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Battery-free ambient FM backscatter nodes reaching 150 metres through digital chirp synthesis and MISO diversity reception

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ABSTRACT

Wireless sensors are often placed where changing a battery is costly or impractical. Powering them from harvested energy is attractive but a conventional radio spends milliwatts to generate a carrier. Ambient backscatter removes that cost by reflecting existing broadcast signals instead of creating new ones. This work turns frequency modulation (FM) broadcast stations up to 34.5 km away into free carriers for a battery-free sensor node that uses Long Range (LoRa) style chirp spread spectrum. The chirps are synthesised in the digital domain by the inter-IC sound (I²S) peripheral of a commercial microcontroller while its processor sleeps. No digital-to-analog converter or voltage-controlled oscillator is needed and a single radio-frequency switch performs the modulation. At a 1% duty cycle the node draws only 22 μ W yet its packets reach 150 metres. A multiple-input single-output receiver built from low-cost software-defined radios captures the same packet reflected from many FM stations at once. Fusing these copies in the frequency domain with soft-decision decoding lowers the bit error rate. Solar harvesting and a supercapacitor power the 45 mm $\times$ 50 mm node along with its environmental sensors plus motion and magnetic sensors. The result is a practical path toward maintenance-free wireless sensing.

Introduction

The proliferation of wireless sensor networks underpinning the Internet of Things (IoT)¹ and fifth-generation (5G)² infrastructure has introduced a fundamental tension between scale and sustainability. Deploying billions of self-maintaining sensor nodes in urban environments, industry, and wearables demands communication approaches that go beyond conventional transmitters³, which employ power amplifiers, oscillators, and frequency synthesisers that consume constant power in the milliwatt-to-watt range. A battery lasts only a few weeks when power is consumed at milliwatt levels. Using batteries at this scale becomes prohibitively costly. The environmental impact^{4,5} presents a further challenge, as the disposal of millions of batteries poses significant harm to the environment. Backscatter communication^{6–9} provides a solution based on the theoretical foundations laid by Stockman¹⁰ in 1948. Rather than producing a radio frequency (RF) carrier, the backscatter node changes impedance at the antenna and reflects the incident signal to communicate data. This eliminates the high-energy-consuming radio parts, reducing energy consumption to the level of a logic switch and paving the way for battery-free communications.

Battery-free energy harvesting requires the node to source energy from the environment¹¹, while the communication power must remain within the budget provided by the harvested energy. In an indoor setting or with intermittent sunlight exposure, harvested solar power typically reaches tens to hundreds of microwatts¹². Energy harvesting from thermal¹³, ambient RF¹⁴, or mechanical sources¹⁵ offers even lower power budgets than photovoltaics. Any communication consuming milliwatt-level power would be impractical, as it would require larger harvesters, making massive deployment infeasible.

Backscatter communication began with radio frequency identification (RFID) applications using passive ultra-high frequency (UHF) tags^{16,17} communicating with readers over a few metres at data rates of tens to hundreds of kbps. Ambient backscatter eliminates the need for a dedicated RF source, making the system much simpler and enabling massive deployments. This technology was first demonstrated using digital television broadcast transmissions¹⁸ as the ambient carrier source. This led to exploitation of other ambient RF sources such as WiFi¹⁹, Bluetooth²⁰, Long Range (LoRa)^{21,22}, and cellular²³. In addition, parallel efforts on frequency modulation (FM) backscatter^{24–29} were made for backscatter communications, especially for the long-range coverage of FM stations often spanning more than 50 km, demonstrating that the FM broadcast network is a good ambient RF source. Early ambient FM backscatter systems relied on simpler modulation techniques such as on-off keying (OOK)²⁴, two-level frequency shift keying (2-FSK)²⁵, four-level pulse amplitude-modulated signal (4-PAM)²⁹, and binary

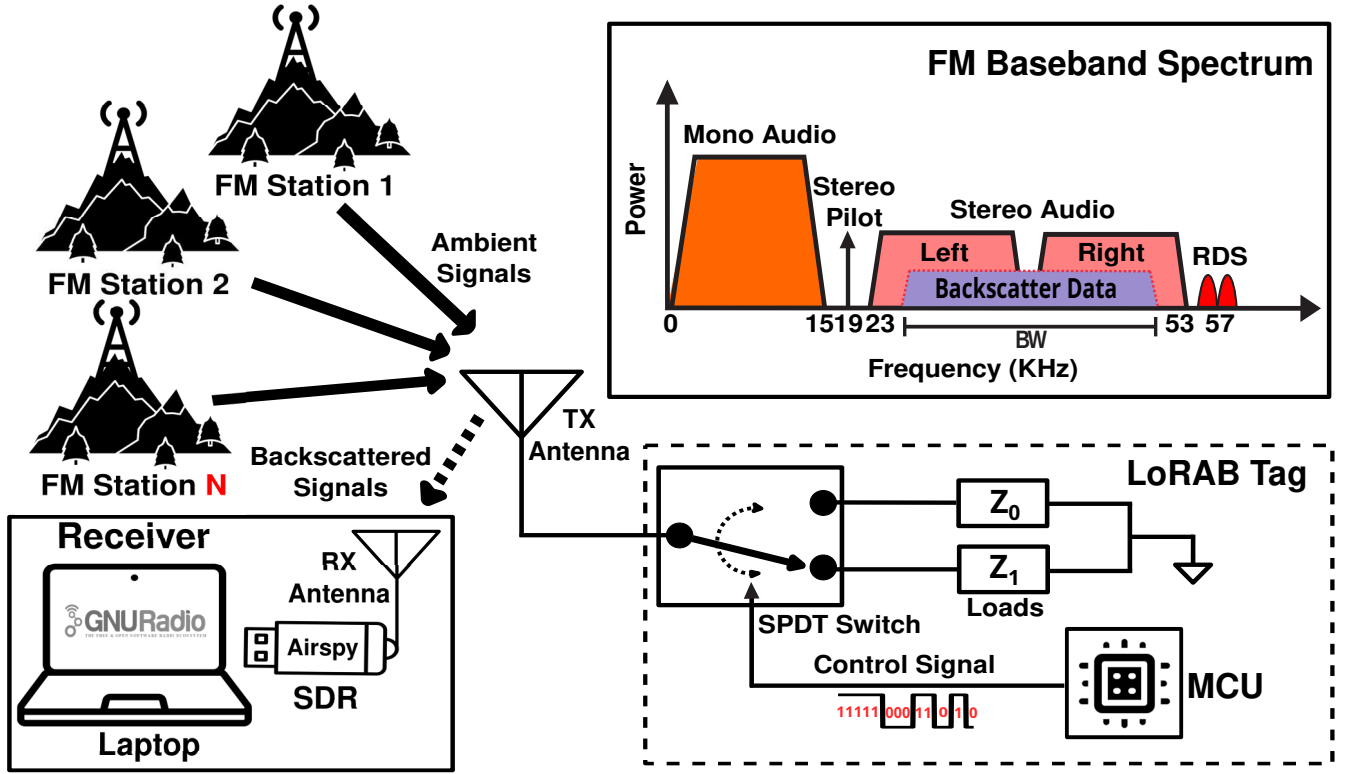


Figure 1. LoRAB 2.0 system overview. Multiple ambient FM broadcast stations illuminate the tag, which comprises a transmit antenna, an ADG919³⁰ single-pole double-throw (SPDT) switch alternating between impedance loads Z_0 and Z_1 , and an nRF52833³¹ microcontroller unit (MCU) generating the binary control signal. Backscattered chirp spread spectrum (CSS) signals are received by an Airspy Mini³² software-defined radio (SDR) connected to a laptop running GNU Radio. The inset shows the FM baseband spectrum, with backscatter data embedded within the 23–53 kHz stereo subcarrier band.

phase-shift keying (S-BPSK)²⁶; however, their range remained below 30 m.

