Ultra-fast quantum-well infrared photodetectors operating at 10µm with flat response up to 70GHz at room temperature

M. Hakl,† Q.Y. Lin,† S. Lepillet,‡ J.-F. Lampin,§ S. Pirota,‖ R. Colombelli,∥ W. J. Wan,∥ J. C. Cao,∥ H. Li,∥ E. Peytavit,∥ and S. Barbieri†,*

†Institute of Electronics, Microelectronics and Nanotechnology, Univ. Lille, ISEN, CNRS, UMR 8520, 59652 Villeneuve d’Ascq, France
‡Centre de Nanosciences et de Nanotechnologies (C2N), CNRS UMR 9001, Université Paris-Saclay, 91120 Palaiseau, France
§Key Laboratory of Terahertz Solid State Technology, Chinese Academy of Sciences, Shanghai, 200050, China
∥Corresponding author: stefano.barbieri@univ-lille.fr

These authors participated equally to this work.

Abstract

III-V semiconductor mid-infrared photodetectors based on intersubband transitions hold a great potential for ultra-high-speed operation up to several hundreds of GHz. In this work we exploit a ~350nm-thick GaAs/Al0.7Ga0.3As multi-quantum-well heterostructure to demonstrate heterodyne detection at λ~10µm with a nearly flat frequency response up to 70GHz at room temperature, solely limited by the measurement system bandwidth. This is the broadest RF-bandwidth reported to date for a quantum-well mid-infrared photodetector. Responsivities of 0.15A/W and 1.5A/W are obtained at 300K and 77K respectively. To allow ultrafast operation and illumination at normal incidence, the detector consists of a 50Ω coplanar waveguide, monolithically integrated with a 2D-array of sub-wavelength antennas, electrically interconnected by suspended wires. With this device architecture we obtain a parasitic capacitance of ~30fF, corresponding to the static capacitance of the antennas, yielding a RC-limited 3dB cutoff frequency >150GHz at 300K, extracted with a small-signal equivalent circuit model. Using this model, we quantitatively reproduce the detector frequency response and find intrinsic roll-off time constants as low as 1ps at room temperature.

1. INTRODUCTION

Thanks to their intrinsically short electron relaxation time, on the ps timescale, mid-infrared (MIR-3-12µm) quantum-well infrared photodetectors (QWIP) based on III-V semiconductor materials were identified as ideal candidates for ultra-high-speed operation at the end of the 80s. Since then, several experiments have been carried out to determine their RF bandwidth using both pulsed mid-infrared excitation or heterodyne detection [1-7].

The exploitation of QWIPS as heterodyne receivers with IF bandwidth of tens of GHz is particularly attractive for a number of applications, including free-space communications, gas sensing and spectroscopy, atmospheric and space science, or military countermeasures [8-11]. Besides enabling the implementation of coherent detection schemes, another advantage brought by heterodyne detection is the possibility to operate QWIPS in the shot-noise regime, overcoming the noise contribution of the thermally activated dark current, which severely impacts the NEP of MIR QWIPS at high temperatures [12]. So far, the highest heterodyne detection frequencies are those reported in 2005 by Grant et al. with a 100-quantum wells (QWs) QWIP operating at a 10µm [6,7]. In this work, the detector was processed in a 16µm-side square mesa illuminated from a 45° polished substrate. Heterodyne detection up to 75GHz and 110GHz was demonstrated respectively at 80K and 300K. Due to the limited tunability of the CO2 lasers used in the experiment, only scattered sets of heterodyne beatnotes could be detected above 30GHz, showing power drops of ~20dB and ~10dB from 35GHz to 75GHz at T=80K and 300K.

In the context of high-speed QWIPs the possibility of coupling the detector element to an antenna opens interesting perspectives. Indeed, the antenna allows a reduction of the detector active volume without sacrificing the radiation collection area, thus avoiding a reduction of the quantum efficiency. Nano-antennas were first applied to MIR bolometers as a way to increase both their sensitivity and speed [13-15]. QWIPS based on arrayed patch antennas resonators (PARs) were first proposed in 2001 [16], and recently demonstrated [17-18]. PARs are ideally suited for QWIPS as they allow illumination at normal incidence, which is clearly advantageous compared to facet illumination [6,7], while confining the electro-magnetic field inside a sub-wavelength volume [16,18,19]. Compared to QWIPS based on standard mesa geometry and of comparable collection area, this enables the realization of “thin” detectors (including a small number of QWs) with a higher detectivity while keeping a small capacitance, which is clearly relevant for high speed operation.

In this work we have fully exploited this possibility by demonstrating PARs-based QWIP detectors specifically designed for ultra-broadband operation. Thanks to this design we demonstrate experimentally that the detector capacitance is dominated by the (unavoidable) static capacitance of the antenna resonators. With these devices we demonstrate at room temperature, heterodyne detection at 10.3µm with a nearly flat frequency response up to 70GHz (limited by...
the detection electronics), and state of the art responsibilities of ~0.15A/W. To the best of our knowledge this represents the broadest experimental RF bandwidth reported to date for a QWIP detector [6,7,21]. Moreover, we develop a small-signal equivalent circuit model that can quantitively reproduce the observed device frequency response, which we find to be strongly dependent on bias and temperature. From this model we extract an RC limited 3dB cutoff of ~150GHz at 300K, and an intrinsic roll-off time constant ~1ps, providing the first experimental evidence that QWIP detectors can indeed reach RF-bandwidths limited by electron capture on the ps timescale at room temperature. These results pave the way to the development of ultrafast MIR optoelectronics.

2. DEVICE DESIGN AND FABRICATION

The structure is grown by molecular beam epitaxy (MBE) on a semi-insulating GaAs substrate: 100nm-thick lattice-matched Ga0.65In0.35P etch-stop layer followed by an Al0.2Ga0.8As/GaAs heterostructure. The PAR active region consists of seven, 6.5nm GaAs quantum wells (QWs) with a central, 5.3nm-thick region, n-doped at a level of 6.7x10^{17} cm^{-3}. The wells width is chosen to obtain a bound-to-bound transition energy of ~120meV. The QWs are separated by 40nm barriers, and active region is sandwiched between 50nm and 100nm-thick top and bottom n-doped contact layers with concentrations of 3x10^{18} cm^{-3} and 4x10^{18} cm^{-3} respectively.

