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Zero-Power Radio Device

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Zero-Power Radio Device

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Abstract

This report describes an unpowered radio receiver capable of detecting and responding to weak signals transmitted from comparatively long distances. This radio receiver offers key advantages over a short range zero-power radio receiver previously described in SAND2004-4610, *A Zero-Power Radio Receiver*. The device described here can be fabricated as an integrated circuit for use in portable wireless devices, as a wake-up circuit, or as a stand-alone receiver operating in conjunction with identification decoders or other electronics. It builds on key sub-components developed at Sandia National Laboratories over many years. It uses surface acoustic wave (SAW) filter technology. It uses custom component design to enable the efficient use of small aperture antennas. This device uses a key component, the pyroelectric demodulator, covered by Sandia owned U.S. Patent 7397301, *Pyroelectric Demodulating Detector* [1]. This device is also described in Sandia owned U.S. Patent 97266446, *Zero Power Receiver* [2].

CONTENTS

NOMENCLATURE

AM	amplitude modulated signal
BPSK	bi-phase shift keying, a type of signal modulation
CDMA	code division multiple access
COTS	commercial off the shelf components
dB	decibel
DC	direct current
FCC	federal communication commission
FET	field effect transistor
HF	high frequency, 3-30MHz
HP	Hewlett-Packard
IDT	interdigitated transducer
IF	intermediate frequency
ISM	instrumentation, scientific, and medical, an FCC unlicensed frequency band
JFET	junction field effect transistor
$LiNbO_3$	lithium niobate
MEMS	micro-electrical mechanical systems
MOSFET	metal oxide semiconductor field effect transistor
NiCr	nickel chromium, or nichrome
PZT	lead zirconate tantalate
RAM	random access memory
RF	radio frequency
RFID	radio frequency identification
RMS	root-mean-square
SAW	surface acoustic wave
TaN	tantalum nitride
VHF	very high frequency, 30-300MHz

1. INTRODUCTION

The idea of a zero-power radio receiver is not new. The crystal-set radio, an entirely unpowered form of radio receiver, was invented in 1906. There are several different types of commercially used wireless radio frequency identification (RFID) tags that are essentially zero-power radios. However, most unpowered wireless RFID tags only have a range of a few meters. There are long-range, low powered tags that draw energy either from a battery or some other form of localized power. This report describes a new form of zero-power radio that operates at long ranges and that does not need any form of localized power. For this discussion, a passive, or zero-power, radio receiver is defined as a receiver that uses no direct electrical power but makes sole use of the power available from a transmitter via its transmitted radio frequency signal. Any divergence from this definition will be clearly stated. There may in some instances be distinct advantages to using low power circuitry in some portion of the receiver, but that use should be clearly understood, as there are always trade-off costs associated with using localized electrical power. The invention described in this report is a true zero-power radio receiver, and no portion of the receiver requires powered circuitry.

Commercial applications for short-range zero-power radio receivers are already common for RFID tags used in applications ranging from low-cost theft prevention devices in stores and libraries to somewhat more expensive devices used to track shipping containers and pallets in warehouses. A long range, zero-power radio receiver has hitherto been unavailable. However, it is not difficult to imagine how such a device might be of value in the commercial world. Right now, cell phones must continually turn their radio receiver circuitry on and off to listen for attempts to be contacted by the cellular base station. Similarly, mobile Internet-of-things devices must cycle their receivers on and quickly off to keep track of changes in location. A long range, zero-power radio receiver with sufficient sensitivity can greatly extend battery life in wireless mobile applications. It can also enable wireless communications in applications that are currently impractical, such as embedding wireless receivers in the walls of new houses or the spinning components of an engine.

A miniature, zero-power radio receiver might even be implantable in the body, communicating directly from its own low voltage output to stimulate inherently low voltage neurons. This could facilitate many areas of current biomedical research and enable a huge number of new applications. There many areas where animal researchers, medical researchers, and doctors are implanting cumbersome, battery-powered radio receivers. These include radio-controlled insects [3], vision replacement communications for the human brain [4], and restoration of control for spinal cord injuries [5]. All of these applications, and many more, could be greatly aided by a miniature, permanent radio receiver with a very small antenna that never needs to have a battery changed out. Applications presently confined to science fiction could be enabled, such as direct nervous system reception of transmitted audio or video signals or direct reception of magnetic, voltage, or other sensor types located outside the body.

A previous report, SAND2004-4610, *A Zero-Power Radio Receiver* [6], described both a general methodology and some specific examples of passive radio receivers. These radio receivers were not capable of receiving weak signals from distant sources. The zero-power radio receivers previously described in [6] made use of commercial-off-the-shelf (COTS) components along with a custom surface acoustic wave (SAW) correlator, to enable relatively short range unpowered communications. The general methodology described in [6] offers some advantages over commercially used unpowered wireless tag techniques, such as those described in Finkenzeller's *RFID Handbook* [7]. Commercial RFID approaches often use a low frequency, inductively coupled signal to power a higher frequency communication receiver. Inductively coupled RFID tags typically have a range of 1-2 meters. There are also commercial RFID tags in use and described in [7], operating in the 434 MHz or 915 MHz ISM bands, that use voltage multipliers to enable operation at up to 10 meters. The zero-power receivers described in SAND2004-4610 have a range of tens to hundreds of meters. As will be shown in this report, the new zero-power receivers potentially have much longer ranges.

Any zero-power radio receiver depends on four key attributes. This is the case whether the zero-power receiver is a 1950's era crystal-set receiver, an unpowered wireless ID tag, or a long-range zero-power wake-up receiver. The first attribute that any zero-power receiver must possess is selective reception of the correct radio signal from a radio frequency environment filled with background noise. The second attribute is isolation of that signal from as much of the received noise as possible. Understanding the third attribute entails first understanding radio waves.

Transmission of radio frequency signals typically occurs in free space or in some facsimile to free space such as air. Electromagnetic waves which travel in free space are generally time varying oscillations transferring energy back and forth between the electric and magnetic field components. The impedance of free space, 377Ω, creates a balance between electric and magnetic field components that keeps the peak power in each component equal. To interface to other electronics or biological systems, the electric field component in that material must be made to be much greater than the temperature induced agitation of localized charges measured in Volts and known as the thermal voltage, V_{th}. The thermal voltage is the average voltage noise induced by thermally excited charges in a material. The thermal voltage in a semiconductor is given by

$$V_{th} = kT/q = 26\text{mV (at room temperature)} \qquad (1)$$

This voltage, though small, is much higher than that produced by any but the highest power received radio signals across a typical receiving antenna. For example, a typical received terrestrial communication signal of -90dBm, or 1 picoWatt, produces 27μV peak across the free space impedance of 377Ω. It takes almost 1μW, or -30dBm, six orders of magnitude more power, to produce a peak voltage equal to V_{th} across the free space impedance. Across a dipole antenna with an impedance of 73Ω, it takes almost -23dBm of received signal power to produce a peak signal voltage equal to the thermal voltage. These concepts point to the third attribute which a zero-power radio receiver must possess. That is, it must present a transformation from the receiving impedance environment of free space to a much higher impedance environment. A received signal of -90dBm, which produces 27μV at the free space impedance of 377 Ω, instead produces a peak voltage of 1.4V across an impedance of 10^{12}Ω. A peak voltage of at least $4V_{th}$ (= 100mV) is needed for reliable circuit operation with a low probability of false alarms. The third attribute then is that the received signal must undergo an impedance transformation from free space to a much higher impedance.

The fourth attribute which a zero-power receiver must possess is the ability to transform the high frequency oscillating received signal into a DC or very low frequency, baseband signal. That is, it must be capable of usefully applying the final received signal. This application process consists of transforming the received RF signal into an appropriate signal type that the final end-user of the signal needs. For a crystal-set radio, the final signal is an audio signal across a high impedance earphone with enough signal power to be intelligible against the background received noise or spurious signals. For a commercial RFID device, the final signal is a coded stream of bits that have been stripped from an RF carrier signal which has coincidentally provided enough power to enable the voltage multipliers on the RFID to power the chip. For a long distance zero-power wake-up receiver, it can be a DC voltage exceeding the gate threshold of an FET transistor. This FET transistor can then be used to power-on an identification decoder, store memory bits in a static RAM, turn on a powered transponder, etc.

2. BASIC ELEMENTS

There are four basic elements of a long-range, zero-power radio receiver, corresponding to the four attributes previously described. These four elements are the antenna, the narrowband filter, the impedance transformer/demodulator, and the detector (figure 1). It is not essential that a zero-power radio receiver have all four of these elements or have an optimized version of all four of these elements, but if any element is lacking or is not optimal, then the receiver will have reduced range and sensitivity.