Increasing the operating range of backscatter systems at an extremely efficient power level has resulted in two parallel areas of research. First, rather than OOK²⁴ or simple 2-FSK^{25,33}, chirp spread spectrum (CSS)^{27,28} waveform is used, which is similar to LoRa’s physical layer (PHY); thus, it offers impressive processing gain at the receiving end, enabling successful reception in the presence of a signal strength well below what narrow-band protocols could tolerate. The key concept is the spreading factor (SF), which controls the trade-off between data rate and link budget performance, providing an option for low data rate applications with ultra-low powered nodes at the edge of coverage. In earlier research on LoRa Backscatter³⁴, CSS was successfully transmitted over a single-tone carrier, reaching ranges up to 2.8 kilometres in proximity to the signal source. However, a complete analog transmitter design was required for a backscatter node, including digital-to-analog conversion (DAC), a voltage-controlled oscillator (VCO), mapping circuitry, and an RF multiplexer operating single pole, eight throw (SP8T) switch. A subsequent simplified baseband generator implementation³⁵ relied on a field-programmable gate array (FPGA) driving the switches; however, it involves complex register-transfer level (RTL) code and thus requires specialised hardware development expertise and dedicated FPGA infrastructure, whereas it is possible to generate the same CSS modulation with only a commercial MCU and a single SPDT RF switch. Second, FM stations can experience blackouts, local multipath effects, or high noise in the carrier when music is played. Using backscatter signals from multiple FM sources²⁸ simultaneously enables redundancy against such events and enhances received energy by means of combining, improving error probability as the number of diversity branches increases. Our preliminary conference work²⁸ combined digital CSS synthesis with a MISO receiver on a microcontroller development board. The present article extends that work with a custom self-powered PCB that integrates environmental, motion and magnetic sensors, solar energy harvesting with supercapacitor storage, a receiver with carrier frequency offset (CFO) and sampling time offset (STO) correction, and an outdoor field test at distances up to 150 metres.

Figure 1 summarises the proposed sensor node, named LoRAB 2.0²⁸, which unifies both directions in a single minimal-hardware platform. CSS chirps, similar to LoRa format, are generated in the digital domain using the inter-IC sound (I²S)

interface of a commercially available nRF52833³¹ microcontroller. Lookup table samples are streamed independently by the direct memory access (DMA) controller while the central processing unit (CPU) remains in deep sleep mode. Neither a DAC, a VCO, nor complex RF circuitry is used in the tag. A distinctive feature is that the tag simultaneously impresses CSS-modulated backscatter onto multiple ambient FM carriers, originating from broadcast stations up to 34.5 km distant, using a single ADG919³⁰ SPDT switch driven by the I²S bitstream and alternating between two antenna impedance states. A multiple-input single-output (MISO) receiver, implemented with portable, low-cost SDRs, processes these backscattered replicas by combining their frequency-domain components followed by soft-decision decoding and achieves a considerable reduction in bit error rate compared to single-station reception. Here, the multiple inputs are the FM transmitters, each of which delivers an independent backscattered replica of the same packet, while the single output is the one receive antenna and SDR at which these replicas are captured and combined, so the diversity arises from the several FM carriers rather than from multiple antennas. The device features environmental, motion, and magnetic sensors, solar-powered energy harvesting, and a supercapacitor-based power storage module miniaturised to a 45 × 50 mm form factor, enabling energy-autonomous operation. Measurements performed outdoors have shown that packets were successfully received up to a distance of 150 metres, using an energy budget of just 22 μ W.

Digital chirp synthesis for two-component backscatter modulation

LoRAB 2.0's operation hinges on the principle that the CSS chirp modulation, despite being considered complex due to the continuously varying frequencies in the waveform, can be described as a binary square wave when considering the behaviour of backscatter. This is because the transitions in the square wave are decided based on the zero crossings of the chirp baseband waveform. The ideal up-chirp and down-chirp waveforms for LoRa are:

$$s_{\uparrow}(t) = \exp\left(j2\pi\left(f_0t + \frac{B}{2T}t^2\right)\right), \quad 0 \leq t < T \quad (1)$$

$$s_{\downarrow}(t) = \exp\left(j2\pi\left(f_0t - \frac{B}{2T}t^2\right)\right), \quad 0 \leq t < T \quad (2)$$

The starting frequency of the chirp is referred as f_0 , its bandwidth as B , and $T = 2^{\text{SF}}/B$ as the symbol duration determined by the spreading factor SF. In an up-chirp the instantaneous frequency rises linearly from f_0 to $f_0 + B$ over one symbol interval, and in a down-chirp it falls linearly from f_0 to $f_0 - B$. As shown in Figure 2 (left), the digital square-wave representation of the chirp waveform produces a fundamental spectral component whose amplitude is scaled by a factor of $4/\pi$ relative to a sine wave of the same peak amplitude, corresponding to an increase of approximately 27.3%. However, it also introduces additional harmonic components at frequencies that are odd multiples of the instantaneous chirp frequency. Under typical conditions, such harmonics would significantly increase out-of-band emissions in conventional radio systems. In the case of ambient backscatter communication, however, the impact of these harmonic components is considerably reduced due to the low signal amplitudes involved. The ambient FM carrier received at the tag antenna has a power level of about -60 dBm. The fundamental CSS component backscattered by the tag and received at a distance of 1 m has a lower power level of approximately -80 dBm. The n -th odd harmonic of a square wave has $1/n$ of the amplitude of the fundamental, so the strongest harmonic, the third, lies about 9.5 dB below the already weak fundamental and effectively merges into the measurement noise floor.

The method through which the tag transfers the chirp pattern onto the incoming FM field is the impedance modulation of the backscatter antenna, as illustrated in Figure 1. The tag's complex reflection coefficient in response to the incident wave is given by

$$\Gamma = \frac{Z_L - Z_{\text{ref}}}{Z_L + Z_{\text{ref}}} \quad (3)$$

where $Z_{\text{ref}} = 50\Omega$ denotes the reference impedance of the antenna port and Z_L represents the load impedance controlled by the ADG919³⁰. Figure 2 (centre) illustrates the measured reflection coefficients Γ_0 and Γ_1 on the Smith chart, referenced to the same 50Ω impedance over the FM broadcast band from 88 to 110 MHz. The median impedance states of the ADG919³⁰ in this range result in reflection coefficients of $\Gamma_1 \approx 0.832 - j0.531$ and $\Gamma_0 \approx -0.775 + j0.130$, respectively. With $|\Gamma_1| \approx 0.987$, the first state lies almost on the unit circle, while $|\Gamma_0| \approx 0.786$, and the two points are separated by a substantial arc on the Smith chart, indicating a sharp contrast between the two states. This variation in the reflection coefficients directly determines the modulation depth of the backscattered signal.

