent dimensions
- Very small optical aperture....coupling can be hard
Saturation in Photodiodes: Field-Screening

back-of-envelope calculation (skipping integrals)

Electron velocity >> Hole velocity --> hole stored charge dominates

\[ Q_{\text{hole}} \approx J\tau_{\text{hole}} = JD/\nu_{\text{hole}} \]

\[ \Delta E \approx Q_{\text{hole}}/\varepsilon = JD/\nu_{\text{hole}}\varepsilon \]

\[ \Delta V \approx \Delta E \cdot D \approx JD^2/\nu_{\text{hole}}\varepsilon \]

\[ V_{\text{bias}} + \phi \approx JD^2/\nu_{\text{hole}}\varepsilon \]

\[ J_{\text{max}} \approx \left( V_{\text{bias}} + \phi \right) \frac{\varepsilon v_{\text{hole}}}{D^2} \]

is called the space-charge resistance

Vbias is limited by breakdown
Photoconductor: Fast Optical-Electrical Converter

Photoconductors are optically-controlled resistors

\[ V_{\text{out}} = \frac{1}{2} \left( \frac{Z_0}{Z_0 + 1/G} \right) \]

- \( GZ_0 \ll 1 \): linear photodetector
- For \( GZ_0 \gg 1 \), laser-gated switch

- Subpicosecond optical-electrical converters
- Subpicosecond electrical sampling gates
Photoconductive Detectors: DC Characteristics

R is the reflectivity, tau is the carrier lifetime, L is the gap length, W the gap width

Responsivity varies as $1/L^2$

$$G(t) \propto (W/L)(\text{thickness})(\mu_n n + \mu_p p)$$

$$n \cdot (\text{thickness}) = p \cdot (\text{thickness}) = (1 - R)\frac{q P_{\text{optical}}}{h\nu WL} \tau$$
Photoconductor Impulse Response (bandwidth)

- **Carrier recombination time**
  \[ n(t) = n_0 e^{-t/\tau}, \quad p(t) = p_0 e^{-t/\tau} \]

  Radiation damaged GaAs
  reduces \( \tau \) and \( \mu \): increases bandwidth, decreases sensitivity.

  Low-Temperature-Growth GaAs:
  reduces \( \tau \) but maintains good mobility, hence good sensitivity.
  (Arsenic clusters as recombination centers?)

- **Gap capacitance charging time**

  Small-signal (\( GZ_0 << 1 \)):
  \[ \tau = 2Z_0 C_{gap} \]
Photodiodes vs Photoconductors

Photodiode

- Contacts block carrier entry into drift region.
- Total charge collected is exactly # photons absorbed.
- Current duration is electron + hole transit times.

Photo-conductor

- N contacts block hole entry, do not block electron entry.
- By charge neutrality, more electrons enter from cathode as holes leave.
- Current pulse duration is hole recombination or sweep-out time.
- Current gain is (hole lifetime)/(electron lifetime).

N-I-N photoconductor (vice-versa for P-I-P photoconductor)

N contacts block hole entry, do not block electron entry.
Metal-Semiconductor-Metal Photodetectors

Unless Ohmic contacts are used, MSMS are photodiodes, not photoconductors

Published Ultrafast MSM detectors show no photoconductive gain

Observed change in responsivity with bias due to increased depletion depth?
The Step Recovery Diode

Under bias, carriers are stored in the intrinsic region

Charge control model:

• Stored charge
  \[ Q_s = Q_0 \left( \exp \left( \frac{qV}{kT} \right) - 1 \right) \]

• Diode current
  \[ I = \frac{dQ_s}{dt} + \frac{Q_s}{\tau} \]

• Widely used as a pulse generator in microwave instruments

stored carrier concentration
Electrical Falltime Compression with SRD's

\[ I = \left( \frac{V_{diode} - V_{in}}{R} \right) \]

\[ dQ/dt \approx -I \]

For a fast-changing input signal, the SRD acts as a nonlinear capacitor.
Transition time limits of SRD's

Depletion Capacitance:
\[ \tau = R_{load} C_{SRD} \]

Carrier Diffusion Time
Time in which minority carriers are swept from the intrinsic region. No simple closed-form expression. Moll (1969) estimates \( \approx 10 \text{ ps/\mu m} \) of intrinsic layer thickness.

![Commercial SRD Impulse Generator](image_url)
Schottky Diodes

Forward bias: nonlinear conductance

Reverse Bias: Nonlinear Charge Storage Element

\[
I(V) = I_s \left( e^{qV/kT} - 1 \right) \quad \text{so if } v(t) = v_1 e^{j\omega_1 t} + v_1 e^{j\omega_2 t}, \text{ then, } I(t) = \sum_{l,m} I_{l,m} e^{j(\omega_1 + m\omega_2) t}
\]

\[
Q(V) = C_0 V + C_1 V^2 + \ldots
\]

• Sum and difference-frequency generation ("mixing"), frequency multiplication.
• Schottky diodes are fastest semiconductor devices, perform as mixers to \( \approx 10-20 \text{ THz} \)

28
"Canononical" mixer and frequency multiplier are shown.

Millimeter-wave implementations in waveguide or on-wafer, submillimeter-wave versions in quasi-optical form.

Mixing and harmonic generation at low THz frequencies.

Circuits shown are narrowband.
Schottky Diode Structure and Parasitics

\[ f_{\text{diode}} = \frac{1}{2\pi C_{\text{diode}} \left( R_{N^-} + R_{N^+} + R_{\text{contact}} \right)} \]
Exponential Hyperabrupt Varactor Diodes

Doping Profile

Normalized Capacitance-Voltage Curve

\[ V_{\text{reverse}} + \phi = \int_{0}^{x_{\text{depletion}}} \left( \frac{qN_d(x)}{\varepsilon} \right) dx \quad , \quad C = \frac{\varepsilon A}{x_{\text{depletion}}} \]

• Hyperabrupt profiles increase capacitive change but decrease cutoff frequency and reverse breakdown voltage.
Bandwidth of Uniform Varactors vs Geometry

Large-Signal Cutoff Frequency, THz

Surface Doping, x10^{17}/cm^3

Avalanche Breakdown Voltage

W=0.5 \mu m, D=0.5 \mu m
W=0.5 \mu m, D=1 \mu m
W=1 \mu m, D=1 \mu m
W=1 \mu m, D=2 \mu m
W=2 \mu m, D=2 \mu m
Bandwidth of Hyperabrupt Varactors vs Geometry

Schottky Diode Bandwidth vs Geometry

Large-signal cutoff frequency, THz

Surface doping, $N_0 \times 10^{17}/cm^3$

Avalanche breakdown voltage

$qN_0 x_0^2/\varepsilon = 14 \text{ V}$

$V_{br}$

$\lambda = 1 \mu m$

$\lambda = 2 \mu m$

$\lambda = 3 \mu m$
THz Mixing and Detection with Schottky Diodes

Whisker contact as antenna coupling to Schottky diode

Mixing of 33rd harmonic of 74 GHz Klystron (2.44 THz) with 119 μm laser radiation

## Antenna-Coupled Mott Mixer Diodes

<table>
<thead>
<tr>
<th><strong>Objectives</strong></th>
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<tr>
<td>• Low-noise focal-plane array receiver/downconverters for 184/203/2500 GHz radiometers</td>
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<tr>
<td>• Advanced mixer diode technology integrable with diode LO frequency multipliers</td>
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<table>
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<tr>
<th><strong>Approach</strong></th>
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<tr>
<td>• Antenna-coupled diode SMMIC</td>
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<tr>
<td>• Low capacitance: 0.05 μm² contact</td>
</tr>
<tr>
<td>• Low transit time: 150Å depletion layer</td>
</tr>
<tr>
<td>• Low resistance: N++ / I / metal diode</td>
</tr>
<tr>
<td>• 12 THz circuit bandwidth</td>
</tr>
<tr>
<td>• ≈20 THz diode bandwidth</td>
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</tbody>
</table>
Resonant Tunneling Diodes

(Sollner, MIT Lincoln Labs)

(Esaki)

Strong current flow when electron reservoir and confined state are aligned in energy.