In Fig. 1(a)(b), we present the SEM images of the fabricated detector. It consists of a 5x5 periodic array of square PARs of side s=1.85μm and period p=3.9μm, sitting on top of a Ti/Au (100/400nm) ground plane. As detailed in the next Section, the values of p and s are chosen to obtain a maximum PAR array absorption as close as possible to the intersubband transition energy. At the same time, to minimize the array capacitance, the number of patches is kept to the minimum needed to allow collecting 100% of the incident radiation (~20μm-diameter laser spot size, see next Section).

Particular care was taken in the detector microwave design, aimed at reducing the effect of parasitic capacitances brought by electrical connections and contact pads, which limited RF operation up to a few GHz in Ref. [18]. As shown in Fig.1(a)(b), this is achieved by connecting the 2D array to the central electrode of a 50Ωtapered coplanar waveguide through an air bridge. Besides minimizing the parasitic capacitance, this solution is ideal for on-wafer testing by means of a 67GHz microwave coplanar probe. Finally, individual resonators are connected by suspended gold wires (Fig.1(b)(c)): compared to keeping the semiconductor beneath [18], this solution allows minimizing the wires capacitance, while simultaneously eliminating the current flow outside the resonators, therefore reducing the dark current. As a result of our design, as shown in Section 4, we find that the detector capacitance is essentially coincident with the static capacitance of the PARs alone, of approximately 30fF.

The fabrication of the PARs begins with the realization of a buried metal layer serving as electromagnetic ground plane and bottom Schottky contact metallization. This is obtained by transferring the epi-layers onto a 2”-diameter high-resistivity (>5 kΩcm) silicon wafer using a Au-Au thermo-compression bonding technique detailed in [22], followed by the wet etching of the GaAs substrate and the etch-stop layer. Next, the Ti/Au (8nm/300nm) top Schottky contact metallization is realized through e-beam lithography, followed by e-beam evaporation and lift-off. The epi-layers are subsequently ICP etched using the top metal layer as etch-mask. The ground metal layer is finally dry-etched by an Ar+ ion-beam around the PARs array down to the silicon substrate. A 100-nm-thick Si3N4 layer is then deposited on the silicon by plasma enhanced chemical vapor deposition (Fig.1(a)(b)).

To electrically connect the patches together, suspended ~150-nm-width Ti/Au (20nm/600nm) wire-bridges are finally fabricated by a two-step e-beam lithography process. A first resist layer is used as support after deposition, e-beam lithography and reflow, followed by a second one to define the wires by standard lift-off process. The same process is used to realize the air-bridge connecting the 2D array to the 50Ω coplanar line. The latter is deposited on the Si3N4 this avoids any leakage currents between the electrodes of the coplanar waveguide coming from the silicon.

3. SPECTRAL and dc ELECTRICAL CHARACTERISATION

In Fig.2 we report the results of the infrared spectral characterization of the PARs array. Fig. 2(a) shows the absorption spectra at 300K, corresponding to the fraction of the incident power absorbed by the QWIP detector with two orthogonal polarizations of the incident light: orthogonal (black) and parallel (red) to the wire bridges [17,18]. The absorption is defined as 1-R(o), where R(o) is the reflectivity spectrum obtained through FTIR micro-reflectivity measurements. At the cavity resonance for the orthogonal polarization (116meV ~ 10.7μm) we find that, 1-R(o) = 0.9, i.e. 90% of the incident photons are absorbed. Indeed, the period p=3.9μm, is selected to operate the QWIP as close as possible to the critical coupling regime, compatibly with the targeted intersubband transition energy [20]. In this condition, the single PAR collection area at the resonant frequency is given by 0.9xπp^2, yielding a total collection area of ~340μm^2 (=18.5μm)^2 for the PAR array. As shown in the insets of Fig.2(a), for the parallel polarization, the spatial distribution of the cavity mode is modified by the presence of the wire bridges, This yields a blue shift of the cavity resonance, as well as a reduced integrated absorption.

In Fig.2(d) we report the measured photocurrent spectrum at 77K, obtained by FTIR spectroscopy with the QWIP structure processed in a mesa geometry, therefore showing the effect of the bare intersubband transition. We find a peak at 107meV, in good agreement with the expected bound-to-bound transition.
Energy. By multiplying this spectrum by the cavity absorptions in Fig.2(a) we obtain the dashed spectra shown in panel (b), in good agreement with the QWIP detector photocurrent spectra measured a 77K, represented by the solid lines. From the black solid line we find that the QWIP operates in the ~10-12μm range, with a maximum response at ~10.8μm (115meV).

In Fig.3(a),(b) we report the dark current and dc photocurrent vs bias characteristics at 77K and 300K, obtained by illuminating the QWIP with a 10.3μm (120meV) DFB quantum cascade laser (QCL), polarized at 45° with respect to the wire bridges. For these measurements, the collimated beam from the QCL was focused on the detector using an AR coated aspheric lens (NA = 0.56; 5mm focal length). At 10.3μm we measured a waist diameter of 20μm using a knife-edge technique, i.e. approximately equal to the side of the 5x5 PAR array collection area (18.5μm ~ 340μm²). Therefore, for the rest of this work, we assume that all the QCL power, measured after the lens, is incident on the QWIP. This corresponds to the power values reported in Fig.3(a),(b).

As expected, at 300K the dark current dominates the photocurrent for all power levels. At 77K the situation is reversed, showing that at this temperature the QWIP can be potentially operated in the photon-noise regime with only a few mW of incident power [2]. At 77K and 3.5-4V (Fig.3(a)) we also observe a pronounced saturation of the photocurrent, that we attribute to negative differential drift velocity, resulting from intervalley scattering [23]. Saturation fields in the 10-20kV/cm range have been found in previous works. Here, at 3.9V (Fig.3(a)) the average electric field is ~100kV/cm, indicating that a large fraction of the applied bias drops on the Schottky contacts.