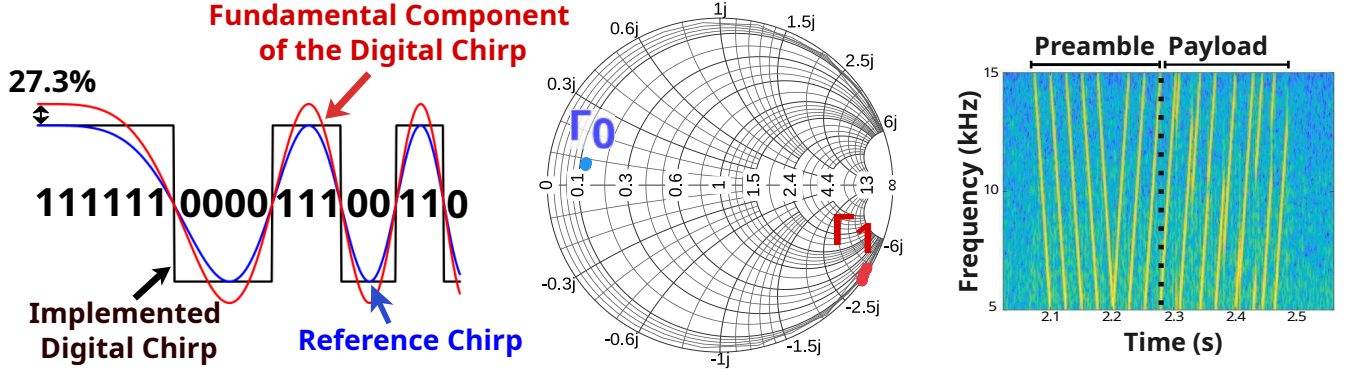


Figure 2. Left: Reference chirp (blue), its binary digital approximation (black), and the fundamental component of the digital chirp (red), illustrating the 27.3% amplitude increase relative to the original analog reference. Centre: Smith chart of the measured ADG919³⁰ reflection coefficient states Γ_0 and Γ_1 across 88–110 MHz. Right: Received CSS frame spectrogram showing the preamble and payload regions.

$$\Delta\Gamma = |\Gamma_1 - \Gamma_0| \quad (4)$$

yielding $\Delta\Gamma \approx 1.74$ in the current implementation. The $\Delta\Gamma$ value establishes the strength of the information-carrying backscattered part compared to the antenna's structural scattering baseline, and maximising it is the main goal of tag antenna and switch impedance design.

The lookup tables for both up-chirp and down-chirp sequences are pre-calculated and stored in flash memory, and later loaded into the random-access memory (RAM) of the nRF52833³¹ microcontroller, since its EasyDMA controller can only read from RAM. Each table entry represents the binary state of the I²S data line at a given sampling instant, determined by evaluating the sign of the real component of the ideal chirp waveform at the I²S clock frequency. During transmission, the DMA controller independently streams these samples to the I²S peripheral without processor intervention, generating a serial bitstream that accurately follows the instantaneous frequency evolution of the chirp. The CPU remains in deep sleep during this process, waking only at packet boundaries to advance the frame state machine, which largely explains the observed low average current of approximately 12.2 μ A. Although I²S is employed here for its convenient clocked serial output, the same binary lookup-table approach is directly applicable to any microcontroller peripheral capable of generating a clocked bitstream at the required bandwidth, including serial peripheral interface (SPI), universal asynchronous receiver-transmitter (UART), inter-integrated circuit (I²C), or pulse-width modulation (PWM) outputs. The I²S clock frequency can be configured, and the current implementation supports chirp bandwidths of 7.5 kHz, 10 kHz, and 15 kHz. At these bandwidths, the lookup table required for a single symbol contains $(2^{\text{SF}} F_s)/B$ samples, where $F_s = 400$ kHz denotes the I²S sampling rate, obtained from a 3.2 MHz master clock with a ratio of 256. The nRF52833³¹ provides 128 kB of RAM, sufficient to accommodate lookup tables for both up-chirp and down-chirp simultaneously. This memory limitation restricts the highest usable spreading factor to SF 9, as SF 10 and higher would require lookup tables exceeding the available RAM for the given bandwidths and are therefore not supported in the current implementation. Within the supported range of SF 5 to SF 9 and across the three supported bandwidths, and with a coding rate of 4/8, LoRAB 2.0 achieves data rates ranging from approximately 70 bps at SF 9 with a bandwidth of 7.5 kHz (corresponding to one packet every 90 s), to about 1.2 kbps at SF 5 with a bandwidth of 15 kHz (corresponding to one packet every 3 s). While the tag hardware can theoretically support higher data rates, broader bandwidths, and larger spreading factors (at higher bandwidths), the generated CSS waveforms are intentionally confined to the FM stereo subcarrier region as shown in Figure 1 (inset), which spans 23 to 53 kHz above each FM station's centre frequency. Restricting operation to this region, as identified in earlier LoRAB studies^{27,28}, achieves two key objectives simultaneously:

- It keeps the backscattered fundamental away from the primary FM audio channel. The odd harmonics of the 23 to 53 kHz chirp fall 69 kHz or further from the carrier and can reach the edge of the adjacent channel at 200 kHz, but they are at least 9.5 dB weaker than the already weak fundamental, which limits the risk of degrading broadcast quality.
- It allows potential reception using commercial FM receivers equipped with radio data system (RDS) or stereo decoding capabilities, since the backscattered signal resides within the same spectral region as the stereo channel.

The tag transmits a frame format based on the LoRa packet structure, featuring a preamble of five up-chirps followed by three down-chirps for synchronisation, along with payload symbols encoded using the selected spreading factor and coding

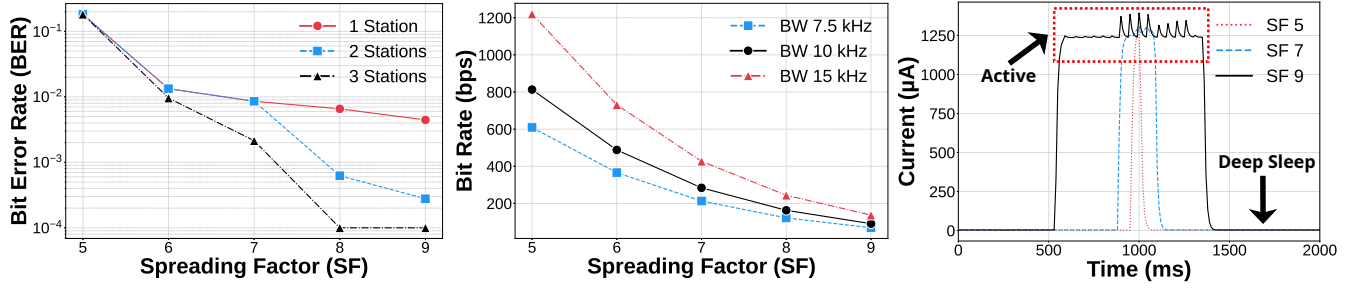


Figure 3. Left: Indoor BER vs. SF at 10 m for 1, 2, and 3 station combining. Centre: Achievable bit rate vs. SF for three chirp bandwidths. Right: Measured current profile showing deep-sleep and active transmission intervals for SF 5, 7, and 9.

rate. Figure 2 (right) shows the spectrogram of a received CSS frame, clearly highlighting the preamble and payload regions. The example uses SF 7 in a 10 kHz bandwidth occupying 43–53 kHz above the FM carrier, downconverted to a 48 kHz audio baseband such that the chirp sweeps from 5 kHz to 15 kHz. A fixed duty cycle of 1% was chosen to keep the tag within the harvested energy budget, and it constrains the average power consumption to $22 \mu W$, independent of the selected spreading factor. This is because the active transmission duration scales with the symbol air-time, while the deep-sleep baseline current remains unchanged. This duty-cycling strategy ensures that channel utilisation per node remains sufficiently low, enabling the simultaneous operation of multiple LoRAB 2.0 tags on the same FM carrier without mutual interference.