Narrower AlAs barriers decrease the electron trapping time, increasing $\Delta E$, and increasing the peak current density. Increasing the emitter doping increases the electron supply and therefore also increases the current density.
Parameters Determining RTD Fmax

\[ f_{\text{max}} = \frac{1}{2\pi R_n C} \sqrt{\frac{R_n - R_s}{R_s}} \]

\[ \frac{1}{R_n C} = \frac{1}{dE} \int dJ \]

-Maximized by designing for high current density (reliability limits)

-Rs then limits \( f_{\text{max}} \):
InAs/AlSb RTD's have very low (10^{-7} \( \Omega \)-cm^2) contact resistance, & have attained \( f_{\text{max}} = 1.3 \) THz.
RTDs as negative-resistance oscillators

Coupled to the conjugate impedance \(-jX(\omega)\), the RTD will oscillate if \(R(\omega) < 0\)

\[
Z(\omega) = R_s + \frac{-R_n}{1 - j\omega R_n C} = R(\omega) + jX(\omega)
\]

\[
R(\omega) = R_s - \frac{R_n}{1 + \omega^2 R_n^2 C^2}
\]

\[
R(\omega_{\text{max}}) = 0
\]

\[
\omega_{\text{max}} = \frac{1}{R_n C} \sqrt{\frac{R_n - R_s}{R_s}}
\]

RTD oscillators have been demonstrated to 712 GHz
Well Transmission Probability vs Energy

AlAs/GaAs, 45Å well
Quantum Well Inductance

Transit time through well:

\[ \tau = \frac{2\hbar}{\Delta E} \]

\[ \Delta E \] is the F.W.H.M. of the transmission probability

Equivalent inductance:

\[ L_W = \frac{\Delta E}{G_N} \]

• In the negative resistance region \( G_N \) and \( L_W \) are negative.
• \( f_{\text{max}} \) is decreased
• Space charge transit time \( \tau_t = \frac{d_{\text{space-charge}}}{v_{\text{electron}}} \) also effects \( f_{\text{max}} \), but with \( d_{\text{space-charge}} \sim 500 \text{\AA} \), \( v_{\text{electron}} \gg 10^7 \text{cm/sec} \) and \( \tau_t \sim 100 \text{fs} \)
Schottky-Collector Resonant Tunnel-Diodes

Top ohmic contact is eliminated. Decreased Rs, increased fmax
Bandwidth of Submicron Schottky-Collector RTDs

RTDs vs. SRTDs in AlAs/GaAs and In$_{0.53}$Ga$_{0.47}$As/AlAs

- Scaling to submicron dimensions increases RTD periphery/area ratio
- Periphery-dependent parasitic resistance terms driven towards zero (bottom ohmic contact resistance, buried N+ layer)
Schottky-Collector RTDs for 0.3-3 THz Oscillators

0.12-μm AlAs/InGaAs/InP device

- Schottky (metal) electron collector, 0.1 -μm geometry.
- Greatly reduced series resistance, large increase in \( f_{\text{max}} \)

AlAs/GaAs SRTDs: 900 GHz \( f_{\text{max}} \) estimated
InGaAs/AlAs SRTDs: 2.2 THz \( f_{\text{max}} \) estimated

Application: 0.3-1.5 THz quasi-optical power oscillator arrays
RTD Switching Speed Considerations

Governing law is
\[ I = C \frac{dV}{dt} \]

Even if \( R_s = 0 \) (infinite \( f_{max} \)), risetime is not zero

For fast risetimes, need high RTD \( I_{peak}/C \) ratio. 8 ps risetimes in AlGaAs \((\approx 10^5 \text{ A/cm}^2\text{current density})\), 1.8 ps in InAs/AlSb \((3.5 \cdot 10^5 \text{ A/cm}^2)\)

\[ T_{10\%-90\%} = \int_{0.9V_{\text{initial}} + 0.1V_{\text{final}}}^{0.1V_{\text{initial}} + 0.9V_{\text{final}}} C dV / \Delta I(V) \]

\[ \approx 3 - 5R_n C \]
3-Terminal Devices (Transistors)
Heterojunction Field-Effect Transistors

Current state of art:

InAlAs/InGaAs / InP HEMTs, 0.1 μm gate (R&D device)
250-500 GHz \( f_{\text{max}} \), 2-3 Volt breakdown
1 dB noise figure amplifiers at 65 GHz

AlGaAs/InGaAs/GaAs HEMTs, 0.25 μm gate (mass production device)
100-150 GHz \( f_t \), 100-150 GHz \( f_{\text{max}} \), 5-8 Volt breakdown
50 GHz \( f_t \), 180 GHz \( f_{\text{max}} \), 10-15 Volt breakdown

Circuits:

5-10 dB gain 150 GHz monolithic narrowband (low-noise) amplifiers
5 dB gain, 1-155 GHz broadband amplifiers (Agarwal et. al. 1998 MTT)

Monolithic mm-wave ICs for radar (mixers, phase shifters, preamps, power amplifiers,...)
Device Circuit Models vs Device Physics: FET as example

\[ I_d = \frac{Q}{l} v_{sat}, \text{ where } \frac{dQ}{dV_{gs}} = C_{gs} \text{ and } \frac{dQ}{dV_{gd}} = C_{d-ch} \]

so:

\[ g_m = \frac{dl_d}{dV_{gs}} = C_{gs} \frac{v_{sat}}{l} = C_{gs} \frac{v_{sat}}{l} \text{ gate} \quad \text{and} \quad G_{ds} = \frac{dl_d}{dV_{ds}} \approx C_{d-ch} \frac{v_{sat}}{l} = C_{d-ch} \frac{v_{sat}}{l} \text{ gate} \]
Heterojunction Field-Effect Transistors

\[ \omega_\tau = \frac{g_m}{C_{gs}} = \frac{v_{sat}}{I_g} \]

\[ \omega_{\text{max}} \approx \frac{\omega_\tau}{2\sqrt{\left(R_g + R_i + R_s\right)/R_{ds}}} \]

current gain cutoff frequency

POWER gain cutoff frequency
• Reducing gate length by lithographic scaling decreases the carrier transit time (Cgs/gm ratio) and increases device bandwidth
• Use of Schottky vertical contact essential if RC time constants are to scale (Schottky MESFET vs PN JFET)
• Gate length must be 5-10 times gate-channel separation to screen channel charge from drain potential
• Minimum gate channel separation (tunneling limit) sets limit on gate length scaling.
Heterojunction Bipolar Junction Transistors

GaAs/AlGaAs devices: $f_{\text{max}} \approx 200$ GHz
AllInAs/GaInAs devices: $f_{\text{max}} \approx 70$-200 GHz (but ...*)
InP/GaInAs devices: $f_{\text{max}} \approx 220$ GHz for $f_t \approx 220$ GHz
Si/SiGe devices: $f_{\text{max}} \approx 70$ GHz (production), 180 GHz (mesa device)

HBTs have generally poorer $f_{\text{max}}$ and noise figure than HEMTs, but are more predictably-behaved devices more suitable for higher-complexity analog and mixed analog-digital applications.

Representative circuit result: 37 GHz static frequency dividers (e.g. master-slave flip-flops), DC-40 GHz amplifiers, chip-sets for 40 Gbit/s fiber transmission, A-D converters at a few Gigasamples/sec.
HBT Device Structure

Note the indicated base resistance and base-collector capacitance.
HBT DC I-V characteristics

\[ n_p(0) = qN_c e^{-qV_{electron}/kT} \propto e^{+qV_{be}/kT} \]

\[ p_n(0) = qN_v e^{-qV_{hole}/kT} \propto e^{+qV_{be}/kT} \]

electron concentration at emitter edge of base

hole concentration at base edge of emitter

heterojunction makes this small

electron current from emitter to collector

electron current from emitter to collector

\[ I_{electron} = \frac{D_n qA e N_c}{T_b} e^{-qV_{electron}/kT} \propto e^{+qV_{be}/kT} \]

\[ g_m \equiv \frac{dI_c}{dV_{be}} = \frac{qI_c}{kT} \]

transconductance
HBT Stored Charge & Diffusion Capacitance

Base Transit Time

<table>
<thead>
<tr>
<th>N</th>
<th>I</th>
<th>P</th>
<th>N- or I</th>
<th>N+</th>
</tr>
</thead>
</table>

Charge densities

\[ n_p(0) = qN_e e^{-qV_{be}/kT} \]

electron concentration at emitter edge of base

\[ I_{electron} = qn_p(0)D_n / T_b \]

electron current from emitter to collector

\[ Q_{base} = QA_e n_p(0)T_b / 2 = I_{electron}T_b^2 / 2D_n = \tau_b I_{electron} \]

stored base charge

"Diffusion Capacitance"

\[ C_{diffusion} = \frac{dQ_{base}}{dV_{be}} = \frac{dQ_{base}}{dI_c} \frac{dI_c}{dV_{be}} = (\tau_b + \tau_c)g_m \]

\[ C_{be, diffusion} = g_m(\tau_b + \tau_c) \]

tific capacitance between base & emitter

modelling charge storage

Collector Transit Time

<table>
<thead>
<tr>
<th>N</th>
<th>I</th>
<th>P</th>
<th>N- or I</th>
<th>N+</th>
</tr>
</thead>
</table>

\[ \delta Q_{base} = \delta Q_{space-charge} / 2 \]

change in base stored charge

\[ \delta Q_{base} = \delta Q_{collector} = \delta Q_{space-charge} / 2 \]

\[ \delta Q_{base} = \delta I_{collector} (T_c / 2 V_{sat}) = \tau_c \delta I_{collector} \]

depletion-layer space-charge
Bandwidth of Bipolar Transistors

\[ f_{\tau} = \frac{1/2\pi}{\tau_{\text{base}} + \tau_{\text{collector}} + \left(\frac{C_{je}kT}{qI_e}\right)} \]

\[ f_{\text{max}} = \sqrt{\frac{f_{\tau}}{8\pi R_{bb} C_{cbi}}} \]

- \( f_{\tau} \), \( f_{\text{max}} \), and \( C_{cbx} \) are all important for high-speed circuits.

- \( R_{bb}C_{cbi} \) and \( R_{bb}C_{cbx} \) must be reduced.
HBT Design Tradeoffs

Thin base & collector depletion layers result in small transit times, high ft.

Thin base & collector result in high base resistance, high c-b capacitance ...low fmax!

There are thus optimum thicknesses for these layers.

Very high base doping used to minimize base sheet and ohmic contact resistance.