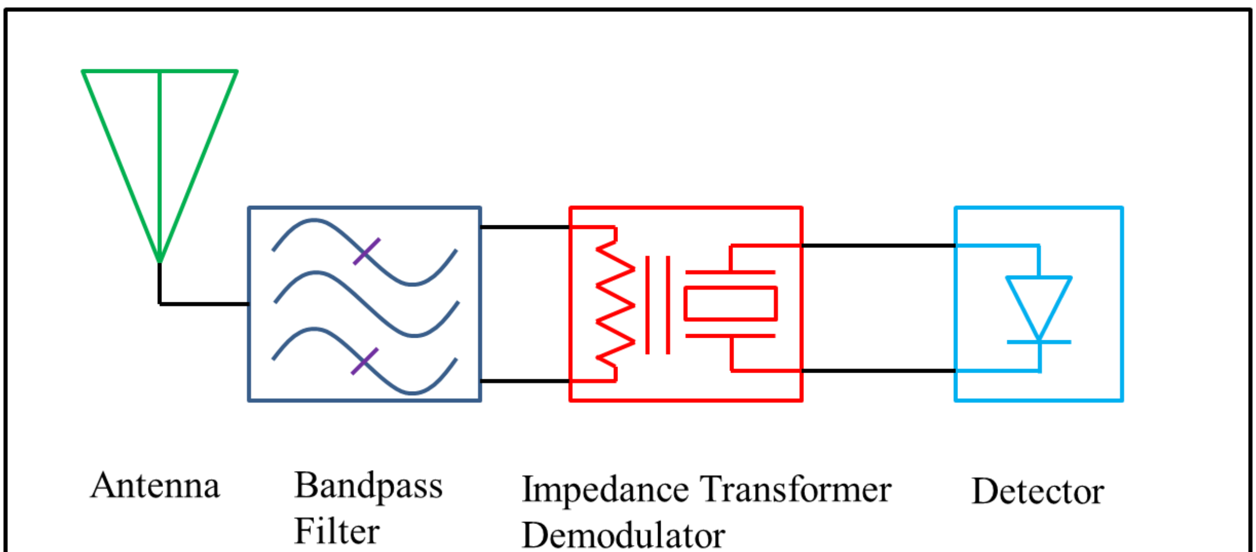

Figure 1: Zero-power radio receiver - Basic elements

It is worthwhile to consider each of the elements in the radio receiver and the role that each serves in signal reception. Each of the basic elements must not only be optimal for the signal handling that it is required to perform, but it must also interact optimally with any components upstream or downstream from it. Since a true zero-power radio works only with the received RF signal power, each component cannot increase that signal power, it can only diminish it. The decrease in signal power through each element must therefore be kept to an absolute minimum. In conventional radio frequency (RF) engineering terms, this means that the insertion loss through each component must be kept to a minimum, that impedances must match at all element interfaces, and that noise sources must be carefully excluded from the signal path. This also means that it is not possible to obtain power gain in the signal path, since no localized source of power is used in a true zero-power receiver. It is possible to obtain voltage gain, process gain, and antenna gain, however. Understanding how these three gains contribute to signal output is critical to understanding a zero-power receiver.

It is also important to consider that a zero-power receiver might use multiple versions of each of the basic elements. For instance, the antenna/ bandpass filter might be a multiple-input-multiple-output (MIMO) version to receive simultaneously on different frequencies. An antenna optimized for a certain frequency band might be paired with a matching bandpass filter. Then multiple versions of these antenna/ bandpass filter pairs can be interconnected into a single demodulator and output detector. This is an example of modification of the simple circuit shown in figure 1 using parallelism.

It is also possible to make serial modifications to the simple zero-power receiver depicted in figure 1. For example, a bandpass filter coupled with a matching demodulator can be considered to form a single stage and this can be connected serially in two or more such stages to downconvert – without the use of a mixer – the incoming signal. This can effectively form a superheterodyne receiver by downconverting a modulated signal to one or more intermediate frequencies (IF) before being finally converted to a baseband signal. An implementation of this concept using bi-phase shift keying (BPSK) modulation with SAW correlators is discussed in some detail in [6].

Using the pyroelectric demodulator lends itself to the downconversion of amplitude modulated (AM) signals to intermediate frequencies through successive stages of downconversion, as in a conventional superheterodyne receiver. This approach is similar to the approach described in [6], except that the SAW correlator is replaced by a pyroelectric demodulator. Also, the BPSK modulated signal is replaced with an AM modulated signal. In this instance, the resulting radio has the same narrow reception bandwidth as a conventional AM superheterodyne receiver.

3. Antenna

The antenna forms a critical part of any radio receiver or transmitter. It is particularly important with a miniature radio receiver, since the antenna can be the largest part of the system. Antenna dimensions typically scale in direct proportion to the frequency of operation and the gain or directivity of the antenna. Consequently, many modern radio receivers are pushed to operate at higher frequencies in order to keep the antenna size small. As will be demonstrated in the section on Example Designs, this poses a problem for zero-power, low power, or RFID receivers, since the range over which such a device will work is directly proportional to the wavelength of the RF signal used. Lower frequency transmitters can be heard at greater distances than higher frequency transmitters, if antenna gains and all other parameters are kept equal. There is an additional effect that enables HF and VHF communications to be heard at very great distances. The atmospheric "E" layer can enable a waveguiding effect between earth and sky that can allow signals in these frequency bands to travel many thousands of miles or even all the way around the earth. The disadvantage of radio transmission at low frequencies is that dimensions for wavelength proportional antennas scale inversely with the frequency of operation and can become quite large. For example, a $\lambda/2$ dipole at the commercially used transmission frequency of 1.62GHz has a gain of 2.15dBi and a size of 9.3cm (3.6in). Similarly, a $\lambda/2$ dipole at the 40.68MHz ISM band frequency also has a gain of 2.15dBi but with a size of 3.7m (12ft). Obviously, it doesn't make much sense to build a very small, portable radio receiver with a 12ft antenna. However, a pyroelectric demodulator-based zero-power receiver has an advantage that enables operation with small aperture antennas.

Very small dipole antennas with lengths on the order of 1/50th of a wavelength can be made to have an effective aperture that is similar or even identical to that of a half-wave dipole antenna [8]. This can be accomplished if the impedance matching and radiative efficiencies of the two antennas are the same. The effective aperture of an antenna is a good measure of that antenna's power collecting ability. This is because the power delivered to the input of a radio receiver is the incident power density multiplied by the antenna's effective aperture. The maximum effective aperture of an antenna is given by [8, page 94]:

$$A_{em} = e_{cd}(1 - |\Gamma|^2)\left(\frac{\lambda^2}{4\pi}\right) D_o |\rho_\omega * \rho_a|^2 \tag{2}$$

Here e_{cd} is the radiation efficiency, the ratio of the radiated power to the sum of the radiated and loss power of the antenna, Γ is the reflection coefficient of the antenna to the input of the receiver, λ is the wavelength, D_o is the antenna directivity, ρ_ω is the polarization of the incoming wave, and ρ_a is the polarization of the antenna. As is shown in the section on Example Designs, it is possible to create a small aperture antenna that has an effective area comparable to that of a half-wave dipole. This can be used in a zero-power receiver either to greatly reduce the size of the antenna at a given frequency or to greatly extend the range of reception by lowering the operating frequency for a fixed antenna footprint.

The first difficulty is that radiation efficiency for a very small dipole antenna declines somewhat below 100%. Radiation efficiency of an antenna is the ratio of radiation resistance to radiation resistance plus parasitic resistance. It is essentially a measure of power transmitted from an antenna to total power accepted by the antenna. Even with the small antenna resistive losses, radiation efficiency can be kept to no less than 70-95%, and this small impact is relatively insignificant. The second difficulty which has hitherto precluded the use of very small aperture antennas in most radio receiver applications, including RFIDs and zero-power receivers, is that the input stage of the receiver must be impedance matched to the very low impedance of the small aperture antenna. However, in the design of a zero-power receiver, it is possible to design the input impedance of the receiver in such a manner as to be a good match to a very small aperture antenna. This can be accomplished by using a pyroelectric demodulator as the input component to the receiver. The input resistor used in the pyroelectric demodulator can be chosen to have a resistance that is equal to the input resistance of the small aperture antenna. A second method to provide a match to a small aperture antenna is to use a specially designed SAW filter as an interface element. The custom SAW filter can have its input interdigitated transducer (IDT) designed to have a very low input resistance to match to the antenna with few or no additional matching circuitry. This can be done by designing the SAW IDT to make use of parallel driver elements so that the input impedance of the filter is dominated by a low resistance element with little parasitic capacitance or inductance. The effect is to enable a small aperture antenna to be integrated with a small filter (figure 2) and demodulator (figure 5).

4. Bandpass Filter

At least one narrowband filter should be used within the detection chain of the zero-power receiver. Its purpose is to limit the input reception frequency band to that of the signal of interest. The bandpass filter used for a zero-power receiver must possess the highest possible performance in terms of insertion loss and passband bandwidth. The passband bandwidth is normally expressed in terms of Q or the ratio of total bandwidth covered divided by bandwidth passed. The bandpass filter in a very low performance zero-power receiver such as a crystal-set radio is formed from the tank circuit created by the tunable inductor/ capacitor combination connected to the antenna (figure 3).

