Title

1 – 50-MHz VHF Electromagnetic Sensor-interface Power-attenuation Detector Circuit

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Abstract

The natural but unwelcome byproduct of modern telecommunication systems is electromagnetic interference (EMI). These communication networks are dynamic and produce unpredictable position- and time-varying electromagnetic fields that interfere with sensitive high-performance electronics, for which shielding is often a necessity. The shield's ability to suppress electromagnetic noise, however, may change not only over time but also across environmental conditions. EMI sensors, as a result, play a critical role because arbitrarily over-sizing a shield to accommodate worst-case conditions is not an option in many portable and mobile applications. This paper presents a logarithmically compressed peak-detection EMI sensor-interface circuit that combines the complementary functional strengths of state-of-the-art power detectors to monitor and sense 1 – 50 MHz of EMI with 5-bit accuracy across 16 dB of dynamic range and under –40 to 40 °C. The proposed circuit and printed-circuit-board (PCB) embodiment compensate for temperature variations as well as diode-induced errors to maintain and improve accuracy across a wide operating range.

Key Words

Electromagnetic Interference, EMI, Power Detection, Sensor, Peak Detection, Power Sensor

1. Shielding against Electromagnetic Interference

With telecommunication networks connecting wireless devices around the globe, there is a growing abundance of signals that produce substantial electromagnetic interference (EMI) across the airwaves. These communication networks are ubiquitous and dynamic in nature, producing an array of unpredictable and difficult-to-suppress position- and time-varying electromagnetic fields that hinder today's increasingly sensitive high-performance electronics [1]. Shielding is often an unavoidable necessity, especially in military and biomedical applications but also in the EMI-plagued consumer market where wireless gadgets thrive. A shield, unfortunately, is not always reliable across temperature, operating conditions, or even time, and over-sizing it is prohibitive in many portable and mobile applications. As a result, sensing and measuring the power attenuated across a barrier before, during, and after deployment is necessary to determine if and when extraneous factors compromise EMI shielding integrity. Of course, post-processing the data collected to dynamically reduce or cancel noise is possible, but also more costly and oftentimes unwarranted for applications where the shield is, by design (but not a given), sufficient.

While over-shielding and sensing its effectiveness once (before deployment) is sensible and a popular practice today, a growing number of mobile applications can neither afford the additional burden nor accept the risk of failures in the field. The fact is wear and tear, temperature, humidity, and other factors compromise the integrity of the shield and its seals, which are especially vulnerable to decay. Sensing EMI penetration as the enclosure moves and environmental conditions change *non-intrusively* (which is to say without requiring considerable space or power) is therefore important. Consider, for instance, the army's Standardized Integrated Command Post System Rigid Wall Shelter (SICPS RWS). The army deploys these physical shelters across wide-ranging spaces under extreme conditions to gather and transmit sensitive intelligence in the field. Non-intrusive antenna-based sensors attached to various points and joints across the shielded enclosure, as shown in Figure 1, decrease the risk of failures that could otherwise compromise the shelter's mission.

The sensor periodically measures the power levels at opposite sides of the enclosure with a pair of planar antennae and compares them with the sensor-interface circuit to determine the severity of EMI penetration in decibels (dB) [2]. The sensor must not dissipate much power, if it is to survive a practical deployment life, or require considerable space, considering the enclosure is already full of equipment. It must also be able to detect a wide dynamic range, if it is to characterize the shelter with sufficient accuracy. To understand how to build such a device, Section 2 first introduces and surveys the state of the art in very-high-frequency (VHF) and radio-frequency (RF) power detectors. Section 3 then presents the proposed RF sensor-interface power-attenuation detector system and circuit embodiment and Section 4 illustrates and discusses how the proof-of-concept printed circuit-board (PCB) prototype built with commercial off-the-shelf (COTS) components performed. Section 5 ends the discussion by drawing relevant conclusions.

Note: Lower case variables with upper case subscripts (e.g., v_S) describe all aspects of a signal, upper case variables with upper case subscripts (e.g., V_S) represent only dc components, and lower case variables with lower case subscripts (e.g., v_s) refer only to ac portions.