Etch & implant techniques used to reduce extrinsic CB capacitance.
Lithographic Scaling & Transistor Bandwidth

- Transistor bandwidth improves as dimensions are reduced
  - 0.1 μm HEMTs
  - 0.25 μm CMOS (VLSI)

- HBTs typically built at ≈1 μm lithography
  - smaller devices not generally faster

- Goal: a scalable HBT
Why are HEMTs smaller & faster than HBTs?

- **FETs have deep submicron dimensions.**
  - 0.1 \( \mu m \) HEMTs with 400 GHz bandwidths (satellites).
  - 5 million 1/4-\( \mu m \) MOSFETs on a 200 MHz, $500 CPU.
  - FET lateral scaling decreases transit times.
  - FET bandwidths then increase.

- **HBTs, RTDs, Schottky diodes have \( \approx 1 \ \mu m \) junctions.**
  - Vertical scaling decreases electron transit times.
  - Vertical scaling increases RC charging times.
  - Lateral scaling should decrease RC charging times.
  - HBT & RTD bandwidths should then increase.

But, HBTs & RTDs must first be modified . . .
Excess Collector-Base Capacitance in Mesa HBTs

- Collector-base capacitance independent of emitter width
- Base resistance independent of emitter width for < 1µm
- $f_{\text{max}}$ does not improve for emitter stripe widths < 1µm
Transferred-Substrate HBTs

- Flip-chip process: narrow collector
- Narrow collectors feasible $\rightarrow$ large decrease in $R_{bc}$
- Consequent large increase in bandwidth
- Submicron collector and emitter scaling $\rightarrow f_{max} \approx 700$ GHz
Transferred-Substrate HBTs: a **Scalable** HBT technology

- Collector capacitance reduces with scaling: \( C_{cb} \propto W_e \)
- Bandwidth increases rapidly with scaling: \( f_{\text{max}} = \sqrt{1/W_e} \)

\[
\begin{align*}
C_{cb} & \propto W_e \\
f_{\text{max}} & = \sqrt{1/W_e}
\end{align*}
\]
Transferred-Substrate HBT Process

Objectives:
- 500 GHz transistor bandwidth
- Thermal management for high power density
- Low wiring & packaging parasitics at 100+ GHz

Approach:
- BCB process: standard IC materials
- Metal substrate, thermal vias
- Microstrip wiring: ground vias backside ground plane $\varepsilon_r=2$: low capacitance

1) Normal emitter, base processes. Deposit silicon nitride insulator.
2) Coat with BCB polymer. Etch vias.
3) Electroplate with gold. Die attach to copper substrate.
AlInAs/GaInAs graded base HBT

Band diagram under normal operating voltages
$V_{ce} = 0.9 \, \text{V}, \, V_{be} = 0.7 \, \text{V}$

- 500 Å 5E19 graded base ($\Delta E_g = kT$), 3000 Å collector
Transferred-Substrate Heterojunction Bipolar Transistor

Device with 0.6 $\mu$m emitter & 0.8 $\mu$m collector
extrapolated $f_{\text{max}} = 560$ GHz (?)

0.25 $\mu$m devices should obtain 600-700 GHz $f_{\text{max}}$
Transferred-Substrate Heterojunction Bipolar Transistor

Device with 0.6 μm emitter & 1.8 μm collector extrapolated fmax at instrument limits

\[ f_{\max} = 470 \text{ GHz} \]
\[ f = 215 \text{ GHz} \]

0.25 μm devices should obtain 600-700 GHz fmax
In Fabrication: Deep Submicron HBTs

0.15 µm emitter

base-emitter diode

IV Characteristics

In development: 0.15 µm emitter/ 0.3 µm collector HBT

Credit: Michelle Lee, Dino Mensa, UCSB; S. Martin, R.P. Smith, JPL
Transferred-Substrate HBT Integrated Circuits

11 dB, 50+ GHz AGC / limiting amplifier

47 GHz master-slave flip-flop

10 dB, 50+ GHz feedback amplifier

7 dB, 5-80 GHz distributed amplifier
Transmission Lines

Geometries

Characteristic Impedances

Group Velocity Dispersion

Skin effect losses

(substrate) radiation losses

Excitation of undesired modes & resulting problems
Types of Planar Transmission Lines

Well designed CPW and CPS have 
\((S+2W)\ll h\) 
(better than 2:1 ratio)

Coplanar Waveguide (CPW)

Slotline: Non-TEM, dispersive

Coplanar Strips (CPS)

Microstrip
Microstrip Line

Dominant transmission medium in microwave IC's

Key advantage: IC interconnections have very low ground lead inductance—more important than signal line inductive parasitics in amplifiers

Key problem: through-wafer grounding holes (vias). Via inductance forces progressively thinner wafers at higher frequencies. Microstrip is used with good performance in 65 GHz monolithic circuits
Ground bounce noise must be 98 dB below full-scale input. Differential input will partly suppress ground noise coupling. ~30 to 40 dB common-mode rejection feasible. CMRR insufficient to obtain 98 dB SNR.

Eliminate ground bounce noise by good IC grounding.
Microstrip IC wiring to Eliminate Ground Bounce Noise

Brass carrier and assembly ground

IC with backside ground plane & vias

interconnect substrate

IC vias eliminate on-wafer ground loops

near-zero ground-ground inductance
Skin Effect Losses, I

\[ \gamma_{metal} = \sqrt{j \omega \mu (j \omega \varepsilon + \sigma)} \]

\[ \approx \sqrt{j \omega \mu \sigma} \]

\[ \alpha_{metal} + j \beta_{metal} = \sqrt{\omega \mu \sigma / 2} + j \sqrt{\omega \mu \sigma / 2} \]

\[ = \left( \frac{1}{\delta} \right)(1 + j) \]

where \( \delta = \sqrt{2/\omega \mu \sigma} \)

\[ Z_{series} = \frac{\gamma_{metal}}{\sigma P} \]

\[ = \left( \frac{1}{\delta \sigma P} \right) + j \left( \frac{1}{\delta \sigma P} \right) \]

Surface impedance of the metal interconnections of a transmission line introduces loss proportional to the square root of frequency.

Dispersion is also introduced, as the skin impedance has equal real and complex parts.
For a coplanar line the effective current carrying periphery \( P \) is approximately the width of the center conductor (if \( S \) is relatively small compared to \( W \), a higher-impedance line)

Following this, the line propagation constant \( \gamma \) can be found, and the transfer function for a line of length \( l \) is

\[
Z_{\text{series}} = \left( \frac{1}{\delta \sigma P} \right) + j \left( \frac{1}{\delta \sigma P} \right)
\]
Skin Effect Losses III

The impulse response of the transmission line can then be found. (Wiginton and Nahman, Proc. IRE, February 1957)

Skin effect causes pulse broadening proportional to distance

\[ h(t) \equiv C \ast U(t/\tau) \left( \frac{t}{\tau} \right)^{-3/2} \exp\left( -\frac{\tau}{t} \right) \]

\[ \tau = \left[ l \sqrt{\frac{\mu}{\sigma}} \frac{1}{4 Z_0 P} \right]^2 \]

Skin effect impulse response
Skin Effect Losses IV

The step response is the integral of the impulse response. Note the initial fast rise and the subsequent "dribble-up" characteristic of skin effect losses.

![Skin Effect Step Response](image)

still hasn't reached 0.9 Volts !
Transmission-Line Radiation Losses

Transmission line velocity is

\[ v = \frac{c}{\sqrt{1 + \varepsilon_r}} \]

Velocity of a plane wave in the substrate is \( v = c\sqrt{\varepsilon_r} \), which is slower.

Power radiates at angle \( \psi \) determined by matching \( K_z \).

With substrate of finite thickness, radiation shows frequency structure due to substrate modes.
Loss (in dB) per wavelength is proportional to frequency$^2$ and to the square of the transverse dimensions of the line

Experimental confirmation -scale model measurements

From Rutledge et al (see reference list)
Skin and Radiation Losses, 50Ω CPW

\[ \alpha_{\text{skin}} \equiv k_1 \sqrt{\varepsilon_r + 1} \sqrt{f / D} \]

\[ \alpha_{\text{rad}} \equiv k_2 \varepsilon_r^{3/2} f^3 D^2 \]
On-wafer interference from line radiation

**Without absorber**

Oscillations are 220 GHz slab modes coupled from other points on the circuit.

**With absorber**

CPW #2

CPW #1

substrate

Wafer Chuck

CPW #2

CPW #1

microwave absorber

Wafer Chuck

72
Transmission Line Parasitic Modes

Nominal Coplanar Waveguide

- Ground
- Signal
- Ground
- Substrate
- Backside ground plane (intentional)
  or wafer chuck (accidental)

Microstrip mode

Nominal Coplanar Strips

- +Signal
- -Signal
- Substrate
- Backside ground plane (intentional)
  or wafer chuck (accidental)

Coplanar strip or slot mode

Coplanar waveguide mode

- Total number of quasi-TEM modes is one less than # of conductors
- Care must be taken to avoid excitation of parasitic modes
- Unexpected results will otherwise arise...
General Rules for Avoiding Parasitic Modes

1) Where do the currents flow?

2) Which conductors have what voltages for which modes?

Be aware that:

- currents must flow in the ground planes of unbalanced transmission lines. The currents flow close to the edge of the ground plane nearest the signal conductor.