MISO diversity reception across ambient FM stations

The receiver architecture of LoRAB 2.0 exploits the fact that the CSS-modulated backscatter signal produced by the tag is simultaneously impressed upon every ambient FM carrier within reception range, generating independent replicas of the transmitted packet on each station frequency. These replicas experience independent noise realisations, independent stereo subcarrier interference profiles, and independent fading conditions, making them well suited to diversity combining. Rather than treating co-channel FM stations as interference, the MISO receiver captures and combines them to reduce bit error rate and provide resilience against station-level outage.

The receiver front-end is an Airspy Mini³² SDR, selected for its 6 MHz instantaneous bandwidth, approximately 3.5 dB noise figure, and compact universal serial bus (USB) form factor. Commercial FM broadcasts occupy 88–108 MHz with a typical inter-station spacing of 200 kHz, meaning the 6 MHz capture window can theoretically accommodate up to thirty stations in a fully occupied band. In practice the number of active stations per 6 MHz window depends strongly on geography; in Edinburgh, where the experiments reported here were conducted, three active broadcast stations were captured simultaneously, and this number could be higher in denser urban environments where FM spectrum occupancy is greater. The received in-phase/quadrature (IQ) stream is processed in GNU Radio, where bandpass filters isolate individual FM channels, each of which is FM-stereo demodulated to recover the CSS-bearing stereo difference signal in the 23–53 kHz subcarrier band.

The per-station baseband outputs are passed to a Python-based CSS decoder that performs correlation-based preamble detection and timing synchronisation^{36,37}, followed by three sequential correction stages that proved critical to achieving low error rates in practice. Carrier frequency offset (CFO) correction compensates the bulk frequency shift between the expected and received chirp spectrum estimated from the preamble. Sampling frequency offset (SFO) correction addresses the fractional clock mismatch between the tag I²S peripheral and the SDR sampling clock, which accumulates progressively over a packet duration. Sampling time offset (STO) correction then refines the symbol timing alignment by estimating the fractional and integer timing offsets from the preamble. The combined application of these three corrections produced a marked reduction in bit error rate relative to uncorrected decoding, directly addressing the elevated BER observed in earlier LoRAB implementations^{27,28} that lacked explicit offset compensation, particularly at higher spreading factors where longer symbol durations render the decoder more sensitive to accumulated timing drift.

Following synchronisation and offset correction, per-symbol fast Fourier transform (FFT) spectra from all admitted stations are combined through signal-to-noise ratio (SNR)-weighted spectral fusion, with cascaded Chase³⁸ and soft-bit decoding fallback for maximum packet recovery. Hard symbol decisions and error-correction decoding are then performed on the fused spectrum. Stations falling below a minimum SNR threshold are excluded from the fusion to avoid noise accumulation from weak branches.

Indoor BER measurements at 10 metres, shown in Figure 3 (left), confirm that each additional station produces a substantial reduction in error rate, with three-station combining suppressing BER by nearly two orders of magnitude relative to single-

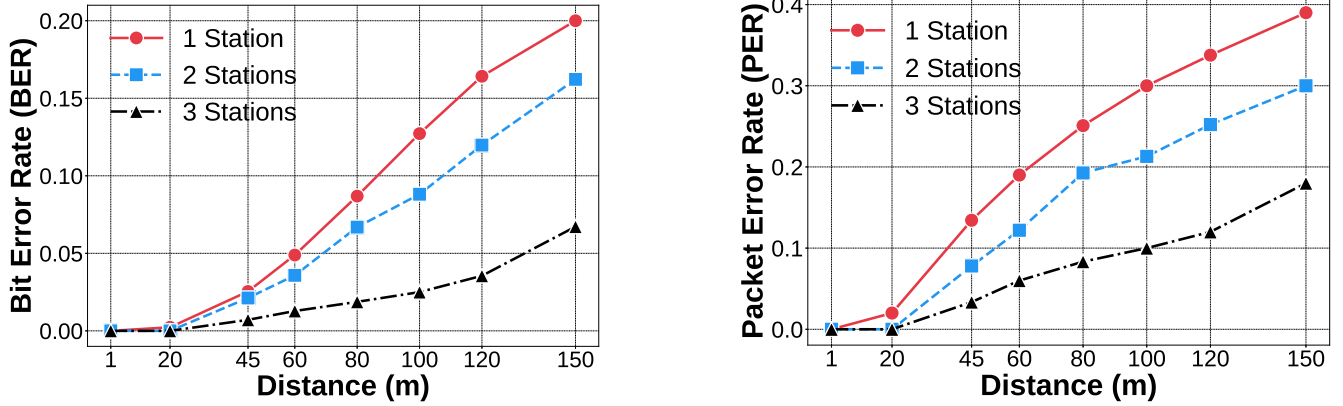


Figure 4. Left: BER vs. Distance. Right: PER vs. Distance.

station reception at SF 7. The corresponding bit rates and current consumption profiles are presented in Figure 3 (centre and right), respectively. Outdoor BER and packet error rate (PER) versus distance results up to 150 metres, shown in Figure 4 (left and right), demonstrate that three-station MISO combining reduces BER to approximately 0.06 and maintains PER below 0.20 at the maximum tested range, satisfying the $\text{BER} < 0.1$ and $\text{PER} < 0.2$ thresholds adopted as figures of merit for successful reception. These measurements used SF 9 packets with a 10 kHz chirp bandwidth and coding rate (CR) 4/8.

A miniaturised self-powered multi-sensor backscatter platform

The LoRAB 2.0 tag is realised on a custom four-layer printed circuit board (PCB) measuring $45 \text{ mm} \times 50 \text{ mm}$, shown in Figure 5, with the PCB reverse side (inset) showing the supercapacitor. The core backscatter functionality, consisting only of the nRF52833³¹ system-on-chip (SoC) and the ADG919³⁰ SPDT switch, could in principle be implemented on a footprint as small as $10 \text{ mm} \times 10 \text{ mm}$. The larger board area is intentional, serving as a research testbed for multiple wireless protocols and sensing modalities simultaneously.