The photocurrent and responsivity as a function of incident power at 77K and 300K are reported in Fig.3(c), respectively at 3.4V and 2.5V. Responsivities are corrected by the polarization factor (Fig.2(c)), and their value corresponds to the situation where the incident field is polarized orthogonally to the wires, which is the ideal condition to operate the QWIP. At low power we obtain responsivities R = 1.5A/W and 0.15A/W at 77K and 300K. From the reflectivity spectrum of Fig.2(a) (black line) and assuming an intersubband transition energy centered at 107meV with FWHM of 10% (see the spectrum of Fig.1(d)), we find that the responsivity measured at 77K at high bias and low optical power, is compatible with a photoconductive gain g = τe/τre ≈ 2.5, where τe and τre are respectively the electron’s capture and transit time (see Supplementary Material) [24]. The decrease of responsivity at 300K is attributed to a decrease of the drift velocity and capture time (see Table 1). Finally, by increasing the power we observe a clear decrease of responsivity at 77K. This is attributed to the presence of a series resistance provided by the Schottky contacts (Rs in the circuit of Fig.3(d), see next Section and Table 1). As a consequence, for a given applied bias, the decrease of the detector photoresistance (Rs in the circuit of Fig.3(d)) with increasing incident power produces a progressive lowering of the electric field across the QWIP active region [25]. At room temperature Rs is instead negligible (see Table 1), therefore the saturation effect is much less pronounced.
4. HETERODYNE MIXING AND FREQUENCY RESPONSE

In Fig. 4 we report the heterodyne frequency response (FR) of the QWIP in the 10MHz-67GHz range, measured at 77K and 300K, at low and high applied biases. To record these spectra, we used a 67GHz-bandwidth cryogenic probe, positioned at the edge of the coplanar waveguide shown in Fig.1(a). The photodetector is connected to a wideband bias-T and simultaneously illuminated by two 10.3μm-wavelength DFB QCLs driven with ultra-low noise (~300pA/Hz^{1/2}) current generators (see Supplementary Material for a schematic of the experimental setup). The current of one QCL was kept constant while the current and temperature of the second one were fine-tuned in order to sweep the heterodyne frequency in the range 0-67GHz. The powers incident on the QWIP from the two QCLs are $P_1 = 27.5mW$ and $P_2 = 6mW$ (33.5mW total). The spectra of Fig.4 correspond to the intensities of the heterodyne beat signals recorded with a spectrum analyzer (SA) set in max-hold trace mode. The traces are corrected by (i) the propagation losses from the QWIP to the SA measured with a vector network analyzer (VNA), and (ii) the power changes (2dB max) of one QCL due to temperature/current tuning (see Supplementary Material).

The top two traces in Fig.4 show the detector FR at high bias, i.e. 3.4V(77K) and 2.5V(300K). From Fig.2(c), the corresponding responsivities are 0.75A/W and 0.13A/W. At 77K we find a monotonic decrease with frequency, with a 3dB-cutoff frequency of ~30GHz, while at 300K the response is much flatter, with a ~2dB increase from 0 to ~40GHz, followed by a 3dB drop at ~67GHz.

At low biases the shape of the FR is rather different. As shown by the two bottom traces, recorded at 1.1V(77K) and 0.9V(300K), the FR is virtually flat up to 67GHz, except at low frequencies where we observe a pronounced drop below ~5GHz(77K) and ~10GHz(300K).

To gain insight in the behavior of the QWIP, we used a VNA analyzer to derive the device impedance vs frequency in the operating conditions corresponding to the spectra of Fig.4. We find that at low bias (1.1V, and 0.9V spectra in Fig.4) the detector’s RF impedance can be well reproduced using the simple small-signal circuit displayed in Fig.3(d) (see Supplementary Material for the complete derivation) [26]. Here $R_0$, $R_s$ and $Z_c$ ~50Ω represent (i) the dc internal photoresistance of the 5x5 PAR array under illumination, (ii) the leakage resistance of the Schottky contact biased in reverse breakdown (the forward biased Schottky junction is considered as a short circuit), and (iii) the internal load impedance seen by the QWIP detector. $C_{PAR}$ ~30fF and $C_i$ ~1pF, are respectively the 2D PARs array and Schottky contact capacitance. The former corresponds to the computed static capacitance of the PARs array.

The current source in the circuit represents the photocurrent generated in the patch array oscillating at the beat frequency $\omega_b$. It is given by:

$$I_s = \frac{m}{\sqrt{1+(\omega_b t)^2}} I_{ph} \frac{R_d + R_s}{R_d} = \frac{m}{\sqrt{1+(\omega_b t)^2}} I_0,$$

where $m$ is a modulation index given by $m = \frac{2}{\sqrt{R_1 + R_2}} I_{ph} = 0.77$, $I_{ph}$ is the dc (i.e. average) measured photocurrent, $R_d$ is the dark resistance, and $I_0 = I_{ph} (R_d + R_s)/R_d$. The term at the denominator takes into account the frequency roll-off of the intrinsic transport mechanism, with $\tau$ approximating the carriers capture or transit time [24,26].

The solid curves corresponding to the two bottom FRs in Fig.4 represent the power dissipated in the load: $P_L = \overline{\Re} |I_s|^2$. They are computed with (i) $R_0 = 200Ω$, $R_s = 350Ω$, for the spectrum at 1.1V (77K) with $I_{ph} = 0.49mA$ and $\tau = 1$-2ps; and (ii) $R_0 = 75Ω$, $R_s = 125Ω$, for the spectrum at 0.9V (300K) with $I_{ph} = 0.14mA$ and $\tau = 1$-ps (see Table 1 and 2nd column). Despite the fairly simple electrical model and the measurement uncertainties the agreement with the experimental FRs is very good, both in terms of absolute power and spectral shape. In particular the observed drop at low frequency reflects the additional conversion losses due to the heterodyne power dissipated in $R_s$ when $f \leq \frac{1}{2\pi R_s C_s}$ (see Fig.3.d). At higher frequencies $R_s$ is instead shorted by $C_s$, de facto eliminating the power loss in the contact resistance. In this case, from the small-signal circuit model, we have that:

$$I_L = \frac{1}{1 + R_{ISR} + i\omega R_L C_{PAR}}$$

yielding a parasitic roll-off time constant $R_L C_{PAR}[(1 + R_L)/R_s] \lesssim 1$ps (see Table 1). We also find (see Supplementary Material) that, for the chosen PAR array size, the QWIP...
impedance is close to 50Ω for frequencies ≥20GHz (30K) and 30GHz (77K), which is ideal for RF impedance matching.