- there are equal and opposite voltages on the 2 conductors of balanced transmission lines. This seriously restricts the types of junctions allowable.
Example of parasitic mode excitation

Both pulse generator and sampling gap excite (sample) mixed microstrip and CPS mode. These propagate at different velocities and will separate in time as they propagate.
Example of parasitic mode excitation

CPS line

transistor gives voltage gain $-A_v$

$+V/2$

$-A_vV/2$

$-V/2$

$+A_vV/2$

$+V/2$

$+V/2$

$+V/2$

$+V/2$

$+V/2$

$+V/2$

- potentials don't match!
- microstrip modes must be excited to equalize potentials.
- circuit load impedances include contributions from microstrip modes
- resonances will result from boundary conditions on microstrip modes

• CPS line is a balanced line with no ground connection. It cannot be used when a common-lead connection (ground) is needed, e.g. in testing a 2-port device.
• microstrip or CPW should be used in this application
Example of parasitic mode excitation

CPS line

photoconductor

CPS line

$+V/2$

$-V/2$

airbridge

CPW line bringing in photoconductor bias

0 V $+V_{bias}$ 0 V

potentials can be reconciled at this point only through excitation of microstrip modes on both the CPS and CPW lines

• CPS is usually a bad idea...
Example of parasitic mode excitation

The problem can be fixed by strapping the junction with air-bridges.
Transmission Lines: In summary

• **Radiation Losses**
  (Coplanar waveguide as example)

\[ \alpha_{\text{radiation}} (dB / mm) \propto (S + 2W)^2 f^3 \]

- Line impedance constrains S/W
- Narrower lines are better

• **Skin Losses:**

\[ \alpha_{\text{skin}} (dB / mm) \propto (W)^{-1} f^{1/2} \]

- Wider lines are better

For any frequency, there is an optimum W for lowest loss. Lower \( \varepsilon_r \)
substrates have lower radiation losses. Alternatives are air-bridge CPW, etched-substrate CPW

Semi-rigid coaxial cable: lower loss than on-wafer lines
cordless bandwidths to 65 GHz. 130 GHz connectors are in development.
Antennas
Picosecond On-Wafer Antennas

Frequency-independent antennas

- tapered CPS ("V" antenna)
- CPS feed
- bowtie antenna (CPW feed)

These are all travelling-wave antennas. Radiation impedance, group delay, and far-field patterns are all frequency-independent above the antenna cutoff.

Resonant antennas

- dipole antenna (CPS feed)
- slot antenna (magnetic dipole)

- metal
- substrate

all these require substrate lenses

Other monolithic antennas:

- log periodic (not truly frequency independent)
- exponential spiral (frequency-dependent polarization)

See Rutledge reference
- Energy primarily radiated into substrate
- Trapped (Snell's law) into substrate slab modes
- Substrate lens allows radiation to escape, but efficiency still relatively poor for high-\(\varepsilon\) substrates. Hyperhemisphere gives some collimation
- Lens must be several wavelengths diameter at longest wavelength of interest.
Radiation Distribution on tapered CPS antenna

- Length ($\Delta z$) of radiation distribution proportional to wavelength.
- Width $W$ proportional to length $\Delta z$, radiating area proportional to $\lambda^2$.
- Far-field radiation pattern independent of wavelength.
Circuit Design

• Summary of fundamental gain-frequency limits.
• A few examples of high-speed circuits

The intention here is to give an overview of

...how active devices are modeled

...relevant device figures of merit & their influence on circuit performance

...this is of importance to researchers involved in optical probing of electronic devices: what measurements provide useful descriptive information to the device physicist and circuit designer?
Device Figures-of-Merit

Top: simplified FET equivalent-circuit model

Center: definition of short-circuit current gain ($I_{out}/I_{in}$)

Current-gain cutoff frequency:
$$\left| \frac{I_{out}}{I_{in}} \right| = 1 \text{ at } f = f_{T}$$

obtaining the maximum power gain requires impedance-matching on input & output.

Power gain cutoff frequency
$$\left| \frac{P_{out}}{P_{in}} \right| = 1 \text{ at } f = f_{\text{max}}$$

Expressions for $f_{T}$ and $f_{\text{max}}$ are given elsewhere in these notes...
Device Figures of Merit

Gains, dB

Frequency (log scale)
f_τ f_max

short-circuit current gain

maximum available power gain

generator match transistor match load

\[
2\pi f_t = \frac{g_m}{C_{gs}} = \frac{1}{\tau_{gate}} \rightarrow \text{depends on carrier transit times}
\]

\[
2\pi f_{\text{max}} = \frac{f_\tau}{2} \sqrt{\frac{R_{ds}}{R_i}} \rightarrow \text{also depends on parasitics}
\]
Resistively-Loaded Transistor Circuits

- Broadly representative of most analog high-frequency amplifiers.
- Bandwidth limited by capacitance-charging time of Cgs.
- Big transistor gives big gm, big Cgs, hence big gain, small bandwidth.
- Small transistor gives small gm, small Cgs, hence small gain, small bandwidth.

Assuming that: (1) Generator and load impedances are equal and (2) \( f_{\text{max}} \gg f_t \) (\( R_i \) and \( R_{ds} \) have small effect)...

Gain-bandwidth product given by ratio of gm to Cgs, eg. by \( f_t \)!
Resistively-Loaded Transistor Circuits 2

- Big transistors give large gain, low bandwidth
- Small transistors give small gain, high bandwidth
- Gain-bandwidth product limited to transistor short-circuit current-gain cutoff frequency if
  (1) Equal generator and load resistances
  (2) $f_{max}$ much bigger than $f_t$
Classes of Circuits: Resistively Loaded

- Advantage: simple elementary circuit --> suitable for building complex ICs
- Limitation: gain-bandwidth products << transistor power-gain cutoff frequency
Representative Resistive-Loaded Circuits

Broadband AGC amplifier for fiber-optic receiver

Stage gain-bandwidth products at or below $f_t$. 

Broadband single-stage resistive feedback amplifier
**Reactively-Matched Transistor Circuits 1**

- Tuning networks on transistor input and output provide maximum power transfer, power gain obtained is maximum available from device.

  *Tuning (impedance-matching) network are fundamentally narrowband (Fano's inequality)*

Maximum available gain is obtained from the transistor, but only over a small frequency range.
Transistor maximum available power gain obtained over narrow bandwidth

Acceptable for narrow-band applications, not for broadband or digital
Classes of Circuits: Reactively Tuned

Advantage: gain equal to maximum available over maximum bandwidth of \( f_{\tau} \) and \( f_{\max} \)

Limitations: bandpass characteristics, large dimensions of tuning networks
A ladder network of Ls and Cs acts as a transmission line of impedance

$$ Z_o = \sqrt{\frac{L}{C}} $$

...for frequencies below its Bragg frequency

$$ f_{Bragg} = \frac{1}{\pi \sqrt{LC}} $$

• Transistor capacitances absorbed into synthetic transmission lines.
• Transistor resistance introduce transmission-line losses.

• Distributed networks eliminate capacitance charging time as performance limit
• Line losses (caused by transistor fundamental, power-dissipating parasitics)-limits performance by introducing line losses.
FET traveling-wave amplifier

- FET capacitances absorbed into synthetic lines.
- Capacitance charging times eliminated
- Feasible gain-bandwidth determined by gate, drain line losses
- Losses result from FET resistive parasitics.
- These resistive parasitics also determine FET fmax.

- Under idealized conditions, circuit gain-bandwidth product approaches transistor fmax.
• Gain-bandwidth product can approach transistor power-gain cutoff frequency
• Somewhat idealized discussion: Real TWAs often limited somewhat below f_{max} if one is not free to choose generator & load impedance independently
• Gain-bandwidth very close to f_{max} always obtainable by capacitive division.
Classes of Circuits: Distributed

Advantage: gain-bandwidth products equal to transistor $f_{\text{max}}$

Limitation: size and complexity, signal delay renders unusable for logic, feedback
Capacitive-Division Traveling-Wave Amplifiers

Input

Output

Normal

Capacitive division

\[ C_{\text{div}} = C_{\text{gs}} \]

Peripheral = W

Peripheral = 2W

same transconductance,
same input capacitance,
but 2:1 improved series input resistance: increases bandwidth

Capacitive Division Increases Gain-Bandwidth

Measured Results: First Design Iteration

Forward Gain, \( S_{21}, \text{dB} \)

Frequency, GHz

10 dB gain, 92 GHz bandwidth

8 dB gain, 98 GHz bandwidth

0.15 \( \mu \text{m} \)

InGaAs/InAlAs HEMT

Insertion Gain, \( S_{21}, \text{dB} \)

Frequency, GHz

10 dB gain, 92 GHz bandwidth

8 dB gain, 98 GHz bandwidth
155 GHz HEMT Distributed Amplifier

UCSB: design, test
Hughes: InP HEMT IC technology

(capacitive division)
**$F_t$ multiplier: Principle of Operation**

**Simple Stage**

\[
\frac{I_{\text{out}}}{I_{\text{in}}} = \frac{1}{j} \left( \frac{f_t}{f} \right)
\]

---

**$f_t$ Doubler**

\[
\frac{I_{\text{out}}}{I_{\text{in}}} = \frac{1}{j} \left( 2 \cdot \frac{f_t}{f} \right)
\]

*Inputs connected in series, outputs connected in parallel, output currents add, current gain is twice that of single stage, $f_t$ is doubled*
$f_t$ Doublers: Implementation

Differential Pair

Single-ended (and emitter-follower)