The central processing element is the Nordic Semiconductor nRF52833³¹, a 64 MHz ARM Cortex-M4F SoC with 512 kB flash and 128 kB RAM, chosen for its sub-microampere deep-sleep current, integrated 2.4 GHz Bluetooth Low Energy (BLE) radio, and native near field communication (NFC) interface. Three sensor subsystems are connected via SPI: a Bosch BME280 for pressure, humidity, and temperature at $3.6 \mu\text{A}$ average current, a ST LIS2DH12 three-axis accelerometer consuming as little as $1 \mu\text{A}$ in low-power mode and a ST LIS2MDL three-axis magnetometer drawing approximately $25 \mu\text{A}$ at 10 Hz. The sensors were kept in their power-down state during the power consumption measurements reported here, since the magnetometer alone would draw about $45 \mu\text{W}$ at 1.8 V.

The energy supply is built around a Texas Instruments BQ25504 ultra-low-power boost converter, which harvests energy from an attached solar cell into a 1 F supercapacitor rated to 5.5 V, implementing maximum power point tracking at 80% open-circuit voltage with a cold-start threshold of only 600 mV and a quiescent current below 330 nA. A Torex XC6504 low-dropout regulator (LDO) provides the regulated 1.8 V supply rail, with the supercapacitor voltage monitored via an analog-to-digital converter (ADC) to allow firmware-level duty cycle adaptation based on available energy. The platform also supports BLE, NFC, and additional novel protocols^{39,40} that generate LoRa-compatible transmissions entirely through microcontroller peripheral manipulation without dedicated radio hardware, making the board a consolidated testbed for next-generation battery-free wireless research.

Results

The outdoor experimental setup is illustrated in Figure 6, showing the Airspy Mini SDR receiver and the LoRAB 2.0 tag separated by 150 metres on the Heriot-Watt University campus, with the ambient FM excitation originating from the Black Hill transmitter near Glasgow approximately 34.5 km distant. Three BBC broadcast stations served as ambient carriers throughout all experiments: BBC Radio 3 at 92.1 MHz, BBC Radio Scotland at 94.3 MHz, and BBC Radio 4 at 95.8 MHz. No dedicated or cooperative RF source was used, and the tag relied entirely on existing broadcast infrastructure.

Table 1 compares LoRAB 2.0 with prior ambient FM backscatter systems. The proposed system achieves a communication range of 150 m at a power consumption of $22 \mu\text{W}$, corresponding to a range efficiency of $6.82 \text{ m}/\mu\text{W}$. The original LoRAB²⁷ implementation achieved a range of 130 m, and its listed power of 100 mW is that of the NI myDAQ⁴² instrument used to generate the chirp waveform, while its RF front-end alone drew 130 nA. Its range efficiency of $0.001 \text{ m}/\mu\text{W}$ therefore reflects

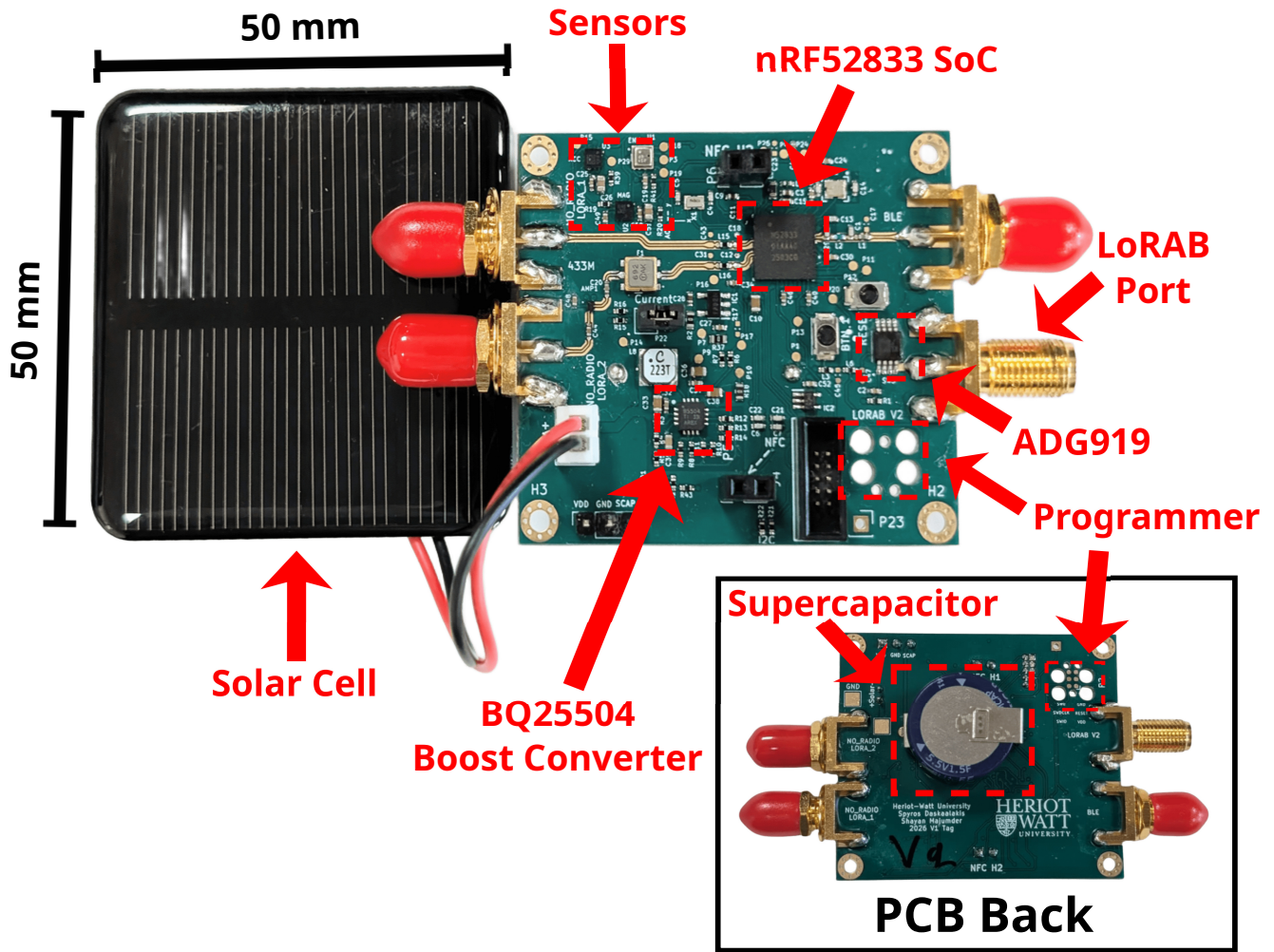


Figure 5. Photograph of the LoRAB 2.0 tag hardware, comprising a custom PCB with an nRF52833³¹ SoC, ADG919³⁰ RF switch, and BQ25504 boost converter, connected to a solar cell for energy harvesting. The PCB reverse side (inset) shows the supercapacitor used for energy storage.

the complete chirp generation chain rather than the switching front-end alone, whereas LoRAB 2.0 generates the chirps on the tag itself within the $22\mu\text{W}$ budget. Furthermore, LoRAB 2.0 is the only system in this comparison that uses a MISO receiver, which provides resilience against the outage of individual FM stations.

Table 2 places LoRAB 2.0 within the broader context of LoRa CSS backscatter systems. It is the only design in this comparison that uses broadcast transmitters as the carrier source, at a source distance of 34.5 km, achieving greater than 100 metres outdoor range at a hardware complexity of just one MCU and one RF switch.