At high biases the effect of C1 is much less pronounced and the power drop at low frequencies disappears (Fig.4, top two spectra). From the small-signal circuit this can be explained by a reduction of R1 due to the Schottky barrier becoming more transparent, therefore effectively shunting C1 at low frequencies. As a result, the QWIP impedance does not display the strong increase at low frequency found at low biases (see Supplementary Material). At 300K, the QWIP is thus virtually almost impedance matched to 50Ω at all frequencies. From the small-signal circuit

\[ T(K) \quad 77 \quad 300 \quad 77 \quad 300 \]

\[ V_{bias}(V) \quad 1.1 \quad 0.9 \quad 3.4 \quad 25 \]

\[ I_{ph}(mA) \quad 0.49 \quad 0.14 \quad 15.2 \quad 22 \]

\[ I_0(mA) \quad 0.49 \quad 0.38 \quad 15.2 \quad 22 \]

\[ R_s(\Omega) \quad 200 \quad 75 \quad 40 \quad 40 \]

\[ R_s(\Omega) \quad 350 \quad 125 \quad 20 \quad 0 \]

\[ \tau(ps) \quad 1.5 \quad \leq 1 \quad 8 \quad 25 \]

\[ \tau_e(ps) \quad 1 \quad \leq 1 \quad 10 \quad 25 \]

\[ \tau_{tr}(ps) \quad 25 \quad 90 \quad 8 \quad 14 \]

\[ v_d(10/cm/s) \quad 1.5 \quad \geq 0.4 \quad 4.6 \quad 2.6 \]

Table 1. Measured photocurrents (Iph), I0, and small-signal circuit resistances (Rs) and (R1) used to compute the solid lines in Fig.4 under different conditions (bias and temperature). The value of the roll-off time constant (τ in Eq.1) is the one yielding the best fit of the experimental data. The capture time (τc) and transit time (τtr) are obtained from τ and the photoconductive gain (see text). The corresponding drift velocity (vd) is obtained from the ratio between the thickness of the QWIP active region (365nm) and τtr.

we find a good agreement with the measured FRs using (i) R0 = 40Ω, Rs = 20Ω, for the spectrum at 3.4V (77K) with, τ ≈ 8ps and Iph = 15.2mA; and (ii) R0 = 40Ω, Rs = 0, for the spectrum at 2.5V, (300K) with τ ≈ 2.3ps and Iph = 2.2mA (see Table 1, 3rd and 4th column).

As shown above, thanks to the very small device capacitance, by fitting the measured FRs using the small signal circuit model we can extract the intrinsic detector response times, which, as shown by the solid curves in Fig.4, dependent on the operating conditions. From the values of τ and from the photoconductive gain derived from the responsivities, we can then obtain the values of τc and τtr shown in Table 1 (see Supplementary Material): except at 77K under high bias, the QWIP intrinsic response time appears to be dominated by electron capture. We also find the expected decrease/increase of τc with increasing bias/temperature [24]. Interpreting the dependence of τc on bias and temperature is beyond the scope of this work and will require more systematic measurements that are presently underway. At the same time, on this subject there appears to be a lack of experimental data in the literature [4].

In Fig.5 we report two examples of heterodyne beatnote spectra recorded in single shot at 77K, under an applied bias of 3V and without any amplification (Fig.5(a)), and at 300K, with an applied bias of 1.1V and with a narrow band amplifier of 50dB gain (Fig.5(b)). In the first case the noise floor is limited by the spectrum analyzer, while in the second spectrum the noise floor is determined by the amplifiers noise. We find instantaneous linewidths of ~100kHz, limited by the QCL thermal and current fluctuations. At 77K the RBW is set to 100kHz, yielding a SNR of ~77dB, while at 300K we find a SNR of 72dB with a RBW of 50kHz. Reducing further the RBW produces a decrease of the beatnote intensity because the RBW goes below the instantaneous heterodyne beatnote linewidth.

The dependence of the SNR of the heterodyne beatnote frequency, obtained with the spectrum analyzer without amplification (P1 = 27.5mW and P2 = 6mW), can be directly extracted from the spectra recorded in max-hold trace mode (see Supplementary Material). At 30GHz and 60GHz, with a RBW of 3.5MHz, we obtain SNRs of 50dB and 35dB, and of 35dB and 25dB, respectively at 77K (3.4V) and 300K (2.5V).

5. CONCLUSIONS

Antenna-coupled QWIP detectors operating in the 10µm-12µm range are demonstrated, exhibiting a flat frequency response up to 67GHz at 77K and 300K. At 300K, from our experimental results and with the help of an equivalent circuit model, we find an RC-limited 3dB cutoff frequency >150GHz and ~ps intrinsic response times.

We believe that the detectors demonstrated here, in combination with QCLs, will open up new perspectives in MIR photonics, namely by extending to the MIR range the possibilities offered by ultra-fast near-infrared optoelectronics, so far the only frequency range benefiting from the availability of ultrafast photodetectors. Envisaged applications are free space communications with data rates >10Gb/s, coherent multi-species gas sensing, high precision spectroscopy and metrology, astronomy, as well as the study of real time dynamics on the 10ps time scale [8-11,27]. More specifically, on this last topic, we expect that broadband devices such as those demonstrated in this work can shed new light on QWIPs ultrafast electron's dynamics.

A final intriguing perspective is the use of these structures as QCL-pumped photomixers for the generation of sub-THz radiation [28,29]. To this end we note that the actual responsivity of ~0.75A/W at 77K, 3.4V (high power – see Fig.3(c)) could be theoretically brought up to ~4A/W if the ISB
transition peak and the PAR resonance were perfectly matched. In this case, from the top trace in Fig.4, we would expect microwave power levels in the mW range. Replacing the Schottky contacts with non-diffusive ohmic contacts (not to increase MIR photon absorption) should also reduce the saturation of the responsivity at high incident powers (Fig.2(c)).

**Funding.** Nord-Pas de Calais Regional Council; Fonds Européens de Développement Régional. RENATECH (French Network of Major Technology Centres). CPER “Photonics for Society”. French National Research Agency, and Direction Générale de l’Armement (project HISPANID). French National Research Agency (project IRENA). European Union FET-Open Grant MIRBOSE (737017) and the European Research Council (IDEASERC) (“GEM”) (306661). National Natural Science Foundation of China (61875220): “From 0 to 1” Innovation Program of the Chinese Academy of Sciences (ZDBS-LY-JSC009).

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Supplementary Material

Ultrafast quantum-well infrared photodetectors operating at 10µm with flat response up to 70GHz at room-temperature.