“mirror doubler”
Circuit Gain-bandwidth Limits: $F_t$ and $F_{\max}$

Lumped circuits: limited by $f_{\max}$ and $f_t$

Distributed circuits: limited only by $f_{\max}$

$f_t$ multipliers: also limited only by $f_{\max}$
Observations: High Frequency Circuit Design

Current-gain cutoff frequency
determined by carrier transit times

Power gain cutoff frequency
also determined by RC charging parasitics

Reactively tuned circuits: classical microwave design
narrow bandwidths, big circuits

Distributed circuits:
“optimal”, but complexity & delay serious limits

Resistively loaded circuits: classical analog design
circuit bandwidths below transistor limits

ft-multiplier brings bandwidths close to fmax-limit
Nonlinear Wave Propagation Devices

Shock-wave nonlinear transmission lines
NLTL-gated diode sampling circuits
Soliton NLTL impulse compressors & frequency multipliers
Traveling-Wave RTD pulse generators

These devices exploit distributed circuit principles presented above
General principle: use of distributed circuit yields performance determined by fundamental limits of the semiconductor device, rather than capacitance charging times.
NLTL Technology

Subpicosecond electronic pulse generation & detection

Principles of operation: nonlinear wave propagation

Basis of performance: THz bandwidth of Schottky diodes

- Shock-wave NLTLs
- Soliton NLTLs
- Diode sampling circuits

Bandwidths now ≈ 700-1000 GHz, ≈ 0.3–0.5 ps pulses
Technological limit is at ≈ 2-3 times greater bandwidth.
Electrical Sampling Circuits
Used in Sampling Oscilloscopes, TDRs, and network analyzers

2 Diode Sampling Bridge

\[ R\cdot C \text{ Risetime at Input} \]
\[ T_{RC} = 2.2 \frac{Z_0}{2C_{Diode}} \]

\[ \text{Aperture Risetime} \]
\[ T_{on} \]

\[ \text{Total Effective Risetime} \]
\[ T_{eff} = \sqrt{\frac{T_{on}^2 + T_{RC}^2}{2}} \]

Schottky diodes are readily made with \( \approx 5 \) fF junction capacitance and \( \approx 2 \) THz R-C cutoff frequencies. The primary bandwidth limitation of sampling circuits is thus the duration of the strobe pulse used to gate the diodes. Previously, silicon step-recovery diodes (\( \approx 25 \) ps \( T_f \)) had been used.
Structure and Equivalent Circuit of NLTL
Wavefront Compression by NLTL

\[ T_{\text{delay}} = \sqrt{L(C_{\text{diode}} + C_{\text{line}})} \]
SPICE Simulation of Shock formation
Limits to NLTL Shock-Wave Transition Time

• Periodic-Network (Bragg) Frequency
The periodic structure results in a sharp filter cutoff inversely proportional to the diode spacing. Within lithographic limits, this can easily be 1-2 THz.

\[ T_{\text{Isolation}} = \frac{1}{f_{\text{Diode}}} \]

1 m

• Diode Cutoff Frequency
The fundamental limit of the technology.
Falltime limited to \( T_f f_{\text{diode}} = 1.4 \, ps \cdot THz \)
10 THz diodes: 0.14 ps predicted shock-waves
Shorted-Line Differentiator for Impulse Generation

Symmetric Impulse Generation with CPW/CPS Balun/Differentiator

Input Voltage

Positive Strobe

Negative Strobe

\[ V_{in} \]

\[ Z_0 \]

\[ C_1 \]

\[ C_2 \]

\[ 2Z_0 \]

\[ V_p/2 \]

\[ t \]
Implementation of the shorted line balun-differentiator exploits the deliberate excitation of CPS modes at the junction between two CPWs.
NLTL-Strobed Sampling Circuit

Sampled outputs

Input

CPW/CPS balun/differentiator

NLTL strobe pulse compressor
NLTL output measured by sampling circuit

Measured falltime, 1.8 ps, 10%-90%
Increasing NLTL bandwidth

- need increased diode cutoff frequency
- need increased Bragg Frequency.
  - small diodes at small spacings then causes problems:
    - narrow transmission lines with very high skin effect losses
    - diode spacing comparable to diode's physical size
    - diode-to-transmission line junction comparable to diode spacing
      hence large junction (pad) parasitics
- difficulties mitigated using elevated coplanar waveguide
  with top-contacted diodes
  - high transmission line velocity increases diode spacings
  - low CPW dielectric constant: wide conductor, lower skin losses
  - diode contacted from TOP, junction (pad) parasitics eliminated
Elevated CPW

Transmission-line fields primarily in air.

Higher velocity.

Wider conductor needed for given $Z_0$, hence lower skin-effect losses

Radiation losses lower (??) because of reduced field in substrate.

Less useful at lower frequencies: to be effective, conductor elevation must be comparable to conductor width. Lower-frequency (non-elevated) CPW can be made WIDER to reduce skin loss.
Perspective drawing showing the air bridged center conductor contacting the top of a diode. The Schottky contact is kept well away from the edge of the $H^+$ implanted region, which ends outside the ohmics.
Elevated CPW III

Air Bridged Center Conductor

CPW Ground

Ohmic

N⁻ Active Layer

N⁺ Layer

Ohmic

CPW Ground

Semi-Insulating GaAs Substrate

Cross section of the air bridge contacted diode. A layer of polyimide is used to keep the post off the substrate during electroplating. The ohmic contacts are recessed through the $N^-$ active layer to a heavily doped $N^+$ buried layer.
Elevated CPW IV

S.E.M. image showing the air bridged center conductor of the coplanar waveguide contacting the tops of the diodes on the NLTL without touching the substrate.
S.E.M. image of the sampling circuit and the output end of the NLTL that provides the strobe pulse to the two $1 \mu m \times 1 \mu m$ sampling diodes.
NLTL output measured by sampling circuit

3.7 Volt step with 0.68 ps measured 10%-90% falltime; 725 GHz deconvolved sampler bandwidth
GaAs Picosecond Optical Waveform Analyzer

Measured Response; Sampling Circuit, 5 μm x 5 μm Detector
Instruments and Measurements
Instrumentation: Motivation

The Problem: Device vs Instrument Bandwidth

Microwave transistor measurements: almost 10:1 frequency extrapolation.

Modern device bandwidth exceeds instrument bandwidth.

High-frequency device models must be:

Poor understanding of devices.

Circuits above 60 GHz hard to measure.
Types of Instruments/ Measurements

Conventional Electronic Measurements

- Network Analyzer
- Time Domain Reflectometer
- Sampling Oscilloscope
- Spectrum Analyzer

Optoelectronic Techniques

- Photoconductive probing
- Electrooptic Sampling: Substrate Probing and Needle Probe

On-wafer network analysis

- GaAs NLTL-based active probes
- Photoconductor-based probes
Network Analysis

• Measures linear 2-port stimulus-response characteristics of a device

• Data usually presented as admittance or wave scattering parameters as a function of frequency.

• 2 Purposes:

Functional measurements of a component (gain-frequency curve, etc.)

Device characterization and modelling
Small-Signal Network Measurements
Restrict to linear devices (or nonlinear device in small-signal regime)

General device model:

Frequency-domain description:

\[ v_1(t) = V_1(\omega)e^{j\omega t}, \]
\[ i_1(t) = I_1(\omega)e^{j\omega t}, \text{ etc.} \]

Two-port Admittance parameters

\[
\begin{bmatrix}
I_1(\omega) \\
I_2(\omega)
\end{bmatrix} =
\begin{bmatrix}
Y_{11}(\omega) & Y_{12}(\omega) \\
Y_{21}(\omega) & Y_{22}(\omega)
\end{bmatrix}
\begin{bmatrix}
V_1(\omega) \\
V_2(\omega)
\end{bmatrix}
\]
Microwave Scattering Parameters I.

Waves on transmission lines:

forward and reverse waves

positive Z direction

a,b: forward and reverse waves

Forward, reverse power: |a|^2, |b|^2

Voltage

Current

Voltage at any point:

\[ v(z,t) = a(z,t)\sqrt{Z_0} + b(z,t)\sqrt{Z_0} \]

Current at any point:

\[ i(z,t) = a(z,t)/\sqrt{Z_0} - b(z,t)/\sqrt{Z_0} \]
Microwave Scattering Parameters II

Two-port parameters described in terms of incident and emanating waves from the device (when connected to transmission lines)

Equivalent S-parameter and Y-parameter models of a 2-port.

Scattering parameter model

$$
\begin{bmatrix}
  b_1(\omega) \\
  b_2(\omega)
\end{bmatrix} =
\begin{bmatrix}
  S_{11}(\omega) & S_{12}(\omega) \\
  S_{21}(\omega) & S_{22}(\omega)
\end{bmatrix}
\begin{bmatrix}
  a_1(\omega) \\
  a_2(\omega)
\end{bmatrix}
$$

Admittance parameter model

$$
\begin{bmatrix}
  I_1(\omega) \\
  I_2(\omega)
\end{bmatrix} =
\begin{bmatrix}
  Y_{11}(\omega) & Y_{12}(\omega) \\
  Y_{21}(\omega) & Y_{22}(\omega)
\end{bmatrix}
\begin{bmatrix}
  V_1(\omega) \\
  V_2(\omega)
\end{bmatrix}
$$

Since $v = a\sqrt{Z_0} + b\sqrt{Z_0}$ and $i = a\sqrt{Z_0} - b\sqrt{Z_0}$, the scattering (S) parameters can be directly computed from the admittance (Y) parameters.
Subsequent notes will discuss laser-based testing of electronics. Key points in transistor measurements:

Are we measuring device or circuit performance? Device measurements are parametric (what $f_t$, what $f_{max}$, what $C_{gs}$...?), while circuit measurements are functional (gain-bandwidth obtained, logic gain propagation delay, pulse amplifier risetime, ...)