2. VHF and RF Power Detectors

Sensing EMI strength amounts to detecting ac power, which is why power detectors and EMI sensors appear hand-in-hand. One basic approach to measuring ac power is to relegate the task to physical forces in nature, that is to say, converting the radiated ac power captured by an antenna into a form that is both proportional and more convenient to read, such as temperature or capacitance. The other method is to burden circuits with the task of deciphering (i.e., processing) how much power is present at its input, in other words, squaring and averaging (i.e., root-mean-square RMS) an ac input voltage.

2.1. Thermal Power Detection

Thermal power-detection schemes rely on how a resistor's temperature relates to power consumption. Amplifier A_{AC} in Figure 2a [3]–[4], for example, sources whatever ac current is necessary to reproduce ac input voltage v_{IN} across resistor R_{AC}, forcing R_{AC} to dissipate RMS power as heat in direct proportion to $v_{IN(RMS)}^2/R_{AC}$. A thermocouple (or a diode) [3]–[5] then generates a temperature-dependent voltage $v_{T(AC)}$ that the negative feedback action of amplifier A_{DC}, resistor R_{DC}, and a matched thermocouple emulate with $v_{T(DC)}$ by sourcing dc current into R_{DC}. This way, since $v_{T(DC)}$ equals $v_{T(AC)}$, R_{DC} generates as much heat and therefore consumes as much dc power (V_{OUT}^2/R_{DC}) as R_{AC} dissipates RMS power in heat. As a result, since resistors require 20 ms or more for their temperatures to rise or fall (i.e., thermal time constants are long), which means $v_{T(AC)}$ and $v_{T(DC)}$ are dc signals, output v_{OUT} is a dc voltage (i.e., v_{OUT} equals V_{OUT}) that is directly proportional to $v_{IN(RMS)}$. Generally, while thermal detection schemes of this sort excel in accuracy (of up to 1% with careful resistor matching [4]) and elegance, and integration is possible [6], they suffer from slow response times (because thermal time constants are long) and a limited dynamic range for which the resistors' temperatures correlate linearly with power before succumbing to second-order effects like thermal runaway, when rising temperatures increase resistance, raising temperature further [5].

2.2. Electrostatic Power Detection

Electrostatic schemes use very-high-frequency (VHF) power to force parallel plates to pull together or apart, causing the capacitance between the plates to modulate and therefore convey input power information. The membrane suspended above VHF signal trace v_{IN} shown in Figure 2b [7]–[8], as an example, which is also output v_{OUT}, forms a parallel-plate capacitor with v_{IN}. Because the mechanical time constant of the membrane is substantially longer than that of the VHF signal, its response to v_{IN} is averaged. Moreover, because the force between the plates is proportional to the square of the voltages across its terminals, deflection (i.e., capacitance) reflects averaged VHF power in input trace v_{IN} (i.e., RMS of v_{IN} across the trace's characteristic impedance). In other words, v_{OUT} is not only dc (i.e., an averaged response) but, given charge Q is constant across the open-circuit capacitor (i.e., Q equals CV), also inversely proportional to capacitance and therefore to VHF input power. In the end, unlike its thermal counterpart, the electrostatic approach only requires that input frequencies exceed the mechanical resonance of the membrane, which means the detectable frequency range (i.e., bandwidth) is wide from maybe 100 kHz to 4 GHz [7]–[8]. Additionally, although monolithic integration is difficult, in-package solutions are possible. Nevertheless, the fundamental problem with this technique is capacitance (i.e., dynamic) range is prohibitively narrow for EMI-sensor applications.

2.3. Peak-signal Power Detection

Peak signals may not reflect absolute power, like mean-squared signals can, but they provide a measure of *relative* signal strength. However, while two *uncorrelated* signals with the same peak voltages do not necessarily imply their power levels equal, two *analogous* signals (that is, two signals whose frequency content and corresponding phase match) with the same peak voltages do. Accordingly, peak-signal power detectors compare the peak voltages of two *similar* signals to determine their relative power levels, which in the end, is the objective of the EMI sensor in Figure 1.