M. Hakl,1* Q.Y. Lin,1† S. Lepillette, 1 J-F. Lampin, 1 S. Pirrotta, 2 R. Colombelli, 2 W. J.Wan,3 J. C. Cao, 3 H. Li, 3 E. Peytavit, 1 and S. Barbieri1,†

1Institute of Electronics, Microelectronics and Nanotechnology, Univ. Lille, ISEN, CNRS, UMR 8520, 59652 Villeneuve d’Ascq, France
2Centre de Nanosciences et de Nanotechnologies (C2N), CNRS UMR 9001, Université Paris-Saclay, 91120 Palaiseau, France
3Key Laboratory of Terahertz Solid State Technology, Chinese Academy of Sciences, Shanghai, 200050, China
*Corresponding author: stefano.barbieri@univ-lille.fr
†These authors participated equally to the work.

1. Computed responsivity

The responsivity of the PAR array for an incident electromagnetic wave of frequency ω, polarized perpendicularly to the wire bridges, can be computed from [1–2]:

\[ R_{QWIP}(ω) = [1 - R(ω)] \frac{A_{IAB}(ω)}{A_{IAB}(ω) + Q_{PAR}} \frac{e\gamma}{h_0} \tag{S1} \]

where \([1-R(ω)]\) is the PAR array absorption spectrum shown in Fig2(a) of the main text, \(e\) is the electronic charge, and \(A_{IAB}(ω)\) is the intersubband absorption coefficient of the PAR given by the following expression:

\[ A_{IAB}(ω) = e^{\frac{\omega}{2\hbar}} \frac{\Gamma}{\hbar^2(ω-ω_{21})^2+\Gamma^2/4} \tag{S2} \]

Here \(f_{op} = 0.088\) is the overlap factor between the PAR mode and the quantum wells; \(h_0 = 275\) meV is the intersubband plasma energy [2]; \(\hbar_0 = 107\) meV is the intersubband transition energy extracted from the photocurrent spectrum (see Fig2(d) of the main text), and \(\Gamma = 10.7\) meV is the FWHM of the intersubband transition, which we assume to be approximately equal to 10% of \(\hbar_0\). \(Q_{PAR} \approx 8\) in Eq(S1) is the quality factor of the PAR array (i.e. excluding intersubband absorption), obtained from the FWHM of [1-R(ω)] (Fig2(a) of the main text). This is a good approximation since, due to the spectral shift between \(A_{IAB}(ω)\) and [1-R(ω)], absorption in the resonators is dominated by ohmic losses (i.e. absorption in the metal and contact layers). We note that the value of \(Q_{PAR} \approx 8\) for the array is compatible with results from FDTD simulations performed on a single resonator, yielding an upper limit of \(Q \approx 15\).

The responsivity \(R_{QWIP}(ω)\), obtained from Eq(S1) and (S2) with \(g = 2.5\), is shown in Fig.S1: for \(h_0 \approx 120\) meV, corresponding to the QCL photon energy (\(λ = 10.3\) µm), we obtain \(R_{QWIP} \approx 15\) A/W, in accordance with the measured experimental responsivity at 34V, 77K and low incident power (Fig.3(c) in the main text).

2. Heterodyne mixing experimental setup and frequency response spectra

The schematic of the experimental setup is shown in Fig.S2. To minimize the linewidth of the heterodyne beatnote the QCLs were driven with low noise current drivers (Koheron, DRV110) with a current noise of 300pA/Hz\(^{1/2}\).

Fig.S3 shows the procedure followed to extract the frequency response (FR) spectra displayed in Fig.4 of the main text. As an example we consider the FR at 300K and 2.5V. First, by sweeping the frequency of one QCL, we have recorded the heterodyne beat signal using the spectrum analyzer (SA) set in max-hold trace mode as described in the main text. The corresponding SA trace is displayed in Fig.S3(a) (solid red line). Here, each vertical line corresponds to an heterodyne beat between the two QCLs. Next, with a VNA analyzer, we have measured the power attenuation from the...
QWIP to the SA, due to the insertion loss in the probes, cables and bias-tee (red dots, in dB). Measured attenuation corrected by the QCL power change (blue dots, in dB). (b) Heterodyne beat spectrum obtained by subtracting the blue dotted trace in panel (a) from the raw heterodyne spectrum. The black circle corresponds to the data reported in Fig.4 of the main text (300K, 2.5V).

QWIP to the SA, due to the insertion loss in the probes, cables and bias-tee (red dots, in dB). This curve has been finally corrected to include the power variation of the QCL due to the frequency tuning. The resulting blue dotted curve in Fig.3(a) has been subtracted by the raw heterodyne spectrum, yielding the spectrum shown in Fig.3(b). The black circles, corresponding to the line peaks recorded every 500MHz, are those displayed in Fig.4 of the main text. We note that the heterodyne spectra were recorded with a RBW of 3.5MHz. This is larger than the actual heterodyne beat linewidth, which was of the order of ~100kHz (see the main text), therefore guaranteeing that the intensity of the heterodyne beats is not reduced by filtering.

In Fig.4 we report the other heterodyne spectra (corrected by losses) used to extract the FRs shown in Fig.4 of the main text.

In an optical heterodyne experiment as described in this work, a dc biased photoconductor is illuminated by two laser beams of power $P_1$ and $P_2$ with a difference frequency $\omega_b$. The incident optical power on the photoconductor can be expressed as:

$$P(t) = P_1 + P_2 + 2\sqrt{P_1 P_2} \sin (\omega_b t). \quad (S3)$$

The photocarrier density in the photoconductor follows the time variation of the incident power. By assuming that the photoconductor exhibits a linear I-V characteristic, it can be
modeled as a time-dependent conductance, which can be written as:

\[ G(t) = G_0 + G_1 \sin(\omega_b t + \phi), \quad (S4) \]

where \( G_0 \) and \( G_1 \) are respectively a dc and a dynamic conductance term. They are given by [3]:

\[ G_0 = G_d + G_{ph}, \quad (S5) \]

\[ G_1 = \frac{m}{\sqrt{1+(\omega_b)^2}} G_{ph}, \quad (S5) \]

with

\[ G_{ph} = \frac{I_{ph}}{V_{dc}}. \quad (S6) \]

In Eq.(5) and (6), the term \( G_d (1/R_d) \) is the dark conductance, while \( G_{ph} (1/R_{ph}) \) is the internal photoconductance given by the ratio of the dc (i.e. average) conduction photocurrent, \( I_{ph} \), generated by the laser sources, and the dc bias voltage, \( V_{dc} \), applied to the photoconductor. In the expression of \( G_1 \), the denominator reflects the frequency roll-off of the intrinsic recombination or transport mechanism, \( t \) approximating the carriers capture or transit time. The term \( m \) is a modulation index given by:

\[ m = \frac{2\sqrt{P_1 P_2}}{P_{tot}}, \quad (S7) \]

where \( P_{tot} = P_1 + P_2 \). For the powers used to record the spectra of Fig.4 in the main text (\( P_1 = 27.5 \text{mW}, P_2 = 6 \text{mW} \)) we obtain \( m = 0.77 \).