Conventional multi-element or resonator types of filters used in microwave signal processing do not possess the necessary circuit Q to provide adequate performance for a long range zero-power receiver. Ideally, the bandwidth of the bandpass filter should be just wide enough to accept the expected modulation of the input signal. This is because any additional bandwidth will admit noise which will obscure detection of the signal. A typical, high performance RF or microwave filter Q is less than 100. This means that the passed signal of the bandpass filter is 1/100th of the center frequency of the filter. For a filter with a center frequency of 1GHz and a Q of 100, the passband will be 10MHz wide. Typically, both much higher and much lower frequencies are blocked by additional filtering elements. A 10MHz passband filter connected to an isotropic antenna with unity gain will see kTB (= 41fW) of thermal noise alone. This may not sound like a lot of power, but it has a peak voltage of around 900mV across a $10^{13}\Omega$ MOSFET gate. The Q of the narrowband filter required for the receiver of a long range zero-power receiver must be at least 100,000 when operating at high RF or microwave frequencies. A filter with a center frequency at 1GHz and a Q of 100,000 will have a bandwidth of 10 kHz. This filter will admit 41aW of thermal noise when connected to an isotropic antenna, and this will produce 29mV peak across a $10^{13}\Omega$ MOSFET gate. This is about $1V_{th}$ and is the maximum thermal noise that can be tolerated in a fully optimized zero-power receiver.

An example of a high Q filter suitable for use as one filter element in a zero-power receiver is shown in figure 2. This is a surface acoustic wave (SAW) filter. It has a 3-dB bandwidth of about 1.3MHz with a center frequency of 2.60GHz, giving it a Q of about 2000. It has an input impedance of 55Ω and an output impedance of 57Ω, a reasonably good match for a dipole antenna's 73Ω impedance. The insertion loss of this device (17.7dB) should ideally be much lower to improve the range of any zero-power receiver that it is used in. When a high Q SAW such as this is used in conjunction with a typical crystal filter with a typical Q of 1000-2000, the combined circuit Q is well over the 100,000x required minimum. Crystal filters are available as COTS components, but commercially available crystal filters are deliberately detuned to lower the Q and increase the passband. A custom crystal filter with a Q of 10,000 to 100,000 is ideal as a lower frequency filter in a zero-power radio receiver.

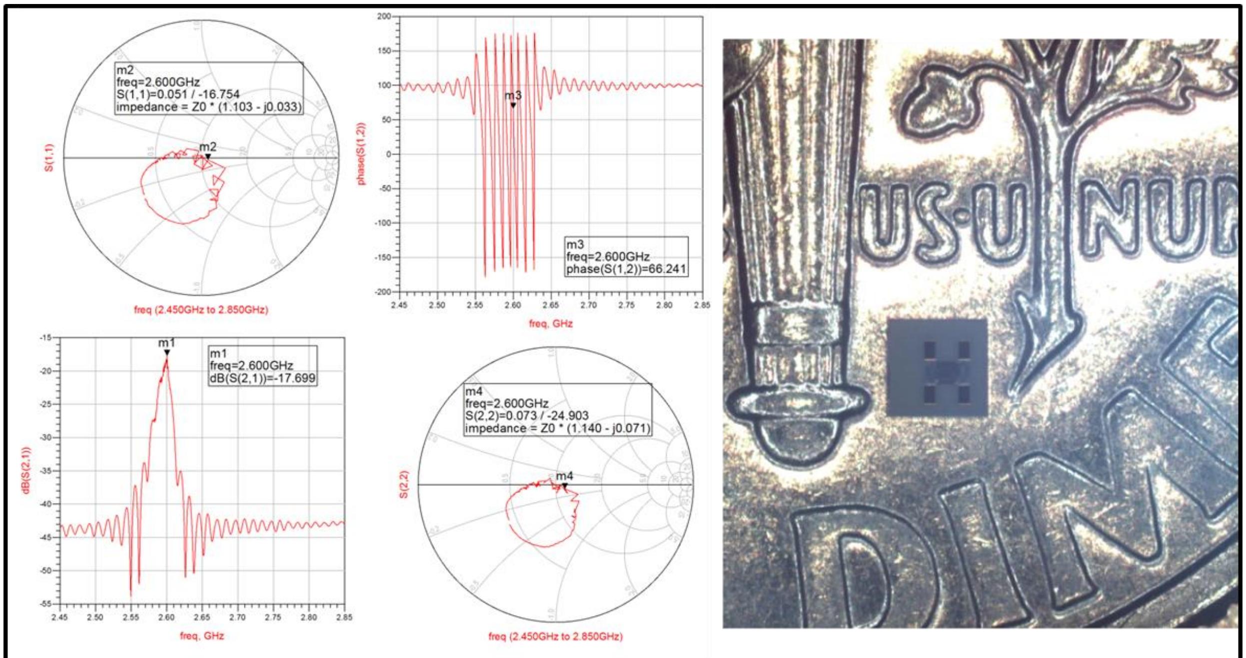

Figure 2: SAW filter S-parameters and single device sitting on the back of a dime.

5. Impedance Transformer/ Demodulator

The next element following the bandpass filter in the zero-power receiver is the impedance transformer/ demodulator. It serves the purpose of stepping the signal from microwave frequency to intermediate or baseband frequency. While doing this it must accomplish two additional steps. First, it must transform the signal from the impedance of the microwave components (typically 50Ω to 300Ω) to the impedance of the intermediate or baseband frequency components (typically 100kΩ to 10^{12}Ω). Second, it must demodulate the RF or microwave signal and transform it into a baseband signal. That is, it must demodulate the signal by removing the microwave carrier and leaving behind the signal modulation at the intermediate or baseband frequency. This must be accomplished because the microwave or RF signal will be greatly attenuated by any parasitic capacitances in a high impedance environment where $Z_o > 10k\Omega$.

A marginal performance demodulator will transform the impedance from the RF environment (50Ω) to an intermediate value impedance (10kΩ - 100kΩ). This has been the approach taken using microwave detection diodes. Microwave detection diodes serve the dual role as demodulator/ detectors, and these typically transform the RF impedance onto a 100kΩ load [9]. Also, much of the input power to a microwave detector diode is often reflected, since the diode small signal resistance is typically 1-10kΩ and is difficult to match to the 50Ω input impedance. The microwave diode creates a direct electrical connection between the 50Ω input impedance and the output load resistor. The video resistance (typically 1-10kΩ) of the diode serves as the small signal transfer element between the input and output, in effect, limiting the amount of voltage gain that can be achieved across the diode. Microwave detection diodes have been the demodulator of choice in commercial zero-power receivers, and this impedance transfer limitation has put a cap on the range and performance of these zero-power receivers for decades.

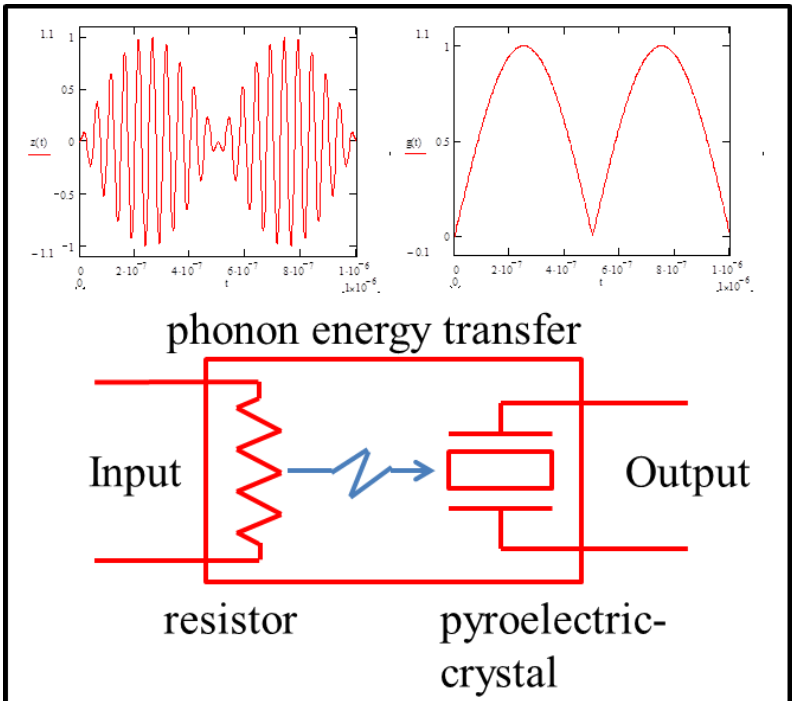

Figure 3: Pyroelectric demodulator

The solution to this problem is to use the pyroelectric demodulator as the impedance transformer/ demodulator, as described in [1], and as pictured in figure 3. The pyroelectric demodulator presents a pure resistive load to the filter or antenna that it is attached to. Its input resistance can be fabricated to be the exact value desired to form a match to the input device. The demodulator accepts an amplitude modulated RF signal across its input resistor. The input resistor is thermally isolated from any path of heat transfer except through the pyroelectric crystal underneath it. The resistor heats and cools rapidly, driven by the modulated RF signal (shown in the upper left-hand side of figure 3). The resistor is small enough to heat up and cool down very rapidly. That is, it is small enough to follow the modulation of the incoming RF signal, but not fast enough to follow the RF signal itself. As a result, the temperature variation of the resistor follows the modulation curve of the RF signal (shown in the upper right-hand side of figure 3). The pyroelectric element that the resistor sits on top of responds to temperature variations by producing an output voltage that follows the alternating temperature of the pyroelectric element. The pyroelectric element behaves like a capacitor with a very low leakage current. In electrical terms, it looks like a capacitor in parallel with a very large resistor, typically about 10^{12} Ω. This means that it is possible to transfer power from an RF source at 50Ω to the input of a MOSFET transistor with an input resistance of 10^{12} Ω. This impedance transfer capability of the pyroelectric demodulator enables effective transfer of power from the input to the output of the device in accordance with the responsivity of the device. The responsivity of a device is the voltage (or current) out of a device for a given power input to the device.