Device measurements should allow the device engineer to extract the linear model, from which figures-of-merit ($f_t$, $f_{max}$) and device physical parameters (capacitances, resistances, transit times) are determined. The circuit engineer will want a circuit model of the device, described either as above (capacitances, resistances, transit times) or by a black-box linear 2-port description.

Pulse "response time" measurements with unspecified bias conditions, signal levels, and generator and load impedances are of little significant value. Linear or large-signal nonlinear operation will give quite different behavior. Bandwidth (risetime) is a function of gain because of gain-bandwidth limits. Very short RC charging times may be obtained by driving the transistor through very low impedances, but the circuit may be providing no power gain under such conditions...
The Microwave Network Analyzer
Measurement of (linear / small signal) 2-port network parameters in the frequency domain.

Swept-frequency sources ($V_{\text{gen}1}$ and $V_{\text{gen}2}$) are alternately applied to the 2-port input and output, and the incident and emanating waves measured with directional couplers.

Calibration: amplitude/phase contributions of cabling (etc.) between the instrument and the d.u.t. are corrected for by first measuring a series of devices of known characteristics in place of the d.u.t., either 50Ω load, open, short, and through line, or a series of through lines of differing lengths ("LRL")
Block Diagram: Microwave Network Analyzer

Channel #1

- Stimulus signal
- Directional coupler
- Microwave forward & reverse waves
- Sampling circuits
- 20 MHz forward & reverse waves
- 20 MHz amplitude and phase measurement

Channel #2

- Device Under Test
- **V**^+\_1 → **V**^-\_2
- **V**^-\_1 → **V**^+\_2

Channel #2...

**V**^-\_1 \& **V**^+\_2

Bandwidth limits include

- Available connectors and cables,
- Sampling circuits,
- Signal source frequency range

\[
\begin{bmatrix}
V^-_1 \\
V^-_2
\end{bmatrix} = \begin{bmatrix}
S_{11} & S_{12} \\
S_{21} & S_{22}
\end{bmatrix} \cdot \begin{bmatrix}
V^+_1 \\
V^+_2
\end{bmatrix}
\]
Performance of modern network analyzers:

After Calibration: DC-110 GHz instrument
(Coaxial-based system, using coplanar microwave wafer probes)

Amplitude accuracy, 0 dB signal: $\approx \pm 0.05$ dB

Phase accuracy, 0 dB signal: $\approx \pm 3^\circ$

Directivity*: $\approx -40$ dB
*Measured reflection magnitude for a zero-reflection device

Given accurate calibration standards, network analyzers can provide very precise device models. Competing optical techniques offering wider bandwidth must attain competitive accuracy. This places stringent demands on laser intensity stability (and often laser pulse timing stability).
Impedance measurement of a 4 THz Schottky diode. Device Q is 100 at 40 GHz, hence S11 differs from 0 dB by ≈0.05 dB. Instrument accuracy is sufficient to observe the diode resistance!
**Time-domain reflectometry:**
Measurement of the 2-port network small-signal parameters in the time domain: Yields same information as the swept-frequency analyzer (?).

First method uses directional couplers to separate incident & reflected waves

\[ s_{ij}(t) \] is simply the inverse Fourier transform of \( S_{ij}(\omega) \):

Using a delay line with delay longer than the duration of the stimulus signal, the incident & emanating waves are separated in time, eliminating (?) the need for directional couplers.
Problems with Time-Domain NWA Measurements, I

While time-gating eliminates spurious responses, frequency resolution is lost.

Frequency Resolution is $\Delta f = 1/\Delta t$
Problems with Time-Domain NWA, II
Spectrum Downconversion by Sampling

- Aliasing of phase-noise sidebands degrades noise performance if scan rate is below phase-noise bandwidth
- System imperfections (phase, amplitude noise) and DUT nonlinearity cause mutual interference between spectral lines under measurement
- Other problems: sharing bits of resolution in A-D converter between harmonics
- Conclusion: Commercial NWAs use swept-frequency stimulus for good reason
Signal Measurements

Samping oscilloscope
Time waveform of signal. Commercial instruments to 50 GHz (NLTL based...).

Spectrum Analyzer
Measures power spectrum of signal. Commercial instruments: 40 GHz with coaxial inputs, to 325 GHz in waveguide.
Optoelectronic Measurement Techniques

Sampling Devices and Systems

Electrooptic Sampling

Photoconductive Sampling
Sampling Devices, Optoelectronic and Electronic

**Electrical**

Input

\[ +V_{\text{bias}} \]

- Strobe

\[ -V_{\text{bias}} \]

Grove, Hewlett-Packard, 1966

\[ +V_{\text{bias}} \]

Input

\[ -V_{\text{bias}} \]

Auston, Bell Labs

\[ +V_{\text{bias}} \]

Output

\[ -V_{\text{bias}} \]

**Optical**

Input

Picosecond Laser Strobe Pulse

Output

Photoconductive Gap

Grove, Hewlett-Packard, 1966

Auston, Bell Labs

Valdmanis and Mourou, University of Rochester

Transmission line on Electrooptic substrate

Photodetector
Sampling
Reducing the repetition frequency (bandwidth) of a signal so that it can be measured with low-frequency instruments.

If the strobe signal has repetition frequency $f_0$ and the input signal has repetition frequency $nf_0 + \Delta f$, the sampled output will be at frequency $\Delta f$. 

Input

Strobe

Output

If the strobe signal has repetition frequency $f_0$ and the input signal has repetition frequency $nf_0 + \Delta f$, the sampled output will be at frequency $\Delta f$. 

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Laser Sampling and Timing Jitter

Mode-locked lasers derive their pulse repetition rate from the cavity round-trip time. This resonator, in terms of the laser intensity modulation intensity envelope, has relatively poor Q (poor finesse) and pulsed lasers have substantial pulse timing fluctuations.

Relative timing fluctuations of the laser and electrical signal source degrade the system time resolution.

Good microwave synthesizer: ≈0.2 ps rms jitter

Mode-locked YAG laser: ≈3-10 ps rms (0.3 ps if phase-locked)
CPM laser: ≈5 ps rms jitter
The pump-probe technique

Because the stimulus and probing signals are derived from the same laser pulse, laser pulse timing fluctuations have no effect on the measurement.

technique limited to stimulus-response measurements
Photoconductive Characterization of Devices

**Purpose:** 2-port small-signal network measurements. **Method:** time-domain reflectometry/transmission using photoconductive pulse generators and sampling gaps: Matloubian et. al., IEEE MGWL Vol. 1. No. 2, Feb. 1991
Electrooptic Probing

Field-Induced birefringence in [100]-cut GaAs:

\[ N_{[01\overline{1}]} - N_{[011]} = N_0^3 r_{41} E_{[110]} \]

A sub-bandgap probe beam is passed in the [100] direction through the substrate, and the birefringence measured with a polarization interferometer. The interferometer output is proportional to the potential difference across the wafer.

Principal axes and cleave planes in (100)-cut Gallium Arsenide.
Gallium Arsenide electrooptic intensity modulator
Electrooptic: Frontside/backside probing

Frontside probing

[100]
Probe Beam

Conductor

GaAs Substrate
≈100 μm thick

Ground Plane

Backside probing

[100]

Signal
Gnd

GaAs Substrate
≈400 μm thick

Probe Beam
Example Measurement: eo sampling

Voltage waveforms at the input (a), source-follower gate (b), and output (c) of a GaAs buffered-FET-logic inverter gate.

100 ps/div.
Electrooptic Sampling of Photodetectors

Mode-Locked Nd:YAG Laser 1.06 μm

Fiber-Grating Pulse Compressor

Autocorrelator

AO Modulator

Pump Beam

Prober Beam

Delay Stage

Photoreceiver tuned to AO modulation frequency

Device Under Test

Microwave Probe

Lens λ/2 λ/4 PBS

Sampling Oscilloscope

85ps FWHM

2ps FWHM

1.06 μm

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Electrooptic Sampling of Photodetectors: Result

Response of InGaAs PIN Photodiode

- Measured response
- Deconvolved response

Time (ps) vs. Voltage (Arbitrary Unit)

- 5 ps
- 3.8 ps
Electrooptic Needle Probe

- J.A. Valdmanis

Diagram showing components of an electrooptic needle probe:
- Longitudinal probe beam position
- Transverse probe beam position
- Integrated circuit
- Electro-optic crystal
- Lased silica

Graph showing relative strength over time (ps):
- Time range: 0 to 4 ps
- Relative strength range: 0 to 1

Additional diagram showing a laser, delay, trigger beam, bias line, probe tip, and eyepiece connected to detectors, analyzer, and compensator.
Laser sources for optoelectronic measurements

Colliding-pulse mode-locked dye laser:
Pulse durations <<100 fs.
Free-running laser (passively mode-locked): 5-10 ps rms timing jitter
wavelengths above-bandgap in GaAs: not for electrooptic substrate
probing. Relatively stable amplitude. Primary laser for femtosecond
optoelectronic experiments.