![Fig.5](image)

**Fig.5**: Electrical circuit model of the heterodyne mixing experiment.

An accurate model of the QWIP detector should include the electrical capacitance of the PARs array, \( C_{PAR} \), in parallel with \( G(t) \). An additional parallel RC circuit should also be added to model the Schottky contact biased in reverse breakdown (the forward biased Schottky junction is considered as a short circuit), with \( R_s \) the junction resistance and \( C_j \) the junction capacitance. The resulting electrical circuit model is shown in Fig.5. This circuit also includes the inductor and capacitor \( (L_{bias-T}, C_{bias-T}) \) of the bias-T used in the experiment, allowing the decoupling between dc and ac currents. Finally, \( Z_L \) is the load impedance seen by the QWIP (see Section 4). Voltages and currents are time periodic (period \( T = \frac{2\pi}{\omega_b} \)) and can be expressed in a Fourier series. By neglecting high orders harmonics [3], the voltage across the photoconductor takes the simple form:

\[ V(t) = V_{dc} + V_{ac} \cos(\omega_b t + \phi), \quad (S8) \]

and the current:

\[ I(t) = I_{dc} + I_{ac} \cos(\omega_b t + \delta). \quad (S9) \]

The quantities \( V_{dc}, V_{ac}, I_{dc}, I_{ac}, \phi, \delta \) can be derived from the circuit of Fig.5 by applying Kirchoff’s laws at \( \omega = 0 \) and \( \omega = \omega_b \) (i.e. exploiting the decoupling between dc and ac currents thanks to the bias-T inductance and capacitance) and by using the constitutive relation:

\[ I(t) = G(t) V(t). \quad (S10) \]

From the equations above, an ac small-signal circuit \( (\omega = \omega_b) \) and a dc circuit \( (\omega = 0) \) can be derived, as shown in Fig.6(a),(b) [3].

![Fig.6](image)

**Fig.6**: (a) Equivalent small-signal ac circuit \((\omega = \omega_b)\), (b) Equivalent dc circuit \((\omega = 0)\).

In the ac circuit (Fig.6(a)) the QWIP detector is modeled by an equivalent ac current source, \( I_s \), with an internal impedance consisting of its dc photoreistance under illumination,

\[ R_0 = \frac{r_{ph}}{r_d^* + r_{ph}^*}, \quad (S11) \]

in parallel with the intrinsic capacitance of the PAR array \( C_{PAR} \). The current source can be computed as [3]:

\[ I_s = V_{dc} \times G_1 = \frac{m}{\sqrt{1+(\omega_b)^2}} V_{dc} \times \frac{1}{r_{ph}}, \quad (S12) \]

where \( I_s \) is in general a phasor (from now on we assume that all currents and voltages are represented by phasors). The dc equivalent circuit (Fig.6(b)), can be used to derive \( V_{dc} \):

\[ V_{dc} = \frac{V_{bias}}{r_d^* + r_{ph}^*} R_0 = I_{dc} R_0. \quad (S13) \]
From this equation we note that, due to the Schottky contact resistance, $V_{dc} < V_{bias}$. From Eqs. (S11) to (S13) we obtain:

$$I_z = \frac{m}{\sqrt{1+(\omega a)^2}} I_{dc} \frac{R_{ph}}{R_d} = \frac{m}{\sqrt{1+(\omega a)^2}} I_{dc} \frac{R_{ph}}{R_d}$$ (S14)

where $I_{ph}$ is the dark current under illumination that can be measured experimentally. It is also useful to express the current source $I_z$ as a function of the $dc$ photocurrent of the QWIP, which is obtained by subtracting the dark current from $I_{dc}$. From the $dc$ equivalent circuit of Fig. S6(b) the photocurrent is given by:

$$I_{ph} = I_{dc} - I_{dark} = V_{bias} \left[ \frac{1}{R_0 + R_s} - \frac{1}{R_d + R_s} \right] = I_{dc} \frac{R_g - R_0}{R_d + R_s} = I_{dc} \frac{R_g - R_0}{R_d + R_s}$$ (S15)

From this last equation we find that the $dc$ photocurrent is equal to the measured $dc$ current only if $R_0$ and $R_s$ are negligible compared to $R_d$. As can be deduced from Fig. 3(a)(b) in the main text, this is the case at 77K for sufficiently high power levels, but not at 300K, where the correction factor $(R_d + R_s)/R_d$ (Eq. S16) cannot be neglected. Finally, by comparing Eq. (S14) and (S15) we obtain:

$$I_z = \frac{m}{\sqrt{1+(\omega a)^2}} I_{ph} \frac{R_g + R_0}{R_d}.$$ (S16)

4. QWP impedance measurements

In Fig. S7 we report the real and imaginary parts of the QWP impedances vs frequency (black and red lines) obtained from the $S_{11}$ parameters measured with a VNA analyzer, after de-embedding the 50Ω integrated coplanar line. At T=77K the $S_{11}$ parameters were measured under the same conditions (bias, temperature and illumination) used to record the FRs, while at 300K they were measured in the dark. This last choice stems from the fact that, contrary to 77K, at 300K the dark current is much larger than the photocurrent even under illumination at high power (see Fig. 3(b) of the main text), i.e. the QWP impedance under illumination is very well approximated by the dark impedance ($R_d < R_{ph}$, see Fig. S6(a)).