A MEMS integrated version of the pyroelectric demodulator geometry is shown in figure 4. The detector is a two-port microwave device. The input port accepts the microwave signal, which is an amplitude modulated microwave sinusoid. The top layer of the structure consists of a thin film resistive material. The material can be any standard thin film resistor material, such as NiCr or TaN. The input leads to the resistor must be thermally isolated to contain and direct the flow of thermal energy dissipated in the resistor. The objective is to thermally isolate the microwave signal in the resistor in all directions except one. The thermal energy must be forced into the thin pyroelectric layer, which should be thermally isolated from all of its surroundings except for the resistor. This can be accomplished by using MEMS processing to thermally isolate the detector in all directions except from the resistor on top. An anisotropic etch can be used to remove substrate material along the walls of the detector. Any heat dissipated in the resistor will be forced to travel down into the pyroelectric material. Connections using suspended bridge structures made of metal or polysilicon can also be created using MEMS processing to further thermally isolate the detector.

The layer immediately under the thin film resistor is the electrode material for the pyroelectric sensing layer. This should be a thin layer of metal or doped, highly conductive polysilicon that is electrically isolated from the overlaying thin film resistor by a very thin layer of insulator, such as Si_2O_3 or SiN. This will serve to electrically isolate the input port from the output port. Capacitive coupling between the microwave input signal line and the demodulated output signal line will still give some parasitic coupling from input to output.

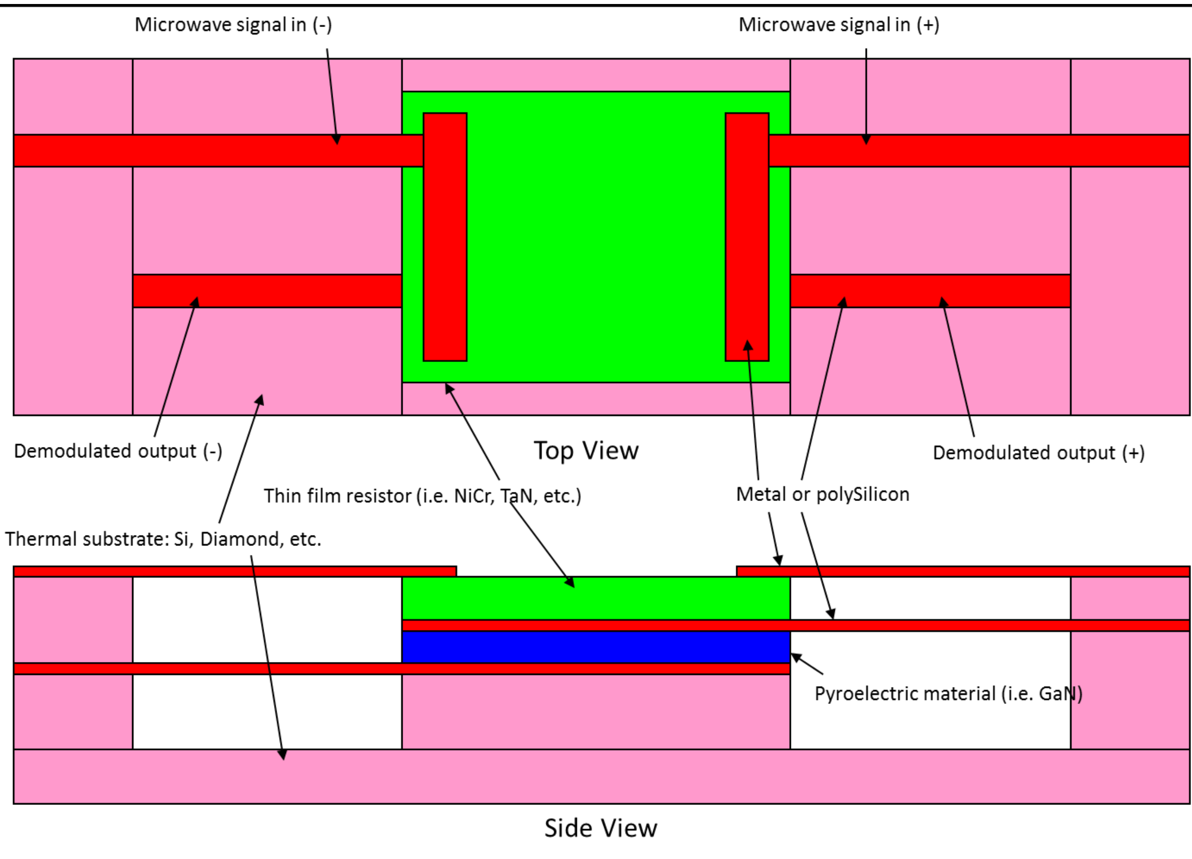

Figure 4: Pyroelectric demodulator integrated circuit fabrication geometry

The pyroelectric demodulator operates by first converting the electrical energy present in an amplitude modulated microwave signal into thermal energy. A potential microwave input signal is shown in the upper left corner of figure 3. This signal will be introduced at the input port. Its electrical energy will be entirely converted to thermal energy. The thermal energy from the microwave signal will couple out of the resistor, principally flowing downward into the pyroelectric material. The dimensions, geometry, and material of the underlying layers govern the time response of this thermal transfer. This thermal energy will induce a temperature change in the underlying pyroelectric layer. The pyroelectric layer will convert the temperature change into a voltage difference across the output port with a very short time lag, on the order of 1 psec. The thermal transfer across the pyroelectric layer can occur with an extremely fast response time. The objective is to tailor the thickness and volume of the pyroelectric layer so that the thermal transfer into it occurs at a rate that is greater than or equal to the maximum modulation bandwidth of a signal of interest.

Figure 5 contains some images of various pyroelectric demodulating detectors and their testing setups and results. There have been four different generations of pyroelectric demodulators built and tested at Sandia National Laboratories. The first-generation device was a hand-assembled, wirebonded device, built using a thick film resistor attached on top of a large lithium tantalate detector. The second-generation device was built up as a single, monolithic device on top of thinned $LiNbO_3$. It was not much smaller than the first-generation device and is shown in the upper left corner of figure 5. The third-generation device was built in a large array on PZT. The test setup and signal outputs from testing the third-generation device are shown in the other frames in figure 5. Fourth-generation devices have been built and will be described in subsequent publications.

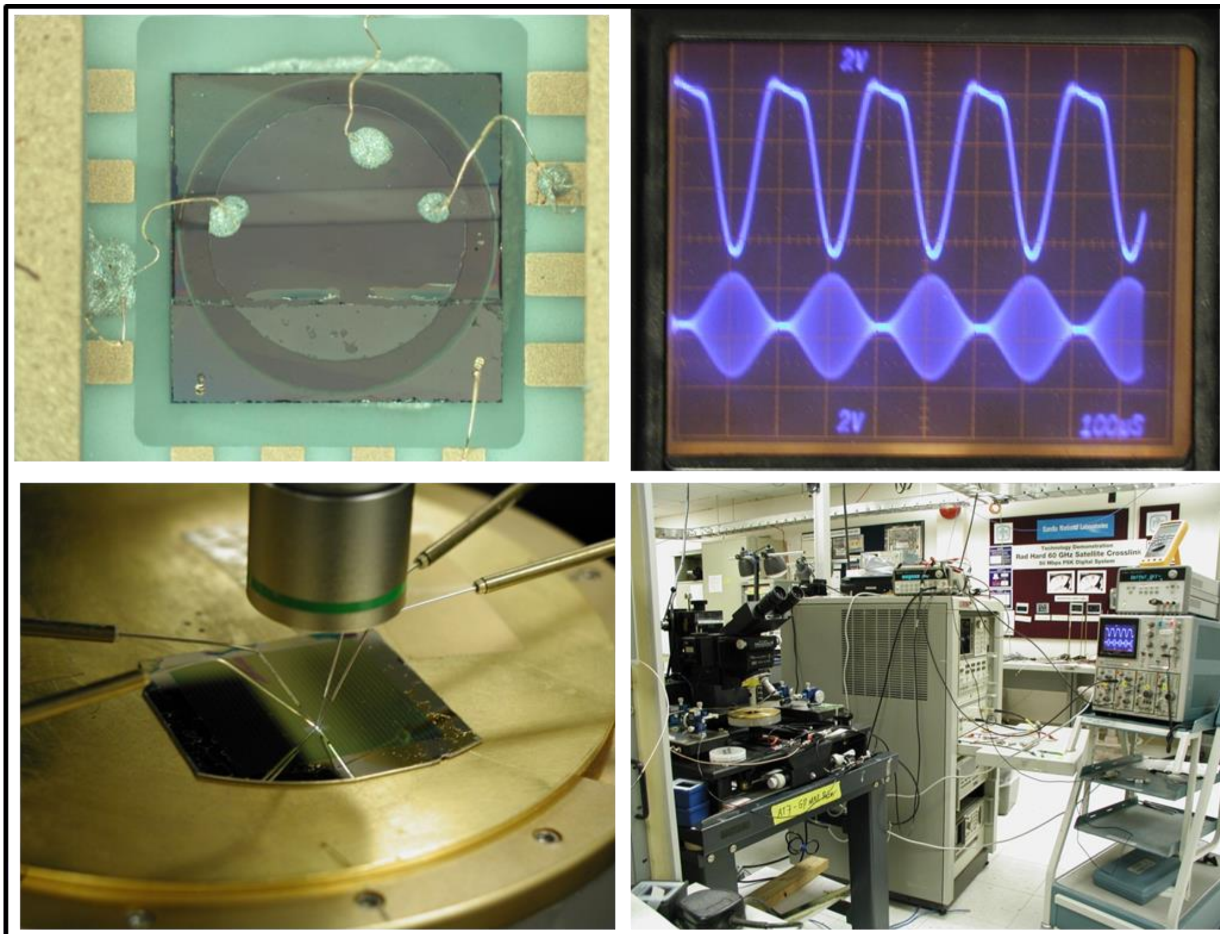

Figure 5: Second generation pyroelectric demodulator, third generation pyroelectric demodulator, probe station test setup, and output over input signals to and from the demodulator (ccw from upper left corner).