The average power consumption on the custom PCB was measured at approximately $22\mu\text{W}$, marginally higher than the $17.6\mu\text{W}$ ²⁸ reported with the nRF52833 development board in earlier experiments. This discrepancy is attributed to stray leakage currents drawn by the onboard peripherals that share the regulated 1.8 V supply rail with the MCU, including the powered-down sensor subsystems, during deep-sleep intervals between transmissions, and is expected to be reducible through more aggressive peripheral power gating in future firmware and hardware revisions.

Discussion

LoRAB 2.0 demonstrates that the combination of fully digital chirp synthesis, ambient FM excitation, and MISO diversity reception can simultaneously satisfy the power, range, and hardware complexity requirements that have individually constrained prior ambient backscatter systems. The $22\mu\text{W}$ operating power and 150 metre outdoor range achieved here suggest immediate applicability in several domains where maintenance-free wireless sensing is of high value: agricultural and environmental monitoring networks where nodes must survive for years without battery replacement, wearable health and activity monitors

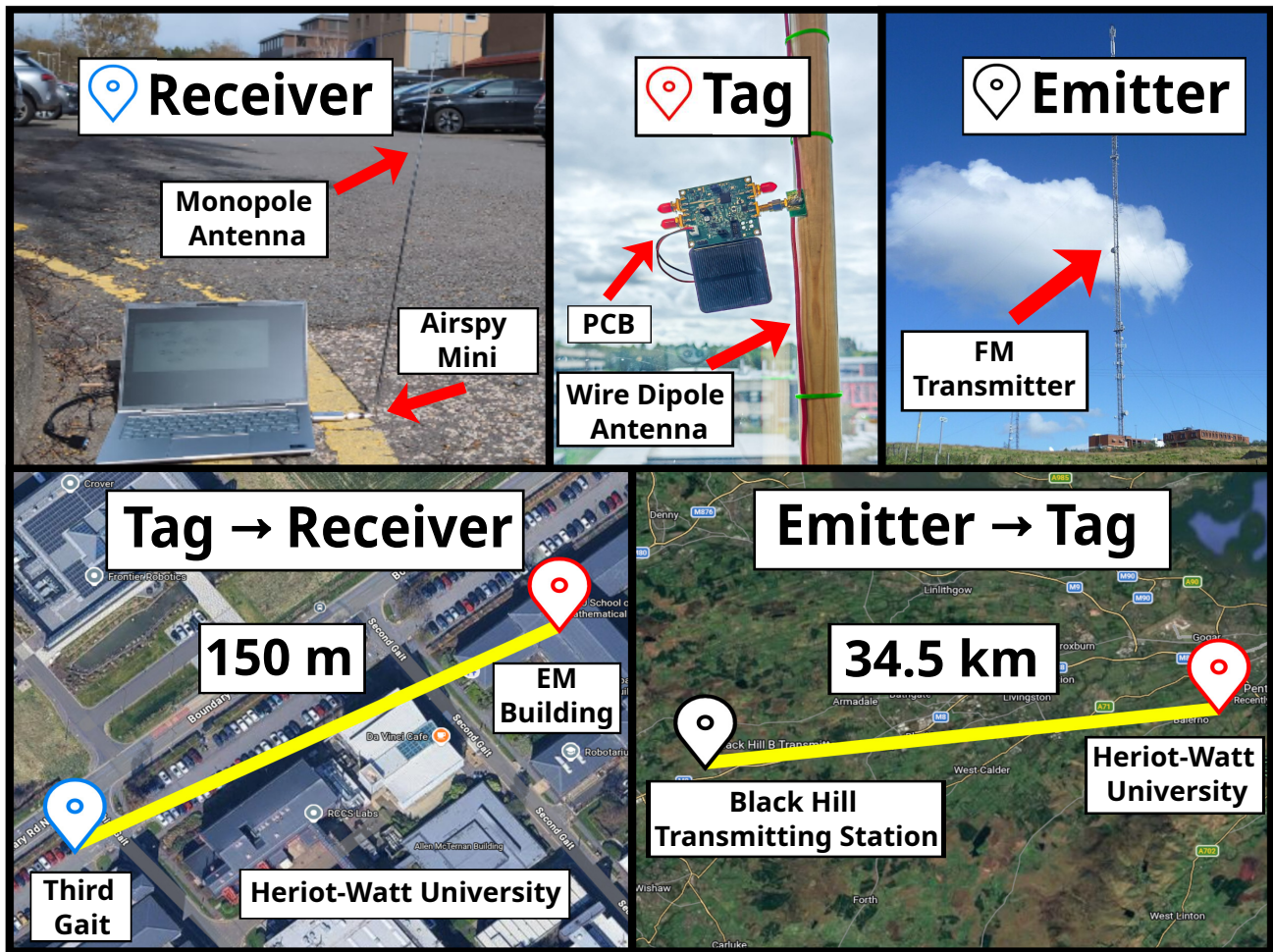


Figure 6. Outdoor experimental setup at Heriot-Watt University, Edinburgh. The receiver comprises an Airspy Mini SDR³² and monopole antenna, the tag consists of the LoRAB 2.0 PCB with a wire dipole antenna, and the emitters are the FM broadcast transmitters located at Black Hill, Glasgow, 34.5 km distant. The tag–receiver separation is 150 m.

where the FM broadcast coverage of urban and suburban environments provides a perpetual ambient carrier, smart building infrastructure where sensor nodes embedded in walls or structural elements cannot be accessed for maintenance, and asset tracking in logistics environments where the density of deployed nodes makes individual battery replacement economically infeasible.

The digital chirp synthesis approach is microcontroller agnostic and can be applied to any microcontroller with a clocked serial peripheral and DMA, as demonstrated by a port of the LoRAB 2.0 tag to the MYOSA board, an alternative low-power microcontroller platform⁴⁶. It is also not inherently limited to ambient FM backscatter and could in principle be applied to LoRa backscatter systems operating against a dedicated single-tone carrier. However, in that context the harmonic content of the binary-approximated chirp waveform, while negligible in the ambient FM case due to the extremely low signal levels involved, could constitute a meaningful risk of spectrum pollution when the incident carrier is strong and the backscattered harmonics fall within occupied adjacent channels. Recent progress in single-sideband (SSB)⁴⁷ backscatter modulation and beamforming-assisted backscatter⁴⁸ systems has opened a promising path toward mitigating this concern, offering the possibility of retaining the hardware simplicity of single-switch digital synthesis while suppressing unwanted spectral products through receiver-side or antenna-side selectivity.

The feasibility of FM backscatter is reinforced by rapid advances in compact, flexible FM-band antenna design. Printed spiral monopoles⁴⁹ occupying as little as 33 mm × 50 mm have been shown to cover the full 88–108 MHz band with omnidirectional radiation, while transparent meandering monopoles⁵⁰ fabricated from micrometric mesh silver film achieve 87% optical transparency with radio performance within 4–5 dB of a conventional quarter-wave antenna. For body-worn

System	Modulation	Range	Power	Range Efficiency [†] (m/ μ W)	MISO Receiver
FM-OOK ²⁴ (2017)	OOK-FM0	5 m	2.84 mW	0.002	×
FM-FSK ²⁵ (2017)	2-FSK	18 m	11 μ W	1.64	×
FM Remodulation ⁴¹ (2017)	Analog FM	26 m	24 μ W	1.08	×
LoRAB ²⁷ (2022)	CSS	130 m	100 mW ^{‡42}	0.001	×
FM Rider ³³ (2024)	2-FSK	107 m	1.4 mW	0.076	×
This Work	CSS	150 m	22 μW	6.82	✓

[†] Range efficiency = $\frac{\text{range (m)}}{\text{power } (\mu\text{W})}$.