To that end, in its simplest form, capacitor C_H in Figure 3a [9]–[11] holds (i.e., stores) the highest (i.e., peak) voltage pn-junction diode D_P or diode-connected MOSFET M_P conducts so that comparing C_H's voltage v_{OUT} against another similarly peak-detected signal establishes the relative signal strengths of the two. A more practical realization, however, replaces the diode with a transistor's base-emitter or gate-source voltage [12], as bipolarjunction transistor (BJT) Q_P depicts in Figure 3b, because the transistor draws supply energy (i.e., collector current i_C from supply V_{CC}) to charge C_H, instead of draining power from v_{IN}. In the circuit shown, to illustrate a more complete example, constant current I_{BIAS} biases Q_P to keep Q_P from shutting off when v_{IN} drops, which would otherwise slow Q_P when attempting to re-engage. V_{CC} and R_{BIAS} (with I_{BIAS}) bias Q_P's base while capacitor C_{AC} couples only the ac portion of v_{IN}. Replica reference BJT Q_{REF} and accompanying bias circuit then sets a reference voltage (i.e., peak signal) against which Q_P's emitter voltage (i.e., the peak voltage of v_{IN}) can be compared. In general, while the fundamental advantage of this signal-rectifying approach is on-chip integration, the basic challenge is limited dynamic range because the voltages D_P, M_P, and Q_P drop (e.g., between 0.5 and 1 V) are a considerable fraction of supply V_{CC}, leaving little margin for v_{IN} to swing and produce linearly detectable signals under moderate to low supply-voltage conditions.

One way of increasing dynamic range is to logarithmically compress input v_{IN} so that a large v_{IN} generates an exponentially smaller counterpart v_{LOG}, where v_{LOG} is proportional to log v_{IN}. In Figure 3c, to cite an example, cascaded and identical linear amplifiers emulate the function of a logarithm, each amplifier A_V clipping (and limiting) its output beyond prescribed input levels. Consider, for instance, that a small v_{IN} keeps all amplifiers in the linear region and their combined effect, as a result, is the output with the largest voltage (i.e., $v_{OUT} \equiv v_{A1} + v_{A2} + v_{A3} \approx v_{A3}$). As v_{IN} surpasses a circuit-defined threshold level, the last amplifier clips its output to a constant V_C (e.g., v_{A3} in Figure 3c: $v_{OUT} \equiv v_{A1} + v_{A2} + v_{A3} \approx V_C$). Increasing Δv_{IN} by a factor of A_V beyond this point causes v_{A2} to also clip, asserting but limiting v_{A2}'s impact on v_{OUT} to V_C (i.e., $v_{OUT} \equiv v_{A1} + v_{A2} + v_{A3} \approx 2V_C$), all of which means large variations in v_{IN} effectively translate to linear V_C changes in v_{OUT}. As before, diodes (D_P) rectify the voltages so that summing transconductor G_Σ can generate a current for holding capacitor C_H to ultimately filter into a voltage. In the end, a dynamic range of 70 dB with a better than ±1 dB log conformance is typical for such a system [3]. Note one additional benefit is that the output is practically already in dB (i.e., in a logarithm). The fundamental drawback here, however, is that the power high-speed linear amplifiers require to process VHF signals is substantially high.

3. Proposed EM Sensor-interface Power-attenuation Detector

3.1. System Design

The system proposed discards thermal and electrostatic schemes in favor of peak-signal detection because EM penetration sensors across a shielded barrier demand both high bandwidth and wide dynamic range [2]. Additionally, the proposed system logarithmically compresses its output to maintain linearity performance across a wide dynamic range and produce a signal that reflects signal attenuation across the enclosure (from Figure 1) in dB. The proposed system is unique and advantageous in that it compresses the already peak-detected signal because, unlike its at-speed incoming ac predecessor in [2], the peak voltage is basically at dc (at low frequencies), so the log amp need not operate at VHF frequencies nor require the considerable power usually attached to such speeds. Additionally, while an external antenna generates a reference ac input v_R, an antenna inside the enclosure generates the sensed attenuated (and correlated) ac voltage v_A against which reference v_R is ultimately compared to gauge EM penetration. Practically, as realized in Figure 3b and now proposed in Figure 4, a parallel replica reference signal path generates the necessary processed peak signal V_{PR} against which v_A's processed peak output V_{PA} is compared.