6. Detector

Conventional microwave diode detectors are built to combine the functions of the demodulator and the detector. That is, they serve both as an impedance transformer to convert the 50Ω microwave signal into a relatively high impedance (10-100kΩ) baseband signal while also serving as an AM demodulator and while simultaneously providing a DC level output that is proportional to the RF signal power. Most microwave detector diodes don't do a very good job performing these three functions, at least if no DC power is provided to serve to bias the detector diode on. The highest performance unbiased microwave detector diodes are Sb-heterostructure backward diodes and are described in [10] and [11]. Before looking at the best performing microwave detector diodes, which are not commercially available, it is worthwhile to look at typical COTS microwave detector diodes.

Zero-bias microwave detector diodes are used in a configuration that is referred to as a crystal video receiver [9]. There are two basic crystal video receiver configurations, making use of one or two zero-bias microwave detector diodes. These are shown in figure 6. The zero-bias microwave detector diode performs the job of the demodulator, impedance transformer, and DC output detector. A bandpass filter can be added to the crystal video receiver in-between the antenna and the microwave diode detector. This will lower the power gain of the entire circuit by the amount of the insertion loss of the bandpass filter, but it will improve the frequency selectivity by the ratio of the passband of the filter to the passband of the diode detector.

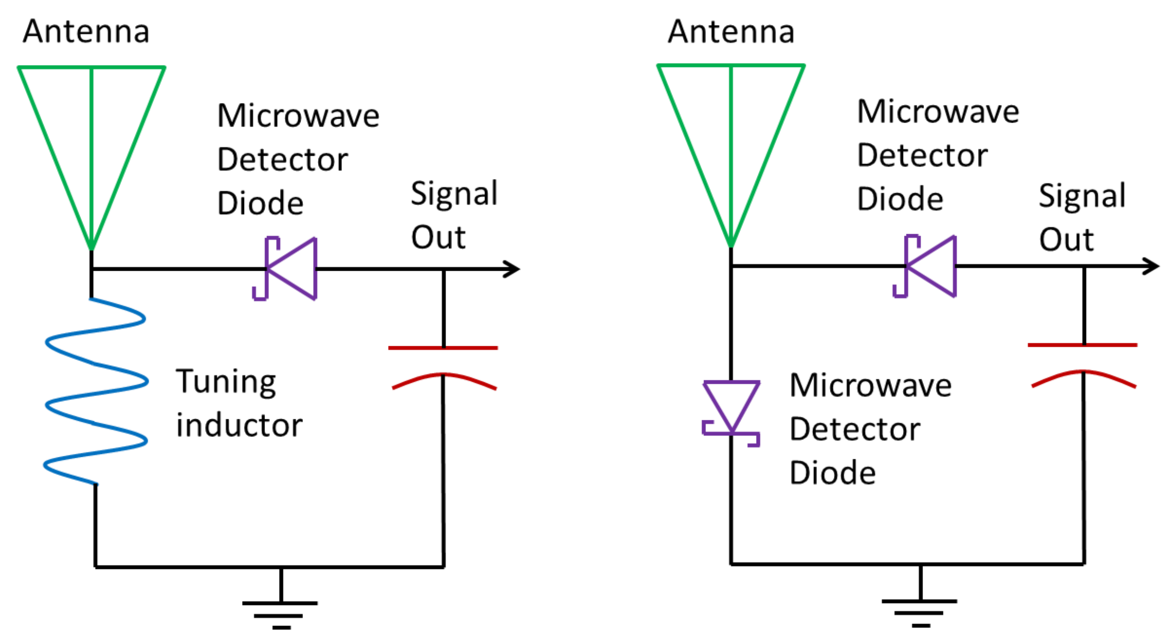

Figure 6: Zero-bias crystal video microwave receiver configurations

The crystal video receiver shown in figure 6 is a variant of the much older crystal-set receiver which was first patented in the year 1906, U.S. Patent 836,531 [12]. The original crystal radio circuit is shown in figure 7. It has the same elements as the zero-bias crystal video microwave receiver. The receiver front-end consists of an antenna with a tunable tank circuit to form a bandpass filter. The antenna is usually a dipole with an impedance that varies around the free-space, half-wave dipole impedance of 73Ω. Low frequency half-wave dipole antenna impedance typically varies between 60Ω and about 98Ω, depending on the height of the antenna above the ground.

For a detector, the crystal-set radio uses a low knee-voltage diode, such as the 1N34 germanium diode. The RF output is demodulated onto a capacitor and the output is received by the human ear through a high impedance audio earphone. The germanium diode offers good responsivity with a potential for some voltage gain in transforming from the low antenna impedance (73Ω nominally) into the high impedance of the headset (typically 2-3kΩ). There have been many diodes created for use in these crystal radios, and modern microwave detector diodes are the follow-on components to this technology. The first detector described in the first patented zero-power receiver (e.g. crystal-set radio) came to be known as the "cat's whisker" crystal detector. This was essentially a Schottky diode formed by manually pressing a thin metal stylus onto a semiconductor crystal. The different crystals marketed included galena (native PbS crystal), carborundum (SiC crystal), and silicon. The metal point had to be moved around to locate the best "hot spot" for reception. This was probably due to variations in the native impurity concentrations in the crystals. Radio manufacturers in the 1920's moved on to using tube-based diodes that utilized heated filaments as thermoelectric emitters. Unpowered crystal-set radios remained the purview of hobbyists. The "cat's whisker" crystal detector was replaced by the standard 1N34A germanium diode and its many variants (1N48, 1N54A, 1N81, etc.) [13]. Hobbyists have experimented with and compared these various crystals, diodes, and others [14]. The conclusion to draw from their work is that the 1N34A germanium diode represents the standard of performance of the early zero-bias, zero-power radio receivers. Powered vacuum diodes improved on the unpowered diodes substantially, since when properly biased, the diode "knee" voltage could be set to zero volts on the current versus voltage, diode I-V plot.

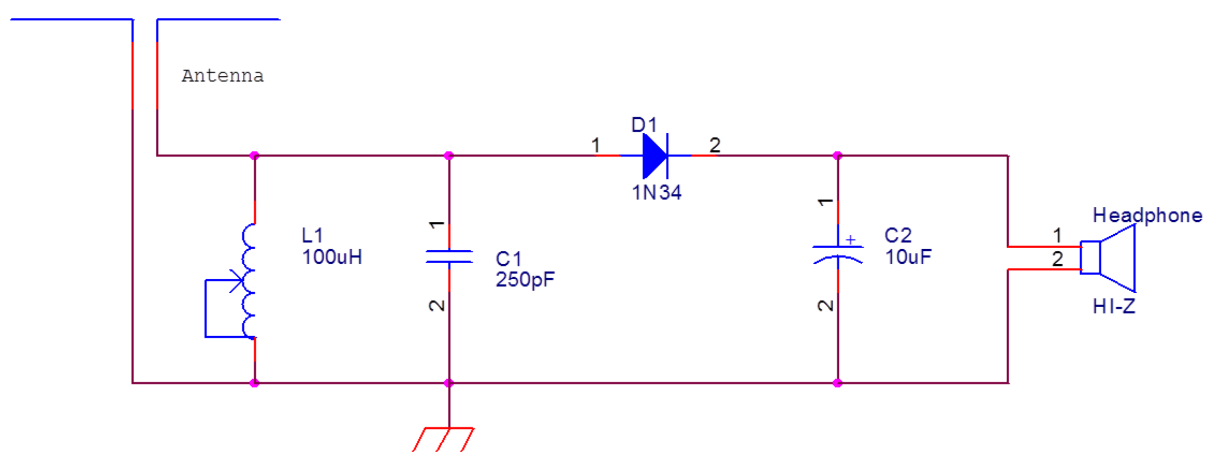

Figure 7: Basic crystal-set AM radio receiver

Zero-bias microwave detectors have been continually refined since the days of the crystal-set germanium diodes. A review of high performance microwave zero-bias detectors is available in [15]. There are two principal limitations with semiconductor, microwave zero-bias detectors. The first, already mentioned, limitation is that a microwave diode has a limited impedance transformation capability, from about 50 Ohms to about 100 kOhms. The second is that the diode knee voltage must be overcome by the received RF signal.