[‡]Power of the NI myDAQ instrument used to generate the chirps.

Table 1. Comparison to existing ambient FM backscatter systems.

System	Excitation	Source Distance	Power	Rx Range	Tag Complexity
LoRa Backscatter ³⁴ (2017)	Single-tone	5 m	9.25 μ W [†]	2.8 km	High
LoRea ⁴³ (2017)	Single-tone	1 m	70 μ W	3.4 km	Medium
PLoRa ²¹ (2018)	Ambient LoRa	0.2–10 m	220 μ W [†]	1.1 km	High
XORLoRa ⁴⁴ (2020)	COTS LoRa	1 m	– [§]	500 m	High
P ² LoRa ²² (2021)	Ambient LoRa	1–100 m	320 μ W	2.2 km	High
Home Surveillance ⁴⁵ (2025)	Dedicated LoRa	2.4 m	635 μ W	84 m	Medium
OsTAG ³⁵ (2025)	Single-tone	1 m	7.8–20 μ W [†]	1.2 km	Medium
This Work	Ambient FM	34.5 km	22 μW	150 m	Low

[†]IC estimate. [§]Not mentioned in text. COTS: commercial off-the-shelf.

Table 2. Comparison to existing CSS backscatter systems.

integration specifically, conformal patch antennas on flexible textile substrates have demonstrated around 84% size reduction relative to a conventional half-wavelength patch at 100 MHz, retaining acceptable efficiency even when worn directly on the body⁵¹. As these designs mature, integrating FM backscatter tags into garments, medical patches, and everyday objects becomes increasingly practical, enabling applications that demand not only microwatt-level power but also the mechanical flexibility and unobtrusiveness that rigid PCB-mounted antennas cannot provide.

Future work will pursue two primary directions. The first is high-density ambient backscatter deployment, exploiting curved chirp⁵² waveforms that occupy non-overlapping time-frequency trajectories to allow multiple LoRAB 2.0 nodes to share the same FM carrier infrastructure simultaneously without mutual interference, extending the scalability of the system beyond what is achievable with standard LoRa spreading factor orthogonality alone. The second is range extension through SSB⁴⁷ backscatter modulation, which suppresses one of the two spectral images produced by the impedance-switching process and concentrates the available backscattered power into a single sideband, providing a gain of approximately 10 dB⁴⁷ relative to double-sideband (DSB) operation and potentially extending the reliable communication range by tens of metres beyond the 150 metre result reported here.

Methods

Instrumentation and implementation tools

The two reflection coefficient states Γ_0 and Γ_1 of the ADG919³⁰ SPDT switch were characterised using a Keysight N5234A vector network analyser⁵³ (VNA) covering 10 MHz to 43.5 GHz. Measurements were conducted across the FM broadcast band from 88 to 110 MHz with a 50 Ω reference impedance, yielding the Smith chart plots shown in Figure 2 (centre). The reflection coefficient difference $\Delta\Gamma$ was extracted directly from the calibrated S-parameter measurements at each frequency point across the band.

Power consumption measurements were performed using the Nordic Semiconductor Power Profiler Kit II⁵⁴ (nRF-PPK2), which provides high-resolution current measurements at the regulated 1.8 V supply rail, which feeds the nRF52833³¹ and the onboard peripherals. The sensors were kept in their power-down state throughout these measurements. Average current values were recorded across the full duty-cycle period, encompassing both the active transmission interval and the deep-sleep baseline,

from which the average power consumption of $22 \mu\text{W}$ at 1.8 V was derived.

The custom four-layer PCB was designed using KiCad EDA (version 6.0.4)⁵⁵. The nRF52833³¹ firmware, responsible for I²S-based CSS chirp generation, DMA-driven lookup table transfer, and duty-cycle management, was developed with artificial intelligence (AI)-assisted coding support from Grok (xAI)⁵⁶. The chirp generation approach was inspired by the open-source *LoLRa*⁵⁷ project by C. Lohr, which demonstrates LoRa-compatible waveform generation through microcontroller peripheral manipulation without dedicated RF hardware.

The multi-station GNU Radio processing pipeline and the Python-based CSS decoder, including CFO, SFO, and STO correction stages, as well as the SNR-weighted spectral fusion algorithm, were developed with assistance from Claude Sonnet 4.6 (Anthropic)⁵⁸. All scientific content, experimental results, and intellectual contributions are solely those of the authors. Claude Sonnet 4.6 was additionally used as an AI-assisted copy-editing tool during the preparation of this manuscript.

Digital signal synthesis

The key parameters governing the chirp waveform are the start frequency f_0 , end frequency f_1 , and the spreading factor SF. The symbol duration T is given by

$$T = \frac{2^{\text{SF}}}{\text{BW}} = \frac{2^{\text{SF}}}{f_1 - f_0}, \quad (5)$$

where $\text{BW} = f_1 - f_0$ is the chirp bandwidth. The chirp rate $k = \text{BW}/T$ describes the linear rate of frequency variation over the symbol interval, and the number of samples stored in memory is $N_s = \lfloor TF_s \rfloor$, where F_s denotes the sampling rate imposed by the I²S peripheral bus constraints.

The underlying analog chirp waveform, from which the digital representation is derived, is expressed as

$$s(t) = \sin\left(2\pi\left(f_0t + \frac{k}{2}t^2\right)\right), \quad 0 \leq t \leq T, \quad (6)$$

with instantaneous phase

$$\phi(t) = 2\pi\left(f_0t + \frac{k}{2}t^2\right), \quad (7)$$

and instantaneous frequency

$$f(t) = \frac{1}{2\pi} \frac{d\phi(t)}{dt} = f_0 + kt. \quad (8)$$

Since LoRAB 2.0 operates in the digital domain, the continuous waveform $s(t)$ must be sampled. Letting $t_n = n/F_s$ for $n = 0, \dots, N_s - 1$, the discrete-time sampled waveform is

$$x[n] = s(t_n) = \sin\left(2\pi\left(f_0\frac{n}{F_s} + \frac{k}{2}\left(\frac{n}{F_s}\right)^2\right)\right), \quad n = 0, \dots, N_s - 1. \quad (9)$$

The binary bitstream $b[n]$ transmitted via the I²S peripheral is obtained by thresholding $x[n]$ against the centre baseline:

$$b[n] = \begin{cases} 1, & x[n] > 0, \\ 0, & \text{otherwise,} \end{cases} \quad n = 0, \dots, N_s - 1. \quad (10)$$

This operation maps the continuously varying chirp waveform to a two-level square-wave sequence whose zero-crossing structure faithfully preserves the instantaneous frequency trajectory of the ideal chirp.