Low-noise amplifiers (LNAs) $A_{LNA.R}$ and $A_{LNA.A}$ first amplify reference and attenuated ac signals v_R and v_A to increase them to detectable levels and reduce the system's overall sensitivity to noise, that is, decrease the system's noise figure. Because they source energy drawn directly from the supply (not their inputs v_R and v_A), there is no need to replace rectifying diodes $D_{P.R}$ and $D_{P.A}$ with their BJT counterparts, as done in Figure 3b. At this point in the circuit, $D_{P.R}$ and $D_{P.A}$ conduct the peak signals of their respective inputs to low-pass-filter holding capacitors $C_{H.R}$ and $C_{H.A}$ to generate outputs V_{PR} and V_{PA} that represent a measure of v_R and v_A's peak voltages $v_{R(PEAK)}$ and $v_{A(PEAK)}$ (and their relative signal strengths): V_{PR} and V_{PA} are approximately $v_{R(PEAK)}A_{LNA.R}$ and $v_{A(PEAK)}A_{LNA.A}$, respectively. Log amps $A_{LOG.R}$ and $A_{LOG.A}$ then logarithmically compress the resulting dc signals in v_{EBR} and v_{EBA} so difference amplifier A_{DIF}'s output V_O (once tuned) conveys EM penetration P_S/P_A in dB:

$$V_O = v_{EBR} - v_{EBA} = K_C \mathrm{Log}\left(\frac{v_{R(PEAK)}}{v_{A(PEAK)}}\right) \propto \mathrm{Log}\left(\frac{v_{R(PEAK)}}{v_{A(PEAK)}}\right)^2 \propto P_{R(dB)} - P_{A(dB)}, \quad (1)$$

where K_C is a constant gain that A_{DIF} tunes to accurately convey the attenuation factor. Notice that discerning RMS from peak voltages is not necessary because, given analogous v_R and v_A signals, peak and RMS ratios (i.e., attenuation factors) $v_{R(PEAK)}/v_{A(PEAK)}$ and $v_{R(RMS)}/v_{A(RMS)}$ equal, which means V_O indeed represents the degree to which a shielded enclosure suppresses EMI.

3.2. Rectification

Since both v_R and v_A traverse through symmetrical signal-flow paths to A_{DIF}, their respective processing circuits mirror one another. Figure 5 therefore illustrates the portion of the path that leads both v_R and v_A to their respective log amps. As a result, A_{LNA} amplifies representative ac input v_{IN} to drive rectifier D_P-C_H. D_P is a Schottky diode because its voltage is lower than a regular diode's, which means its adverse impact on dynamic range is ultimately lower. Note that, unlike Q_P in Figure 3b, there is no quiescent current flowing through D_P because the diode is sufficiently fast to switch at speed without the aid of additional quiescent power. Unity-gain amplifier A_{BUF} buffers rectified signal v_R to prevent the log amps from loading v_R and otherwise discharging C_H. A_{BUF}, however, presents a finite leakage resistance R_{LEAK} that drains C_H so preventing C_H from drooping below its sensing range places a lower limit on C_H with respect to v_{IN}'s frequency f_{EMI} and R_{LEAK}:

$$C_H \geq \frac{i_{LEAK(MAX)}}{\left(\frac{dv_{C(MAX)}}{dt}\right)} = \frac{v_{R(MAX)}}{R_{LEAK}}\left(\frac{1}{f_{EMI}\Delta v_{R(MAX)}}\right) = \frac{1}{f_{EMI}R_{LEAK}\left(\frac{\Delta v_{R(MAX)}}{v_{R(MAX)}}\right)} \equiv \frac{1}{f_{EMI}R_{LEAK}(0.01)}, \quad (2)$$

where maximum allowable droop voltage $\Delta v_{R(MAX)}$ is half the resolution or 1% of V_{DD} and maximum voltage $v_{R(MAX)}$ is V_{DD}. Because v_{IN}'s frequency varies within the bandwidth of the EMI sensor, the lowest detectable frequency sets the minimum possible value of C_H, which in the foregoing case is 100 pF for 1 MHz and 100 MΩ.

3.3. Logarithmic Compression

Logarithmic compression, as mentioned earlier, amounts to translating linear changes into exponentially smaller variations. Instead of using multiple high-speed amplifiers, which require substantial power, to compress the input signal before filtering it, as previously reported, the proposed system peak-detects and filters the signal before compressing it, using only one lower speed (and lower power) amplifier to emulate the logarithmic function. To be more explicit, because base-emitter voltages (v_{BE}) in BJTs convey exponentially smaller collector currents (i_C), translating low-pass-filtered dc peak voltage V_P in Figure 6 into dc collector current I_C (with the help of op amp A_V via negative feedback) means steady-state changes in i_C manifest as exponentially smaller dc variations in Q_N's v_{BE}:

$$V_{BE} = V_t \ln\left(\frac{I_C}{I_S}\right) = \left(\frac{kT_Q}{q}\right)\ln\left[\frac{V_P - (-V_{DC})}{R_{QN}I_S}\right] = \left(\frac{kT_Q}{q}\right)\ln\left[\frac{(v_{IN(PEAK)}A_{LNA} - V_{DP}) + V_{DC}}{R_{QN}I_S}\right], \quad (3)$$

where V_t and kT_Q/q represent the thermal voltage, k is Boltzmann's constant, T_Q is Q_N's temperature in Kelvin, q is elementary charge, and I_S is Q_N's reverse-saturation current.