The impedances at low biases (Fig. S7(a)(b)) are well reproduced by the equivalent circuit of Fig. S6(a), where the QWP impedance (blue lines) is given by the sum of the PAR array and Schottky contact impedances:

$$Z_{QWP}(\omega) = \frac{R_0}{1+i\omega R_0 C_{PAR}} + \frac{R_s}{1+i\omega R_s C},$$ (S17)

In Fig. S7(a)(b), $Z_{QWP}(\omega)$ is computed using the values of $R_0$, $R_s$ reported in the first and second column of Table 1 in the main text, with $C_{PAR} = 30\Omega$ and $C = 0.7\mu\Omega$ (see next Section). In particular, when $f \rightarrow 0$, we see clearly the effect of $C$, producing a fast increase of the real part of $Re[Z_{QWP}(\omega)]$, until, at $f = 0$, $Re[Z_{QWP}(\omega = 0)] = R_0 + R_s$. At higher frequencies ($f_s \gg (2\pi R_s C)^{-1}$), $R_s$ is shunted by $C_s$ and $Z_{QWP}(\omega)$ coincides with the impedance of the PAR array, with a roll-off corresponding to a time constant equal to $R_s C_{PAR}$ (first term in Eq. S17).

As shown in Fig. S7(c), at high biases the QWP impedances change completely. Firstly, the fast increase as $f \rightarrow 0$, disappears, which we interpret as the evidence that the Schottky junction becomes more transparent, i.e. $R_s$ shunts $C_s$ at all frequencies (see the next Section). At higher frequencies both the real and imaginary parts of $Z_{QWP}(\omega)$ show a maximum, followed by a slow decay. As shown by the blue lines this behavior cannot be fully reproduced by our simple circuit model using the parameters reported in the third and fourth column of Table 1 of the main text. In particular the imaginary part becomes inductive around 15-30 GHz. This phenomenon is probably linked to the fact that the QWP is operated close to the onset of the intervalley scattering. A more detailed analysis is needed which is beyond the scope of this article.

In Fig. S8 we report the measured load impedance, $Z_L$, i.e. the impedance seen by the QWP in the plane of the coplanar probes. This was extracted from $S_{11}$ parameter measurements. As can be seen, $Z_L$ can be approximated by its real part $Re[Z_L] = R_L \approx 50\Omega$. 

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**Fig. S7.** Real (black) and imaginary (red) parts of the QWIP impedance, extracted from the $S_{11}$ parameters measurements, after de-embedding the 50Ω integrated coplanar line. The measurements at 77K (panels (a), (c)) were recorded under illumination with a power $P_{ill} = P_1 + P_2 = 33.5\mu\text{W}$. The measurements at 300K (panels (b),(d)) were done in the dark. The blue and purple lines (see text) represent the impedance computed from the small-signal equivalent circuit of Fig. S6(a) using the values of the resistances shown in the legends, with $C_{PAR} = 30\Omega$ and $C = 0.7\mu\Omega$.

**Fig. S8.** Real and imaginary part of the impedance seen by the QWIP in the plane of the coplanar probes ($Z_L$).
5. Determination of QWIP circuit parameters

The various elements, $R_p$, $R_s$, $C_{PAB}$ and $C_s$ in the circuit of Fig.S6 depend in principle on the QWIP operating temperature, bias, and illumination conditions. To determine their values we rely on the experimental FR spectra displayed in Fig.4 of the main text and on the corresponding QWIP impedances shown in Fig.S7.

The first equation used to determine $R_0$, $R_s$ is given by (see Fig.S6):

$$ R_0 + R_s = \text{Re}[Z_{QWIP}(\omega = 0)]. \quad (S18) $$

The second equation is obtained instead by noting that the values of the experimental FRs in Fig.4 of the main text, correspond to the power dissipated in $Z_L$:

$$ P_L = \frac{1}{2} \text{Re}[|Z_L|^2] = \frac{1}{2} R_L \cdot |I_L|^2. \quad (S19). $$

Indeed, we actually find that the FRs are corrected by the attenuation from the QWIP to the SA which, in turn, has an impedance of 50 Ohm, i.e. perfectly adapted to $Z_L$ (Fig.S8). Since $P_L$ in Eq.(S19) depends on $R_s$ and $R_p$, by comparing it with the power levels in the FRs of Fig.4, we find that the second equation, which, together with Eq.(S18), allows the determination of the QWIP and Schottky resistances separately.

Concerning the capacitances, we begin by fixing $C_{PAB} = 28 fF$, which corresponds to the computed static capacitance of the PAR array using a parallel plate model. $C_s$ is instead determined by fitting the decay of the experimental FRs at low frequency.

The details of the calculations are given below, respectively at low bias (1.1V, 77K, and 0.9V, 300K) and high bias (3.4V, 77K, and 2.5V, 300K). The values of the measured $dc$ photocurrent of the QWIP, $I_{phs}$, and the values of $R_0$ and $R_s$ can be found in Table 1 of the main text.

**Low bias**

We start by assuming to be at a sufficiently high frequency such that $R_s$ is shunted by the parallel capacitance $C_s$ ($f_0 \gg (2\pi R_s C_s)^{-1}$) and can therefore be neglected (see Eq.(S17)). In this case we obtain:

$$ I_L = I_s \frac{1}{1 + j \omega R_s} \frac{1}{1 + j \omega R_s C_{PAB}}, \quad (S20) $$

where we have approximated $Z_L$ with its real part $R_s \approx 50\Omega$ (see Fig.S8). Now, provided that the frequency is not too high, e.g. $f_0 \approx 10$GHz, the last term at the denominator can also be neglected thanks to the extremely low value of $C_{PAB}$ (note: the validity of these last two assumptions can be verified a posteriori from the values of $R_0$ and $R_s$). The power dissipated into the load is then given by:

$$ P_L = \frac{1}{2} \text{Re}[R_L \left( \frac{R_0}{R_0 + R_s} \right)^2]. \quad (S21) $$

At $T = 77$K we have that $R_s \gg R_s$ (see Fig.3(a) in the main text), hence, from Eq.(S16), we have that $I_s = \frac{1}{\mu} I_{ph} = 0.53mA$ (at $f_0 \approx 10$GHz, $\omega_2 \approx 77$K). At this point Eq.(S21) can be used to determine the value of $R_0$ by comparing $P_L$ with the measured value of the FR at 10GHz (1.1V, 77K curve in Fig.4 of the main text). The value of $R_0$ can finally be obtained from Eq.(S18) with $Re[Z_{QWIP}(\omega = 0)] = 550\Omega$ (see Fig.S6(a)). We find $R_0 = 2000\Omega$ and $R_s = 35\Omega$ (first column of Table 1).