The pyroelectric demodulator does not have either of these disadvantages. First, a very large transformation in impedance can be made available by using the pyroelectric demodulator with a separate low threshold-voltage, high impedance field effect transistor (typically a MOSFET or JFET) as the detector. The MOSFET transistor provides an input impedance of 10^{12} Ω, giving a very large voltage gain potential due to the transformation from the input impedance of 50Ω. Second, the pyroelectric demodulator is a thermal device, rather than a semiconductor device, and consequently, it has no diode knee voltage to be overcome. If a MOSFET transistor is connected to the output of the pyroelectric demodulator, then it can be designed with a threshold voltage tailored to the application. For the design of a zero-power receiver, the threshold voltage could be set to about 100mV, or $4V_{th}$. The output transistor threshold voltage will be the same limitation for either a microwave detector diode-based receiver or a pyroelectric demodulator-based receiver, since either receiver will need a powered device to turn on.

7. The Invention: High Performance Zero-Power Radio Receiver

Using these concepts, it is possible to make a very sensitive zero-power radio receiver capable of detecting and decoding RF signals from a much greater range than has previously been possible. To briefly describe this new zero-power receiver: it is the use of one or more pyroelectric demodulators, paired with one or more high Q, narrowband filters, and a high input impedance detector. The combination of these elements provides a large range advantage over traditional zero-power radio receiver approaches. The two advantages of the new approach are voltage gain through the pyroelectric demodulator and the lack of a need to overcome a diode knee voltage. The wide modulation bandwidth of the pyroelectric demodulator can provide a third advantage for applications that require it. By pairing a pyroelectric demodulator with a narrowband filter, the desired signal can be separated from background noise and can be given voltage gain while only using power contained in the received signal. By using multiple stages of pyroelectric demodulators paired with narrowband filters, it is possible to demodulate to one or more IF frequencies before finally demodulating to the baseband frequency. The use of multiple stages with one or more IF frequencies enables filtering to a narrower final frequency than is possible using a single filter. That is, the use of multiple pyroelectric demodulator/filter pairs in cascaded stages downconverting to one or more IF frequencies can provide better signal detection and noise rejection than is possible using a single pyroelectric demodulator/ filter stage. The use of multiple stages can provide a higher overall circuit Q than is achievable with a single filter.

The typical single stage implementation is to combine an antenna, a narrowband SAW filter, a pyroelectric demodulator, and a FET detection transistor. A two-stage implementation might combine an antenna, a narrowband SAW filter, a pyroelectric demodulator, a crystal filter, a second pyroelectric demodulator, and a FET detection transistor. Other implementations are possible and are covered in [2]. In either the single or the two-stage implementation, it is important to match the input impedance of each stage to the output impedance of the preceding stage. For the configuration shown in figure 1, the input SAW filter should have an input impedance that is matched to the receiver's antenna. The input resistor used in the pyroelectric demodulator that connects to the SAW filter should be matched to the output impedance of that SAW filter. The pyroelectric demodulator then provides an impedance transformation to the following stage, whether it is a low impedance crystal filter used in a multiple stage design or a high impedance FET transistor used in a single stage design. The detection transistor needs to be a high input impedance device with a low threshold voltage and low off-stage leakage current. This can be provided by a MOSFET or JFET transistor. The use of a custom SAW filter at the front end of the receiver provides an additional advantage in that it may be designed with a very low input impedance matched to a small aperture antenna. The small aperture antenna can be a short dipole or small loop antenna.

These components can be integrated on a single substrate, with or without an integrated antenna. The narrowband SAW filter, the pyroelectric demodulator, and a detector transistor are shown in figure 8. The components shown in this figure were designed and optimized as separate elements and are shown only to give a concept of the size of the zero-power receiver under consideration. As described in the Basic Elements section, it is possible to cascade multiple filter/ demodulator sections to achieve a narrower bandwidth reception than is possible using a single section. This approach mimics the action of a superheterodyne receiver in that the output of the first filter/ demodulator section will be an intermediate frequency (IF) rather than the baseband frequency. The output of the final filter/ demodulator section is at the baseband frequency. For example, one could create a two-stage design with the output of the first stage being at the commonly used 10.7 MHz IF frequency. The first section could consist of a SAW bandpass filter centered at the carrier frequency followed by a pyroelectric demodulator with input impedance matched to the output impedance of the SAW filter. The output of the first pyroelectric demodulator will be at the IF frequency of 10.7 MHz. The output of the pyroelectric demodulator will couple into a high Q crystal filter at the IF frequency. These are available commercially with low in-band insertion loss (typically < 3dB) and high stopband attenuation (typically > 65dB).

The impedance of the crystal filter is typically a few kΩ's, intermediate between the impedance of the RF section and the output detector. This provides a means to step the voltage up part way from the input to the final output section. The output of the crystal filter is then followed by a second pyroelectric demodulator which transforms the impedance of the output of the crystal filter to the impedance of the detector. In the case of the example discussed later, the impedance of the output detector is made to be as high as possible, about $10^{12}\ \Omega$. The communication signal for the multiple-stage version consists of a carrier frequency with amplitude modulation at the IF frequency which is, in turn, modulated at the baseband frequency.

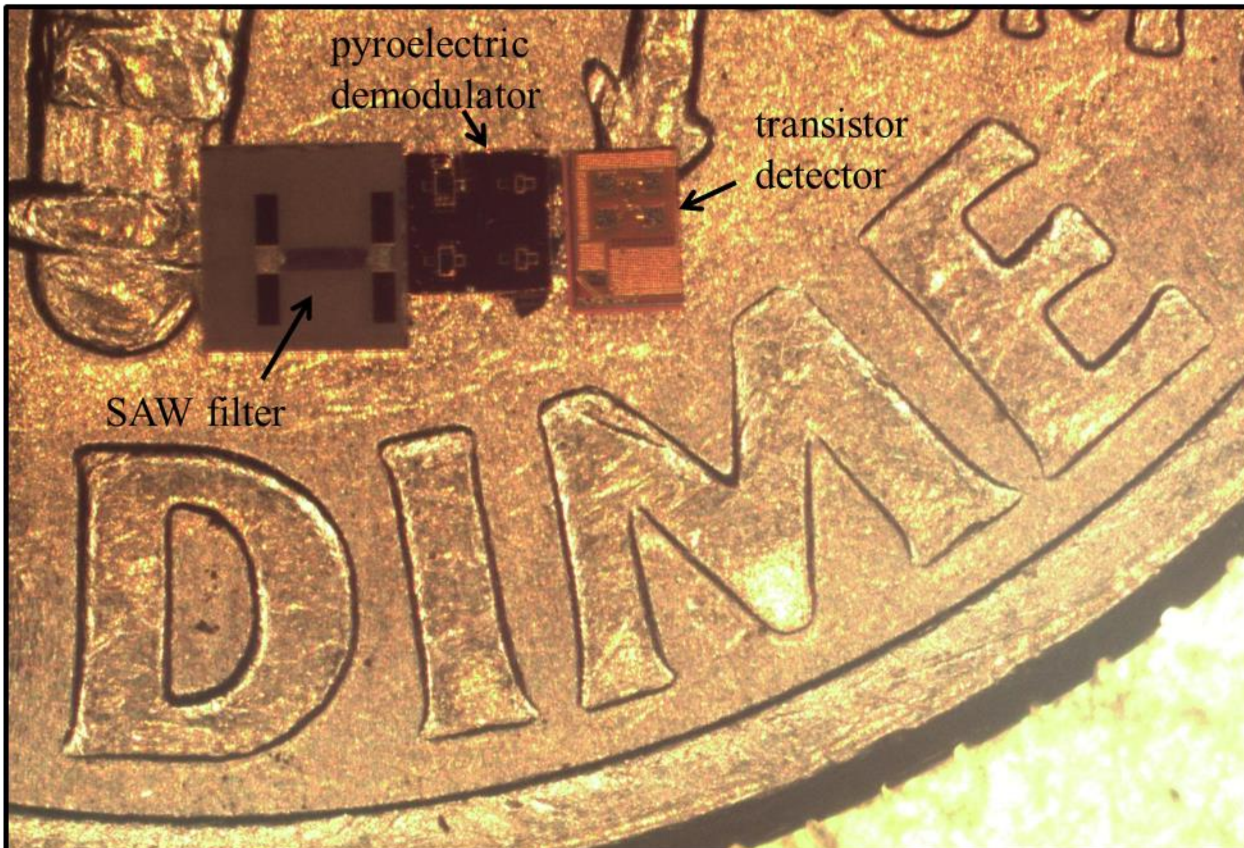

Figure 8: Zero-power receiver components on the back of a dime.

It may be desirable to further encode the baseband signal to enable activation of a decoder by a coded signal. This approach will create a form of code division multiple access. One application of this approach is using a transmitter to send a coded signal to awaken a receiver. To decode the signal, a powered decoder can be added to the receiver, as is done with most commercial wireless systems. However, it is also possible to add an unpowered SAW correlator to the back end of a high performance zero-power receiver to provide a coded unlocking mechanism.