Note that the circuit's output is emitter-base voltage v_{EB} (not v_{BE}) and op amp A_V equates (via negative feedback) its terminal input voltages so the difference between V_P and negative diode voltage $-V_{DC}$ (and R_{QN}) set I_C. The reason for adding compensating Schottky diode D_C's voltage V_{DC} is to cancel the offset voltage rectifier diode D_P in Figure 5 introduces in V_P (as V_{DP}), which would otherwise introduce a non-linear error into the system. As a result, while $v_{R(PEAK)}$ and $v_{A(PEAK)}$ (and their corresponding V_{PR} and V_{PA}) in Figure 4 may differ by orders of magnitude, logarithmically compressed V_{EBR} and V_{EBA} are comparable and therefore easier for A_{DIF} to process linearly.

3.4. Difference Amplifier

Op amp A_V in Figure 7 subtracts logarithmically compressed reference and attenuated dc voltages V_{EBR} and V_{EBA} to produce in V_O the logarithm of their ratio:

$$V_O = \frac{R_{DF}}{R_D}\left(V_{EBA} - V_{EBR}\right) = \frac{R_{DF}}{R_D}\left[V_t \ln\left(\frac{V_{BER}}{V_{BEA}}\right)\right] = \frac{R_{DF}}{R_D}\left[V_t \ln\left(\frac{V_{PR} + V_{DCR}}{V_{PA} + V_{DCA}}\right)\right], \quad (4)$$

where two LNAs amplify reference and attenuated inputs v_R and v_A by A_{LNA} before their respective rectifiers introduce offset diode voltages V_{DPR} and V_{DPA} in peak voltages V_{PR} and V_{PA}, expanding and approximating V_O to

$$V_O = \left(\frac{R_{DF}}{R_D}\right)\left(\frac{kT_Q}{q}\right)\ln\left[\frac{v_{R(PEAK)}A_{LNA} - V_{DPR} + V_{DCR}}{v_{A(PEAK)}A_{LNA} - V_{DPA} + V_{DCA}}\right] \approx \left(\frac{R_{DF}}{R_D}\right)\left(\frac{kT_Q}{q}\right)\ln\left(\frac{v_{R(PEAK)}}{v_{A(PEAK)}}\right). \quad (5)$$

Because the differences in diode voltages are considerably smaller than the amplified peak voltages, V_{DPR}, V_{DCR}, V_{DPA}, and V_{DCA} disappear and A_{LNA} cancels. At this point, extracting the log relationship from the ln function and tuning R_{DF}/R_D accordingly sets gain proportionality A_{EMI} to one to produce an output that is equivalent to the power-attenuation factor between ac reference v_R and its attenuated counterpart v_A:

$$V_O = \left(\frac{R_{DF}}{R_D}\right)\left(\frac{kT_Q}{q}\right)\left(\frac{\ln 10}{20}\right)\left\{10\log\left[\left(\frac{v_{R(PEAK)}}{v_{A(PEAK)}}\right)^2\right]\right\} = A_{EMI}\left(\frac{P_R}{P_A}\right)_{(dB)} \equiv P_{R(dB)} - P_{A(dB)}. \quad (6)$$

Notice that gain factor A_{EMI}, which should remain at one, is proportional to absolute temperature (PTAT) T_Q. To counter this temperature drift in V_O, the type of resistor used for R_D is selected so that its resistance also increases linearly with temperature (with respect to feedback resistor R_{DF}), that is, so that R_D/R_{DF} also increases with T_Q, effectively canceling the effect V_t's T_Q has on A_{EMI}:

$$A_{EMI} = \left(\frac{R_{DF}}{R_D}\right)\left(\frac{kT_Q}{q}\right)\left(\frac{\ln 10}{20}\right) \propto \frac{T_Q}{T_Q} \neq f(T_Q). \quad (7)$$