Symbol generation for arbitrary payload data exploits the cyclic shift property of LoRa modulation. Two complete up-chirp periods are stored adjacently in RAM, and each payload symbol of value v is rendered by reading from a base address offset by vF_s/B samples, recovering the correctly frequency-shifted chirp through a simple memory-pointer operation without any real-time computation. The DMA employs double buffering, preparing the next symbol buffer while the current symbol is being transmitted, enabling continuous packet output without CPU involvement.

Multi-station signal reception and combining

Within the 6 MHz capture window of the Airspy Mini³² SDR, individual FM channels are isolated in GNU Radio using a Frequency Xlating Finite Impulse Response (FIR) Filter, which simultaneously performs frequency translation and channel filtering. The translated baseband signal for the i -th FM station centred at frequency f_i is

$$y_i(t) = [x(t) \cdot e^{-j2\pi f_i t}] * h(t), \quad (11)$$

where $x(t)$ is the received IQ stream and $h(t)$ is the FIR low-pass filter impulse response. Each isolated channel is then FM-stereo demodulated in GNU Radio to recover the CSS-bearing stereo difference signal residing in the 23–53 kHz subcarrier band above each station's centre frequency.

The per-station baseband outputs are passed to a Python-based CSS decoder. Following^{36,37}, the demodulation of a received LoRa symbol y^l is efficiently performed via a dechirping step followed by a discrete Fourier transform. The symbol estimate $\hat{s}^l$ is obtained as the bin with maximum magnitude:

$$\tilde{\mathbf{Y}}^l = \text{DFT}(\mathbf{y}^l \odot \mathbf{x}_0^*), \quad (12)$$

$$\hat{s}^l = \underset{k \in \{0, \dots, 2^{\text{SF}} - 1\}}{\text{argmax}} \left| \tilde{Y}^l[k] \right|, \quad (13)$$

where $\mathbf{y}^l = [y[lN] \dots y[N + lN - 1]]$ contains the samples of the l -th received symbol, $\mathbf{x}_0 = [x_0[0] \dots x_0[N - 1]]$ is the reference upchirp, $\odot$ denotes element-wise multiplication, and $N = 2^{\text{SF}}$.

Preamble detection proceeds via correlation of the received signal against a reference down-chirp template. Following the approach of^{36,37}, the CFO Δf_c and sampling time offset (STO) τ are each decomposed into integer and fractional components as

$$\Delta f_c = \frac{B}{N}(L_{\text{CFO}} + \lambda_{\text{CFO}}), \quad \tau = \frac{L_{\text{STO}} + \lambda_{\text{STO}}}{B}, \quad (14)$$

with $L_x \in \mathbb{Z}$ and $\lambda_x \in [-0.5, 0.5)$, where B denotes the chirp bandwidth. The fractional CFO $\hat{\lambda}_{\text{CFO}}$ is estimated from the phase rotation between consecutive dechirped preamble symbols³⁷:

$$\hat{\lambda}_{\text{CFO}} = \frac{1}{2\pi} \angle \left(\sum_{l=2}^{N_{\text{up}}} \sum_{p=-2}^2 \tilde{Y}^l[i+p] \cdot \tilde{Y}^{l-1}[i+p]^* \right), \quad (15)$$

where $i = \arg \max_k |\tilde{Y}^l[k]|$ is the peak bin of symbol l and N_{up} is the number of preamble upchirps. This estimator is independent of the integer offsets, since only the fractional CFO contributes to the linear phase rotation between consecutive symbols. The fractional CFO is compensated by multiplying the received signal by a complex exponential:

$$y'[n] = y[n] \cdot e^{-j2\pi \frac{\hat{\lambda}_{\text{CFO}}}{N} \frac{B}{f_s} n}, \quad n \in \mathbb{N}_0. \quad (16)$$

After fractional CFO compensation, the fractional STO $\hat{\lambda}_{\text{STO}}$ is estimated using the rational combination of three spectral lines method³⁷:

$$\hat{\lambda}_{\text{STO}} = \frac{N}{2\pi} \cdot \frac{P[i+1] - P[i-1]}{u(P[i+1] + P[i-1]) + vP[i]}, \quad (17)$$

$$P = \sum_{l=1}^{N_{\text{up}}} \left| \text{DFT}_{2^{\text{SF}}}(\mathbf{y}^l \odot \mathbf{x}_0^*) \right|^2, \quad (18)$$

where $i = \arg \max_k P[k]$, $\text{DFT}_{2^{\text{SF}}}$ denotes a DFT with a zero padding of size 2^{SF} , and the constants are $u = \frac{64N}{\pi^5 + 32\pi}$ and $v = \frac{u\pi^2}{4}$.

The integer CFO and STO are jointly estimated from the upchirp and downchirp preamble symbol indices. The demodulation of the preamble upchirps yields $s_{\uparrow} = (L_{\text{CFO}} + L_{\text{STO}}) \bmod N$, while the downchirps yield $s_{\downarrow} = (L_{\text{CFO}} - L_{\text{STO}}) \bmod N$. The integer parts are then estimated as^{36,37}:

$$\hat{L}_{\text{CFO}} = \frac{1}{2} \Gamma_N[(s_{\uparrow} + s_{\downarrow}) \bmod N], \quad (19)$$

$$\hat{L}_{\text{STO}} = (s_{\uparrow} - \hat{L}_{\text{CFO}}) \bmod N, \quad (20)$$

where $\Gamma_N[k] = k - N \cdot \mathbf{1}[k > N/2]$ maps the raw estimate to the signed integer range.

Following synchronisation and offset correction, each payload symbol is dechirped and transformed via an N -point FFT. The per-symbol power spectrum from each admitted station is accumulated into a shared fusion buffer using SNR-weighted combining:

$$\mathbf{F}[m] = \sum_{i \in \mathcal{A}} w_i \cdot |\text{FFT}(r_i[m] \cdot d^*[m])|^2, \quad (21)$$

where $r_i[m]$ is the m -th symbol segment from station i , $d^*[m]$ is the conjugate reference down-chirp, $w_i = 10^{\text{SNR}_i/10}$ is the linear SNR weight, and $\mathcal{A}$ is the set of stations whose payload SNR exceeds the minimum admission threshold. Hard symbol decisions are taken as

$$\hat{s}[m] = \arg \max_k \mathbf{F}[m, k]. \quad (22)$$

If hard-decision decoding fails, a cascaded Chase³⁸ decoder followed by a soft-bit fallback operate on the normalised fused probability spectrum to maximise packet recovery probability.

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Author contributions

S.N.D. conceived the initial idea. S.M. designed the backscatter system, performed the measurements, interpreted the results, and wrote the initial version of the paper. S.N.D. designed the PCB. S.N.D., M.M.T. and G.G. supervised the research and contributed to the general concept and interpretation of the results. S.N.D. and S.M. revised and edited the final version of the paper. All authors reviewed the manuscript.

Data availability

The datasets generated and analysed during the current study are available from the corresponding author on reasonable request.

Competing interests

The authors declare no competing interests.

Code availability

The multi-station receiver developed in this work, including the GNU Radio flowgraphs and the Python-based CSS fusion decoder, is open source and available at <https://github.com/ShayanMajumder/LoRAB-Receiver> (available upon acceptance of this work).

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