A SAW correlator operates by summing the voltage of a time delayed sequence of the input signal. A SAW correlator can be designed to operate at a specific frequency chosen from a wide range of possible frequencies. It operates by summing phase or amplitude modulated RF or microwave carrier signals. Figure 9 shows the output of a SAW correlator (top) and the envelope detected final signal (bottom). The peak signal can be used as the final output to turn on a circuit that is external from the zero-power receiver. The use of the correlator provides additional coding gain that helps to differentiate a signal from background noise. This is explained in detail in [6].

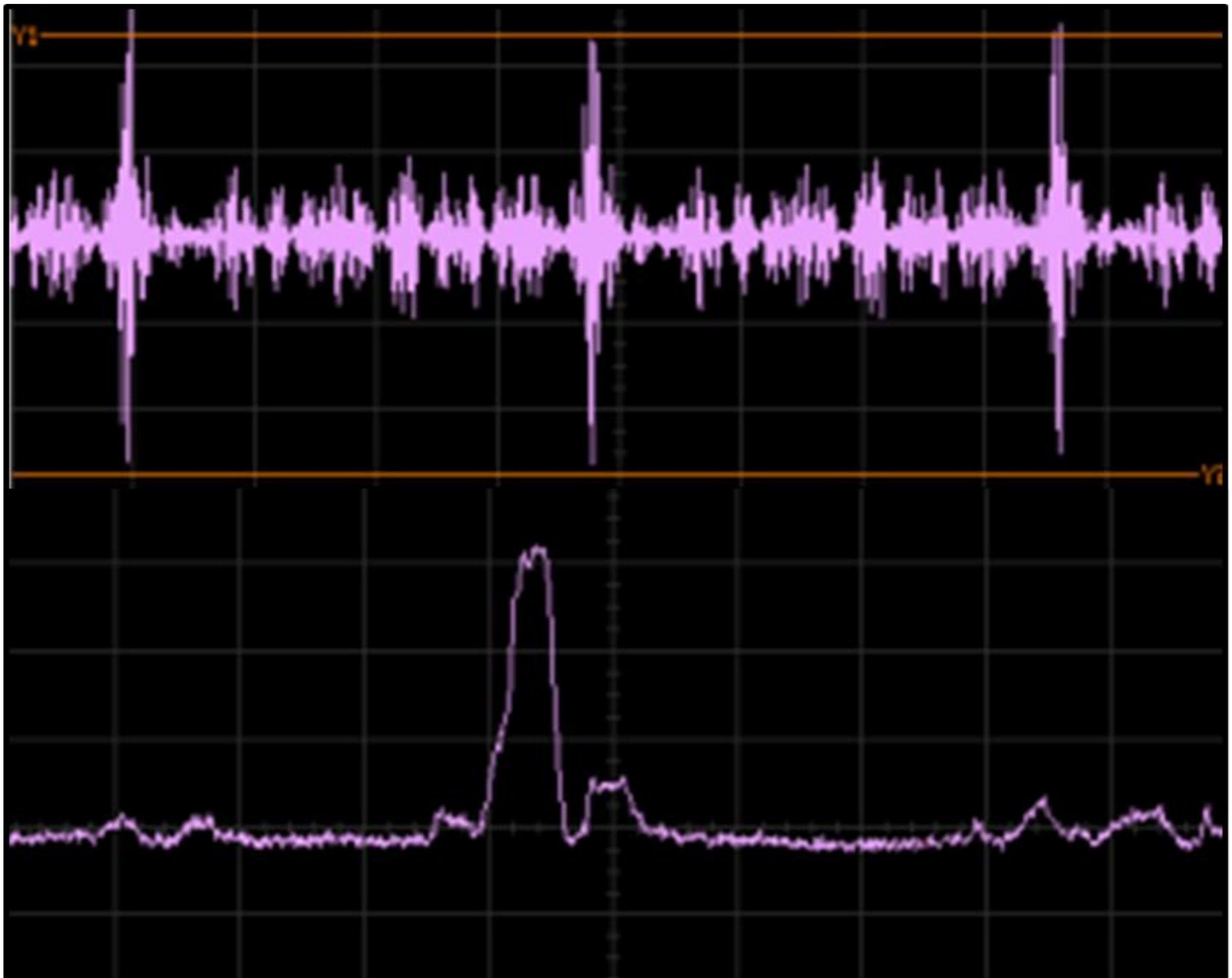

Figure 9: SAW correlator raw output (top) and demodulated (bottom).

The addition of a SAW correlator to the zero-power receiver adds little to the physical area required to fabricate a receiver. The basic components needed to fabricate a zero-power receiver with a decoded output are shown in figure 10. These include the SAW filter, pyroelectric demodulator, MOSFET transistor, and SAW correlator. Useful SAW correlators will typically have chip lengths up to 15 BPSK bits long. We have fabricated longer SAWs, but those with longer codes exhibited degraded performance. If a long code is needed for a particular application, it will be necessary to build a low power decoder that operates on battery power.

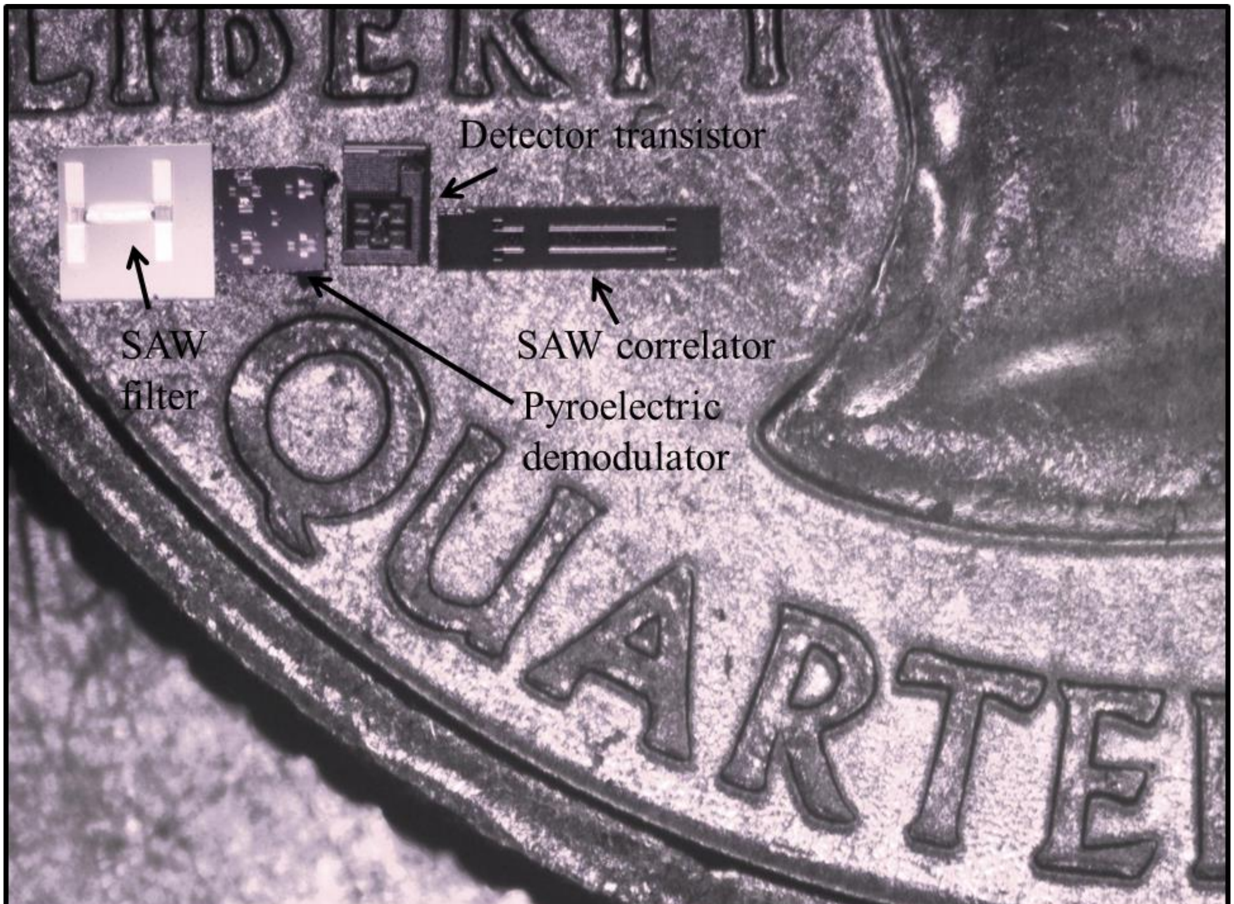

Figure 10: Zero-power receiver with SAW correlator for device ID.

8. Example Designs

Several numerical examples of zero-power receivers are presented here. These are used to demonstrate what has been historically possible, what is routinely used in commercial applications today, and what is possible using the zero-power receiver proposed here. All zero-power receivers need to reference an equation for range.