This cancellation, which amounts to complementary-to-absolute-temperature (CTAT) compensation, is imperfect because finding two resistors whose ratio increases linearly with temperature linearly is, in practice, difficult. Nevertheless, choosing a resistor type for R_D that increases with T_Q at a faster rate than another for R_{DF} mitigates the impact temperature has on the system. In the prototyped PCB, R_{DF} was a cermet resistor with a temperature drift of +2 Ω/°C and R_D was a combination of ceramic and silicon resistors, the latter of which had a strong PTAT behavior at +7.6 Ω/°C. The reason for splitting R_D in two is because the drift of the resistor found was excessively PTAT so adding another lower drift resistor decreases the overall temperature coefficient.

4. Experimental Measurement Results

The photograph in Figure 8 shows the prototyped printed-circuit board (PCB) of the proposed EM sensor-interface power-attenuating detector in Figure 4 and comprised of the rectifier, logarithmic compression, and differential-amplifier sub-systems in Figures 5 – 7. All components in the prototyped proof-of-concept PCB, as specified and desired, are commercial off-the-shelf (COTS) parts. The tunable portion of the gain-adjusting feedback resistor R_F in difference amplifier A_{DIF} is a 100-turn 10-kΩ potentiometer. To test the system, a very-high-frequency (VHF) function generator emulated the shielded barrier's correlated reference and attenuated antennae ac signals v_R and v_A at various frequencies and an oven set the temperatures at which measurements were to be taken to ascertain accuracy performance. Using actual antennae and their corresponding matching networks to produce v_R and v_A, although of value for the finished product, is not necessary in the proof-of-concept phase, as long as emulated v_R and v_A signals carry frequencies of interest at the desired power levels. For demonstration purposes, the five least significant bits (i.e., 5 LSBs) of the analog/digital (A/D) converter's eight-bit output (i.e., $D_1 – D_5$) feed into an array of light-emitting diodes (LEDs) to visually display the digital bit stream as a sequence of lights. For accuracy measurements, to determine the practical performance limits of the system, the set-up monitored A_{DIF}'s analog output v_O directly.

4.1. Frequency Response

Figure 9 illustrates the frequency response of the system when subjected to signals attenuated by 5, 10, and 15 dB. The system's minimum –3 dB bandwidth $f_{–3dB}$ is approximately 50 MHz. Note that after $A_{LNA.R}$ and $A_{LNA.A}$ and $D_{P.R}$-$C_{H.R}$ and $D_{P.A}$-$C_{H.A}$ amplify, rectify, and low-pass-filter reference and attenuated ac signals v_R and v_A, log amplifiers $A_{LOG.R}$ and $A_{LOG.A}$ and differential amplifier A_{DIF} process *low-speed* signals so the limiting factors in frequency response are the LNAs and rectifiers, which operate at speed, at f_{EMI}. In the prototyped case, the LNAs' bandwidths limited the system's response to below roughly 60 MHz. The rectifying diodes' transitional frequencies and related non-linear junction and parasitic capacitances further reduced the system's bandwidth to the measured 50 MHz. This latter effect is not a surprise because the diodes must shut off completely and fully engage at speed, which quiescent current would improve but only at the expense of additional power.

4.2. Accuracy

Figure 10 illustrates the accuracy of the prototyped system with a reference source power P_R of 8 dBm for attenuated signals (v_A) whose power levels are up to 20 dB below P_R. As shown, the system achieved five bits or ±0.5 dB of equivalent accuracy across a dynamic range of 16 dB. Mismatches between the reference and attenuated signal paths limit accuracy to this level. The two logarithmic compressing transistors (i.e., Q_N in Figure 6), for example, introduce an offset, but since they are both on the same chip (in DMMT5551), their currents match to within 2%, which means they do not account for the inaccuracy measured.

Although A_{BUF} in the rectifiers, logarithmic-compression circuits, and difference amplifier in Figures 5, 6, and 7 also introduce uncorrelated input-referred offsets, each amplifier used (i.e., OPA2277 and OPA2266) has less than 10 μV of offset, so their impact on mismatch error is considerably below 1%, which is negligible. The front-end low-noise amplifiers (A_{LNA} in Figure 5) also introduce an offset, but again, not enough to account for the error seen.