Synchronously-Pumped dye lasers
0.1-5 ps pulse durations. Primarily for wavelengths less than 900 nm (no
substrate probing in GaAs).
Significant (1-3 ps rms) timing jitter. Stable amplitude

Ti-Sapphire Laser
≈200 fs pulse widths, negligible wings, tunable wavelengths
Sibbet et. al. have timing-stabilized such lasers. Good choice.
Laser sources for optoelectronic measurements II
Nd:YAG laser & fiber-grating pulse compressor
Wavelength (1.06 μm) below-GaAs-bandgap; suitable for substrate probing.
Subpicosecond timing jitter attainable with feedback stabilization.
Poor spectral flatness (measurement error) due to wings on pulse.
Considerable low-frequency amplitude fluctuations, drift.
Power-dependent Raman scattering adds excess noise above shot noise, limiting input power to fiber, and thereby limiting usable pulse compression to ≈1 ps outputs with single-stage compression.

Mode-locked & gain-switched semiconductor diode lasers
Area of intense research. Low phase noise for actively mode-locked devices, particularly those with external optical cavities. Pulse widths ranging from ≈0.1 ps to 2-5 ps, dependent on pulse repetition frequency. Adiabatic soliton compression looks like a breakthrough
Compressed YAG Autocorrelation: "wings"

Nd:YAG laser with fiber-grating pulse compressor

≈1.5 ps FWHM pulse duration
"Wings" cause errors in low-frequency measurement response. Intensity fluctuations from compressor degrade repeatability. Raman scattering adds (≈10-50 dB) to shot noise, limiting compression ratios attainable with low noise.
phase noise theory and measurement

Laser timing and amplitude fluctuations:

\[ I(t) \equiv P\bar{T}(1 + N(t)) \sum_{-\infty}^{+\infty} \frac{1}{\sigma_t \sqrt{2\pi}} \exp\left[ -\frac{(t - nT - J(t))^2}{2\sigma_t^2} \right] \]

Laser Spectrum

\[ S_I(\omega) \approx \bar{P}^2 \exp(-\omega^2 \sigma_t^2) \sum_{-\infty}^{+\infty} \left[ 2\pi \delta(\omega - \omega_0) \quad \text{...laser harmonics} \\
+ S_N(\omega - \omega_0) \quad \text{...AM sidebands} \\
+ n^2\omega_t^2 S_J(\omega - \omega_0) \quad \text{...FM sidebands} \right] \]
phase noise theory and measurement

I(t) laser intensity
P=average intensity
T=pulse repetition period
\( \sigma_t \) =RMS pulse width
N(t)=intensity fluctuations
J(t) timing fluctuations
phase noise integration

\[ J^2(t) = \sigma^2 = \frac{1}{\pi} \int_{\omega_{\text{low}}}^{\omega_{\text{low}} + \omega_i/2} S_f(\omega) d\omega \]

\( \omega_{\text{low}} \) is the low-frequency limit of integration

\( \approx \pi / T \), where \( T \) is the period of observation

In words: Power in sidebands divided by power in carrier is mean-squared phase deviation in radians. Divide by the radian frequency to obtain the timing deviation
Laser Timing Stabilization

This method is applicable to any actively mode-locked laser. Passively mode-locked lasers can be phase-locked by introducing an electrically-controlled cavity tuning element.
On-Wafer Network Analysis

- photoconductive sampling probes
  - external probes for S-parameter measurements
  - internal-node probes
- NLTL/diode sampling IC probes for waveform & network measurements
- STM/AFM probes
Active Probes

Highly Reproducible Optoelectronic Wafer Probes with Fiber Input
M. D. Feucht, S. C. Stumpf, P. R. Smith, H. H. Lau, C. A. Barros, and M. C. Mass
AT&T Bell Labs, 101 Crawfords Corner Rd., Holmdel, NJ 07733-3030
1. AT&T Bell Labs, 600 Mountain Ave., Murray Hill, NJ 07974-0636
2. AT&T Bell Labs, 791 Holmdel-Kennett Rd., Holmdel, NJ 07733-0400

Figure 1. Schematic layout of optoelectronic wafer probe with ground bridging strips. Optical pulses are introduced through the back of the probe substrate. The system provides high bandwidth, throughput and accuracy.

Figure 2. S11 of a thin-film resistor with a dc resistance of 12.5 Ohm, from dc to 125 GHz at 2.5 GHz intervals, after full correction with vector accuracy enhancement.

Subpicosecond GaAs Wafer Probe System
M. S. Shakouri, A. Black +, D. M. Bloom

Edward L. Ginzton Laboratory, Stanford University
Stanford, CA 94305
+ New at Gageon Microsystems, Los Gatos, CA

Low Frequency
Ribbon Connector
Parallelplate Flexures

50 GHz Coax Connectors

Coax to CPW Transition Circuit
GaAs IC with Micromachined Probe Tip

FIG. 1. GaAs Wafer Probe Assembly

10-90% Fall Time = 880 fsec
5 ps/2 div
On-Wafer Signal Measurements

Novel High-Impedance Photoconductive Sampling Probe for Ultra-High Speed Circuit Characterization

Joungba Kim, Yi-Jen Chan*, Steven Williamson, John Ness, Shin-ichi Wakana, John Whitaker, Dimitris Pavlidis*

Center for Ultrafast Optical Science, *Center for High-Frequency Microelectronics
The University of Michigan, EECS
2260 Bonisteel, 1006 SIT
Ann Arbor, MI 48109-2109

Fig. 1 Experimental set-up to measure short electrical pulses propagating along a coplanar strip. We used 120-fs Ti-Sapphire laser for pump and probe beam.
NLTL-based Probe for On-Wafer Network Analysis

- **signal routing substrate**
- **NWA integrated circuit**
- **quartz CPW probe tip**

<table>
<thead>
<tr>
<th><strong>NLTL-based IC</strong></th>
<th>connected to DUT through wideband CPW-on-quartz probe tip</th>
</tr>
</thead>
</table>

**Stimulus Signal Generator**

**Attenuator**

**Directional Sampler**

**Device Under Test**

**Forward and Reverse Waves (Sampled)**

**Strobe Signal Generator**

**Strobe**

**6-dB Attenuator**

Measurements made in frequency domain with pulsed stimulus signal. Forward & reverse waves separated by directional sampler, not by time-gating.
System Block Diagram

- **controller (workstation)**
- **digitizing oscilloscope**
- **IF signal processing electronics**

**device under test**

**7-200 GHz signals**

**downconverted forward & reverse waves (10-500 kHz)**

**7-14 GHz NLTL drive signals**

**stimulus**

**strobe**

**active probe**
NLTL-based active probe pulse response

DUT: 0.7 ps NLTL pulse generator

3.1 ps probe risetime
110 GHz probe bandwidth

active probe

measured signal (V)

Time (ps)
Measurements with active probe
Measurements with active probe

| Frequency (GHz) | |S11| (dB) |
|---------------|---|-------|
| 0             | -80 |
| 40            | -60 |
| 80            | -40 |
| 120           | -20 |
| 160           | 0   |
| 200           | 40  |

1st measurement

2nd measurement
Measurements with active probe

Measured Results: First Design Iteration

Forward Gain, $S_{21}$, dB

Frequency, GHz

10 dB gain, 92 GHz bandwidth

8 dB gain, 98 GHz bandwidth

(capacitive-division TWA shown earlier.)
AFM active probe

Figure 1. High-speed SFM probe system.

- Ho et al, 1995 Ultrafast Electronics & Optoelectroncis conference
- Force on AFM tip proportional to voltage squared--provides nonlinearity to sample DUT signal with pulse train applied through AFM tip.
- Bettina et al, 1997 UFEO conference: NLTL pulse generator, ~ 1 ps resolution
**pico-/femto-second laser spectroscopy**
(mm-Wave and sub-mm-wave Gain-Frequency Measurement System)

- Broadband: ≈1.5 THz demonstrated
- Need picosecond laser (big, $$)

Purpose: measurements of materials and sub-mm-wave components
(sub)picosecond Laser Spectroscopy: Exter et al.

Fig 3-2
(a) Measured electrical pulse of the freely propagating terahertz beam in pure nitrogen. The inset shows pulse on an expanded time scale. (b) Measured electrical pulse with 1.3 Torr of water vapor in the enclosure. The inset shows pulse on a 20X expanded vertical scale.

Fig 3-3
(a) Amplitude spectra of Figs. 2(b) and 2(c). (b) Amplitude absorption coefficient obtained from Fig. 3(a). (c) Relative phase of the spectral components of Fig. 3(a).
NLTL-based Transmitter and Receiver

Transmitter

Receiver

sampled output

NLTL sampled output

sampling circuit

100 Ω termination

60°

100 Ω termination

NLTL

bow-tie antenna

bow-tie antenna
Free-Space mm-Wave Measurement System

transmitter IC

phase reference
mixer

trigger

oscilloscope

IF signal

receiver IC

synthesizer (RF)

synthesizer (LO)

hyper-hemispherical lens

attenuators

off-axis paraboloidal mirror

material or array under test
Measured Signal with NLTL-based system

- Received Waveform (mV)
  - 270 mV
  - 2.4ps risetime, 10-90%
measurement with NLTL-based system

9.5-Period Alumina-Teflon Bragg Filter

Insertion Loss (dB)

Frequency (GHz)

measured
theory
measurement with NLTL-based system
Appendix sections:

Soliton NLTL impulse compressors
Traveling-wave RTD pulse generators
RTD Array Oscillators for THz signal generation

- Power per device is small.
- Combine outputs from array of \( \approx 1000-10,000 \) SRTD oscillators.
- Cavity resonator defines oscillation frequency, provides strong coupling between array elements.