The last step consists in determining the value of $C_s$. This is obtained by fitting the decay of the experimental FR at low frequency (see Fig.4 in the main text), yielding $C_s = 0.7pF$. We note that this value is in agreement with the theoretical capacitance expected for a Au/GaAs Schottky junction with a doping density of $4\times10^{10} cm^{-3}$ (~15nm depletion region width) [4]. The computed QWIP impedance is represented by the blue curves in Fig.S6(a), showing a good agreement with the impedance derived from the $S_1$ parameter. Also, the computed FR using Eq.(S18) reproduces very well the experimental one as shown in Fig.4 for $\tau \approx 1ps$.

Concerning the measurement at $T = 300K$ and 0.9V, by comparing the dark current and the photocurrent $I/V$ characteristics in Fig.3(b) of the main text, we can clearly see that $R_s \ll R_{ph}$, i.e. $R_s \approx R_s$ from Eq.(S11). At the same time, contrary to 77K, $R_s$ cannot a priori be neglected compared to $R_s$. In this case, from Eq.(S16), Eq.(S19) and Eq.(S21) we obtain:

$$ P_L = \frac{1}{2} \text{Re}\left[ R_L \left( \frac{R_0 + R_s}{R_0 + R_s + R_{ph}} \right)^2 \right], \quad (S22) $$

where again, we used the fact that $f_0 \approx 10GHz$, $\omega_2 \approx 77K$. In this last equation the term $R_0 + R_s$ is known from Eq.(S18) and Fig.S7(b) ($R_0 + R_s = Re[Z_{QWIP}(\omega = 0)] = 2000$). Then, again, $R_s$ is determined by comparing $P_L$ in Eq.(S22) with the measured value of the FR at 0.9V (0.3V, 300K curve in Fig.4 of the main text). From this procedure we obtain $R_0 = 105\Omega$ and $R_s = 95\Omega$, which, however, do not allow to reproduce the QWIP impedance in a satisfactory way, as shown by the purple traces in Fig.S7(b) (here we used $C_s = 0.7pF$). We find that the values $R_0 = 75\Omega$ and $R_s = 125\Omega$, allow to obtain the closest agreement with $P_L$ compatible with a good fit of the QWIP impedance (blue traces in Fig.S7(b)). The resulting computed FR, shown in Fig.4 of the main text, is ~2dBm above the measured FR. This spectrum was obtained with $C_s = 0.7pF$, yielding, as for the 77K,1.1V FR, a decay at low frequency in good agreement with the measurement.

**High bias**

As already pointed out, at high bias we don't observe anymore the drop in the FR as $f \rightarrow 0$. In other words, $C_s$ is shunted by $R_s$, which can be taken as the Schottky contact impedance at virtually all frequencies. As we did at low bias, we also assume that the frequency is sufficiently low that $2\pi f R_s C_{PAR} \ll 1$ (e.g. $f \approx 1GHz$). Under these assumptions we have that with $I_L = I_s \times \frac{R_0}{(R_0 + R_s + R_{ph})}$, yielding:

$$ P_L = \frac{1}{2} \text{Re}[R_L \left( \frac{R_0}{R_0 + R_s + R_{ph}} \right)^2], \quad (S23) $$

where, as usual, $R_0 + R_s = Re[Z_{QWIP}(\omega = 0)]$. At $T = 77K$ we have that $I_s = 11.7mA$, and $R_0 + R_s = 600\Omega$ (see Fig.S7(c)). By using $P_L$ from Eq.(S23) to fit the value of the measured FR at 1GHz (3.4V, 77K curve in Fig.4 of the main text) we obtain $R_0 = 40\Omega$ and $R_s = 20\Omega$. As shown in Fig.4 of the main text, from Eq.(S19) we obtain an excellent agreement with the measured FR using $\tau \approx 8ps$.

At $T = 300K$ we still have $R_s \ll R_{ph}$, i.e. $R_s \approx R_0$. Hence, from Eq.(S16) and (S23) we have:
\[ P_L = \frac{1}{2} m^2 f_{ph}^2 \left( \frac{R_0 + R_s}{R_0 + R_s + R_L} \right)^2 R_L \]  
(S24)

with \( R_0 + R_s = \text{Re}[Z_{QWIP}(\omega = 0)] = 40 \Omega \) (see FigS7(d)). Using \( m \times I_{ph} = 1.7 \text{ mA} \), the best agreement with the measured FR is obtained with \( R_0 = 40 \Omega \) and \( R_s = 0 \) (Fig 4 in the main text).

As shown in FigS7(c), (d), contrary to what happens at low bias, the computed impedances at high bias provide only an approximated value of the actual QWIP impedance (see Section 4).

6. Evaluation of carriers capture and transit times

77K, 3.4V.
From the responsivity reported in Fig3(c) of the main text at \( P_{tot} = P_1 + P_2 = 33.5 \text{ mW} \), we obtain \( g(77K, 3.4V) = \tau_r / \tau_c \approx 1.25 \) (i.e. ~ half the value at low incident power \( g \approx 2.5 \) — see Section 1). The roll-off time constant \( \tau \) can therefore be approximated by the transit time \[ \text{[5]. From the fit of Fig 4 in the main text, we then have that } \tau = \tau_r \approx 8 \text{ps} \text{ and } \tau_c = 1.25 \times \tau_r \approx 10 \text{ps} \]

77K, 1.1V.
We have that \( I_{ph} \propto g \), therefore (see Table 1 in the main text):

\[ g(77K, 1.1V) = g(77K, 3.4V) \times \frac{I_{ph}(77K, 1.1V)}{I_{ph}(77K, 3.4V)} \]

\[ \approx 125 \times \frac{0.49 \text{mA}}{15.2 \text{mA}} = 0.04 \]  
(S25)

The roll-off time constant \( \tau \) can therefore be approximated by the capture time. From the fit of Fig 4 in the main text, we then have that \( \tau = \tau_c \approx 1 \text{ps} \text{ and } \tau_r = \tau_c / 0.04 \approx 25 \text{ps} \).

300K, 0.9V.
Following the same procedure described above we obtain \( g(300K, 0.9V) \approx 0.011 \). From the fit of Fig 4 in the main text, we then have that \( \tau = \tau_c \approx 1 \text{ps} \text{ and } \tau_r = \tau_c / 0.011 \approx 90 \text{ps} \).

300K, 2.5V.
Following the same procedure described above we obtain \( g(300K, 2.5V) \approx 0.18 \). From the fit of Fig 4 in the main text, we then have that \( \tau = \tau_c \approx 2.5 \text{ps} \text{ and } \tau_r \approx \tau_c / 0.18 \approx 14 \text{ps} \).

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