The main culprits here are compensating diodes $D_{C.R}$ and $D_{C.A}$ (i.e., D_C in Figure 6) because, while $D_{C.R}$ and $D_{C.A}$ carry constant bias currents, rectifying diodes $D_{P.R}$ and $D_{P.A}$ do not, which means compensating diode voltages V_{DCR} and V_{DCA} do not exactly offset rectifying diode drops V_{DPR} and V_{DPA}, as they meant to (in Equation 5). Accuracy is therefore worse at higher attenuation levels because the diode voltage-difference error is greatest when the reference and attenuated signal strengths differ most, that is, when the worst disparity between V_{DPA} and V_{DCA} exists (in response to low $v_{A(PEAK)}$ voltages). As a result, accuracy and dynamic range are generally design tradeoffs in the system, except when the difference between V_{DPR} and V_{DCR} also dominates, which may cause irregularities in the response.

4.3. Temperature Drift

Figure 11 illustrates the temperature drift of the system's normalized output from –40 to 80 °C when subjected to 1- and 10-MHz input signals and attenuated by 10 dB. Experimental results show the system is within 0.03 dB of its target up to 40 °C, which proves the efficacy of R_D's PTAT-compensating resistance in canceling the rise thermal voltage V_t would have otherwise induced in v_O with increasing temperatures. The reason why accuracy degrades past 40 °C is more than likely the tuning potentiometer in R_{DF}. The thermal coefficient of expansion of R_{DF}'s plastic package is probably sufficiently high to stress R_{DF} and cause (via piezoelectric effects) its temperature coefficient to shift. In other words, R_{DF}'s temperature dependence changes past 40 °C and causes gain-setting resistor ratio R_{DF}/R_D in difference amplifier A_{DIF} to also deviate from its tuned point.

4.4. Discussion

In comparing the system proposed with the state of the art, it is important to recognize that the EM power-detection functionality required and achieved is unique and not previously reported in literature. Within the context of power detection and logarithmic compression, however, comparing the corresponding portion of the foregoing system with the state of the art is a reasonable exercise. With this in mind, while the logarithmically compressed power-detection scheme proposed did not outperform the previously reported embodiment in resolution or dynamic range (±0.5 versus ±1 dB of resolution with 16 versus 70 dB of dynamic range [3]), the proposed system considerably reduces complexity and power consumption. The fact that only one *low-frequency* amplifier-BJT combination achieved the functionality of *several high-speed* amplifiers carries considerable weight in PCB real estate (i.e., foot-print dimensions), cost (for parts), and power, the latter of which is critical in mobile battery-operated devices, such as in wireless microsensors, because power determines operational battery life. To be more specific, literature shows that cascading high-speed amplifiers, for example, consume 0.75, 1.06, and 5.2 W [13], whereas the low-frequency amplifier-BJT combination can dissipate only 7.5 mW for a bandwidth of 110 MHz [14]. What is more, the constituent components of the system proposed are compatible with system-on-chip (SoC) silicon-based technologies, which means on-chip integration and more power (and cost) reductions are possible.

5. Conclusions

The proposed (and prototyped) proof-of-concept electromagnetic (EM) sensor-interface power-attenuating detector system measures the power attenuation between two 1 – 50-MHz signals within five bits of accuracy for up to 16 dB of attenuation (i.e., dynamic range) across a temperature range of –40 to 40 °C. The system constrains the burden of high-frequency operation, which demands considerable power, to the front end only, logarithmically compressing *low-frequency* signals to (*i*) increase dynamic range and (*ii*) generate an output that is convenient to both read and process: in dB. The circuit compensates not only the error that the rectifying diodes introduce by subtracting similarly defined voltages later in the processing chain but also the temperature drift the log amp induces in its output by carefully introducing a complement-to-absolute-temperature (CTAT) term in its response. Ultimately, EM sensors of this sort play an important role in maintaining the integrity of state-of-the-art applications that are sensitive to very-high-frequency (VHF) and radio-frequency (RF) noise by monitoring and measuring the extent to which EM interference (EMI) penetrates their protective shields, which is critical in military and biomedical applications as well as consumer markets whose insatiable appetite for wireless gadgets seems to border on the extreme. Additionally, by steering away from thermal and electrostatic sensors and relying on peak-signal-processing techniques, robust in-package and even on-chip integration of most of the EM power-attenuating detector is possible, potentially reducing the footprint to micro-scale levels.

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Vitae

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