<table>
<thead>
<tr>
<th>Array oscillator layout</th>
<th>Quasi-optical Array Oscillator:</th>
</tr>
</thead>
<tbody>
<tr>
<td><img src="image1" alt="Array oscillator layout" /></td>
<td><img src="image2" alt="Quasi-optical Array Oscillator" /></td>
</tr>
</tbody>
</table>

- Metal heatsink and reflector
- RTD array
- Output aperture
- Semi-confocal resonator
A Critical Issue: Suppression of Bias Circuit Oscillations

Oscillator: array in resonator

Array unit cell: RTD with bias stabilization

- Without stabilization, bias circuit stability issues limit power to \( \approx 0.5 \text{ mW} \).

- integrated Schottky stabilizer diodes
Maximum RTD Output Power

- Cubic Polynomial fit to RTD I-V characteristics
- Peak negative conductance: \( G_n = \frac{3\Delta I}{2\Delta V} \)
- Maximum RTD output power:
  \[ P_{\text{max}} = \frac{3}{16} \Delta I \Delta V (1 - \frac{f^2}{f_{\text{max}}^2}) \]
- In terms of \( G_n \):
  \[ P_{\text{max}} = \frac{\Delta V^2 G_n}{8} (1 - \frac{f^2}{f_{\text{max}}^2}) \]
Conditions for Stability

- Find complex frequencies \( s_i = \sigma_i + j\omega_i \) at which \( Y_{ckt}(s) = G_n \)
- Necessary and sufficient condition for stability is \( \sigma_i < 0 \)
- A sufficient condition for stability is
  \[
  \text{Re}[Y_{ckt}(j\omega)] = G_{ckt}(j\omega) > G_n
  \]
- At all frequencies where stability is demanded the external circuit should present a low impedance
Power Limitations Imposed by Stability Requirements

A sufficient condition for stability is
\[ \text{Re}[Y_{ckt}(j\omega)] = G_{ckt}(j\omega) > G_n \]

DC stability demands that \( \frac{1}{R_{bias}} > G_n \):
\[ P_{\text{max}} = \frac{\Delta V^2}{8R_{bias}} \]

Stability at other frequencies constraints \( R_{bias} = Z_0 \):
\[ P_{\text{max}} = \frac{\Delta V^2}{8Z_0} \]

Power limitation eliminated if \( L < \lambda / 4 \): On wafer bias stabilizer at submm-wave frequencies
RTD Bias Stabilization using shunt Schottky diodes

- 2 SRTDs share a stabilizer with $R_{stab} = 25\Omega$
- Easily extended to larger SRTDs by area scaling
Slot Antenna Coupled RTD Oscillator

- Slot Antenna: Resonating and Radiating element
- SRTD capacitance detunes the slot length
- Easily extended to arrays
InGaAs/AlAs SRTDs with InGaAs/InAlAs Schottky stabilizer diodes

- Schottky and Interconnect metal
- Buried N++ contact layer
- Undoped layers
- Ohmic Metal
- Schottky Diode
- Posts and Airbridge metal
- SiN Capacitor
- Semi-insulating InP
- InGaAs/InAlAs SRTD
Oscillator Array
Quasi-optical Setup for Testing Submm-wave Oscillators
Bolometer output Vs Mirror displacement for a 64-element array at 800 GHz

Output power (A.U.)

Fabry-Perot Mirror Displacement (mm)

λ/2 0.23 mm

650 GHz oscillation

Bolometer output Vs Mirror displacement for a 64-element array at 800 GHz
Soliton NLTLs: Impulse Compression

Shock-wave devices:
competition of nonlinearity against dissipation
wavefronts compressed into shock-waves
picosecond step-functions are formed

Soliton-propagation devices:
competition of nonlinearity against dispersion
input waveforms decompose into sets of solitons
appropriate scaling: compression of impulses
picosecond large-amplitude impulses are formed.
Signal Distortion on NLTLs

Diode Nonlinear Reactance

Periodic-Network (Bragg) Dispersion

Diode Capacitance

Group Delay

Voltage

Frequency

$f_{\text{Bragg}}$
Solitary Waves on NLTLs

• Pulse waveforms for which nonlinearity and dispersion are in opposition

• Propagate without dispersion

• Larger amplitude solitary waves propagate faster

• Larger amplitude solitary waves have shorter duration

• Solitary wave duration inversely proportional to Bragg frequency.
Soliton Collision
Solitons: solitary waves which are undistorted after collisions.
Compression through Soliton Decomposition

Input Signal

Output Signal

longer duration, lower amplitude

shorter duration, larger amplitude
Line has $\approx 35 \text{ GHz}$ Bragg frequency.

Pulse widths $\approx 1/4f_{\text{Bragg}}$ will split into 2 solitons.

Longer pulses will split into 3 or more solitons per cycle.

Compression ratio limited to approximately 2.5:1
Compression by Repeated Soliton Splitting
Compression on a 2-Step Line

Input:
31.5 GHz, 27 dBm

Output:
8.1 V, 87 mA, 0.70 W peak, 4.5 ps FWHM
Continuous Compression on a Tapered Line
Result: Tapered Impulse Compressor

Output: 11.4 V, 126 mA, 5.15 ps FWHM
Soliton Impulse Compressors: Summary

• Current performance
  \( \approx 2:1 \) amplitude gain but limited bandwidth (5 ps measured, 2 ps theoretical).

• Fundamental limits:
  Soliton (as opposed to shock) formation requires \( f_{\text{Bragg}} < f_{\text{diode}} \).
  2 THz (14 V Breakdown) diodes then should permit 2 ps FWHM pulses.
  Faster diodes have lower breakdown voltages.
  (Epitaxial or wired) series diodes may permit high \( f_{\text{diode}} \) and high \( V_{\text{BR}} \).

Compared to Shock-Wave NLTLs:
  Poorer compression/length (bigger die, expensive)
  \( \approx 2:1 \) poorer (theoretical) pulse performance (above)
  much poorer experimental pulse performance
  \( \approx 3 \) times larger output voltage (9 times in power)
Traveling-Wave RTD Pulse Generator

TWRTD is a distributed structure

Capacitance charging times are eliminated

Risetimes limited by \( f_{\text{max}} \)

if \( R_s \ll R_n \), TWRTD is several times faster than lumped RTD

NOT useful for logic: fast risetime, long delay.

\[
T_{10\%-90\%} \equiv (\ln 0.9 - \ln 0.1)2C\sqrt{R_n R_s} = 0.70 / f_{\text{max}}
\]
Lumped- vs Traveling-Wave RTD Risetimes

For RTD with low $R_S$, TWRTD is several times faster.
TWRTD Measured Output Waveform

Traveling-Wave RTD Pulse Generator

Output Amplitude, volt

( measured with an .NLTL-based active probe )
References
CLEO Short Course, Ultrafast Electronics and Optoelectronics, 1998

Photodetectors


D.G. Parker, P.G. Say, A.M. Hanson and W. Sibbett, "110 GHz High-Efficiency Photodiodes Fabricated from Indium Tin Oxide/GaAs", Electronics Letters, 22 November 1989, pp s66-s67


Photoconductors

*Photoconductive Devices*


*TeraHertz Spectroscopy using photoconductors*


On-Wafer Network Analysis Using Photoconductors


Step-Recovery Diodes:


Resonant Tunnel Diodes


Terahertz Technology (Radio Astronomy and Imaging)

See the IEEE Proceedings, special issue on Terahertz technology, November 1992. Review Papers on Schottky diodes, Transmission lines, antennas, multipliers, quasi-optical arrays, and Terahertz spectroscopy

100 GHz-5 THz Schottky Diodes


Transmission Lines and Antennas


Reinmut Hoffmann, Handbook of Microwave Integrated Circuits, Artech House, Norwood, MASS, 1987. This is a good general reference for transmission line design formulas.

Transistors

U. K. Mishra, A.S. Brown, and S.E. Rosenbaum: "DC and RF performance of 0.1 µm Gate Length AllInAs-GaInAs Pseudomorphic HEMTs" In Technical Digest, 1988 International Electron Device Meeting, Dec. 11-14, San Francisco.


S. Yamahata, K. Kurishima, H, Nakajima, T. Kobayashi and Y. Matsuoka,


**Electrooptic Sampling**


**Diode Sampling Circuits (non-NLTL-based)**


**Nonlinear Transmission Lines**

*Basic Principles: Electrical Shock-Waves and Solitons*


A. Scott, Active and Nonlinear Wave Propagation in Electronics, Wiley-Interscience, New York, 1970. This provides a wide overview of the literature with many references.


**Shock-Wave NLTLs and Sampling Circuits**


**Soliton Devices**


**Applications: Network Analysis, Spectroscopy, Photodetector Integration**
Ruai Yu, Madhukar Reddy, Joe Pusl, Scott Allen, Michael Case, and Mark Rodwell, "Full Two-Port On-Wafer Vector Network Analysis to 120 GHz Using Active Probes", 1993 IEEE Conference on Microwave Theory and Techniques, June, Atlanta, Ga


**Laser Timing Stabilization:**