A communication link operating in free space at wavelength λ with a transmit antenna having a gain of G_T, a receive antenna having a gain of G_R, a transmitted power of P_T, where the receiver and the transmitter are separated by a distance of r, the power received at the front end of the receiver, P_i, is given by [16, page 255]:

$$P_i = \frac{P_T G_T G_R \lambda^2}{(4\pi r)^2} \tag{3}$$

Moving around terms in this equation gives the range, r:

$$r = \frac{\lambda}{4\pi} \sqrt[2]{\frac{P_T G_T G_R}{P_i}} \tag{4}$$

An effective signal-to-noise ratio at the front end of the receiver can be defined as:

$$SNR_{in} \triangleq \frac{P_i}{N_i} \tag{5}$$

Here the received noise, N_i, is given by [ref 16, page 258]:

$$N_i = kTB \tag{6}$$

where k is Boltzmann's constant (= 1.38 x 10^{-23} W/^{o}K-Hz), T is the temperature of the thermal noise source, and B is the bandwidth of the receiver. For a passive radio receiver, the bandwidth, B, must include the entire signal bandwidth that provides noise input to the receiver. That is, a 3-dB bandwidth will not provide a complete picture of the effects of noise on the received signal. A passive radio receiver has a very narrow 3-dB bandwidth, but it still receives some noise power from a wide range of frequencies that are outside of its 3-dB bandwidth and are heavily attenuated. The wide bandwidth of attenuated noise input is like multiplying a large number by a small number; the effect is not negligible and must be considered with care.

A passive radio receiver will introduce a loss in signal power between the front-end of the receiver and the point at which a decision is made between signal and noise. If the loss of the receiver is expressed as L_R, and the power received at the back-end of the receiver where the detector is present is P_D, and P_i is the input signal power at the front-end of the receiver, then:

$$L_R \triangleq \frac{P_D}{P_i} \tag{7}$$

The receiver noise figure, F, is defined as the ratio of the input to output signal-to-noise ratios [ref. 16, page 271]:

$$F \triangleq \frac{SNR_{in}}{SNR_{out}} \tag{8}$$

Using the result that the receiver noise figure is equal to the receiver loss (F = L_R), when the receiver network is a lossy line [ref. 16, page 275], the input power can be expressed as:

$$P_i = SNR_{in} N_i = SNR_{in} kTB = F\, SNR_{out} kTB = L_R SNR_{out} kTB \tag{9}$$

Radio receiver range then can be expressed as:

$$r = \frac{\lambda}{4\pi}\sqrt[2]{\frac{P_T G_T G_R}{L_R\, SNR_{out}\, kTB}} \tag{10}$$

where λ is the wavelength of the carrier frequency signal, G_T is the transmitter antenna gain, G_R is the receiver antenna gain, L_R is the receiver input chain loss, SNR_{out} is the required detection signal-to-noise ratio at the output, k is Boltzmann's constant, T is the receiver temperature, B is the receiver bandwidth. Equations (4) and (10) can be used to calculate maximum range for different zero-power receivers.

A: Example 1, Historic Crystal-set Radio

The first example is that of a crystal-set radio. The crystal-set radio has been around for over 100 years [12]. The parameters for the one used in this example are from a high-performance version. The crystal-set considered here uses a 1N34A germanium diode, a high impedance earphone, and a high-Q tank circuit, all of which were available during the crystal radio heyday era of the 1950's and 1960's. This radio uses a large, though easy-to-construct, dipole antenna that is a 50-ft. long wire. It is set up to receive in the center of the 31m band, the most popular shortwave band. The calculations included below indicate that the range figure-of-merit is 1.068 km/ $Watt^{1/2}$. That is, a 1 Watt transmitter can be received by this receiver at a range of 1 km, and it will produce a minimum audible sound pressure level of 50 dB out of the headphones. The range figure-of-merit is pretty good, which is as expected, since these receivers were used to listen to shortwave radio all over the world with the benefit of atmospheric waveguiding at various times of the day.

The range calculations performed here assume a $1/r^2$ path loss from the transmitter to the receiver. This is a typical free space path loss calculation, but actual path loss can vary greatly from this. As mentioned, atmospheric waveguiding in the HF bands can lower path loss below the $1/r^2$ value, increasing the reception range. Also, transmission through urban environments can increase path loss to up to $1/r^4$, greatly diminishing the reception range. The range-per-$Watt^{1/2}$ figure of merit is intended only for comparison purposes. The range calculation is a good estimate of what to expect, but actual results may vary.

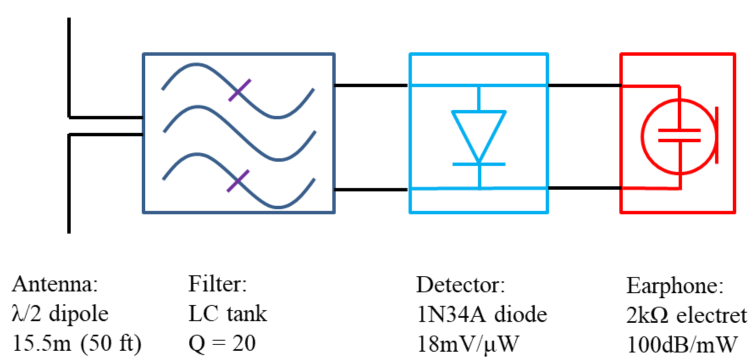

Figure 11: High end crystal-set receiver, set up to receive on shortwave band at 9.7MHz
size: huge, receiver figure-of-merit: 1 km/$Watt^{1/2}$

Simple receiver example: Signal reception for a crystal set radio

$P_T := 1$ Watt transmitter power to calculate figure-of-merit

$G_T := 1.64$ transmitter antenna gain, dipole, 2.1dBi

$G_R := 1.64$ receiver antenna gain, dipole, 2.1 dBi

$S_e := 100$ dB/mW earphone sensitivity, 2kOhm electret

$f_o := 9.7 \cdot 10^6 \quad \lambda := \dfrac{3 \cdot 10^8}{f_o}$ transmit frequency, 9.7 MHz, 31m band

$I_s := 2 \cdot 10^{-6}$ 1N34A diode saturation current

$nV_t := 0.026$ volts, diode thermal voltage

$C_j := 0.8 \cdot 10^{-12} \quad R_s := 20$ 1N34A diode junction capacitance and resistanc

$\gamma_1 := 5000$ V/W, measured 1N34A voltage responsivity

$R_L := 2000$ Ohms, earphone resistance

$R_v := \dfrac{nV_t}{I_s} \qquad R_v = 1.3 \times 10^4$ Ohms, diode video resistance

$\gamma_2 := \dfrac{\gamma_1 \cdot R_L}{R_L + R_v} \qquad \gamma_2 = 666.667$ V/W detector responsivity, w/ output load

$Q_t := 20 \qquad Z_{ant} := 73$ receiver tank circuit Q and antenna impedance

$R_t := Q_t \cdot Z_{ant}$ Ohms, LC tank circuit impedance = Q * Zantenna

$Z_d := R_s + \dfrac{R_v}{1 + (2 \cdot \pi \cdot f_o)^2 \cdot C_j^{\,2} \cdot R_v^{\,2}} \qquad Z_d = 9.294 \times 10^3$ Ohms

diode input impedance

Crystal set radio example, continued

$$\rho := \frac{Z_d - R_t}{Z_d + R_t}$$ mismatch between LC tank circuit and detector diode

$$\gamma_3 := \gamma_2 \cdot \left(1 - \rho^2\right) \qquad \gamma_3 = 312.884$$ V/W detector responsitivity, with mismatches

$$S_m := 50$$ dB, minimum threshold sound level out of earphone

$$P_d := 10^{\left(\frac{S_m - S_e}{10}\right)} \cdot 10^{-3}$$

$$P_d = 1 \times 10^{-8}$$ Watts, minimum power into earphones to produce threshold sound level

$$V_d := \left(R_L \cdot P_d\right)^{0.5} \qquad V_d = 4.472 \times 10^{-3}$$ detector volts to produce threshold sound

$$P_i := \frac{V_d}{\gamma_3} \qquad P_i = 1.429 \times 10^{-5}$$ Watts, received to produce threshold signal

$$m2km := 10^{-3}$$ range conversion factor for meters to kilometers

$$r := \frac{\lambda}{4 \cdot \pi} \cdot \left(\frac{P_T \cdot G_T \cdot G_R}{P_i}\right)^{0.5} \cdot m2km$$

$$r = 1.068$$ km/W^0.5, range-per-root Watt figure of merit for receiver

B: Example 2, Best Possible Commercially Available RFID Electronics

The next example that is considered is that of a high-performance RFID device created from commercially available components (Fig. 12). It uses a combination of different commercially available RFID technologies, including a dipole antenna, an LC front-end filter, an HSMS-2850 microwave detector, and an EPC RFID device serving as a back-end detector. The frequency of the RFID tag is 900MHz. The EPC RFID tag shown in figure 12 does not use the high-Q filter and high performance HSMS-2850 detector. These are commercially available components that were added to the front end of the EPC RFID tag to increase its range. Consequently, the tag shown at the top of figure 12 has a much shorter range than the hardware described at the bottom of figure 12 and that was used in the calculations that follow. The EPC RFID tag is shown to give a sense of how the antenna and tag chip are packaged together.

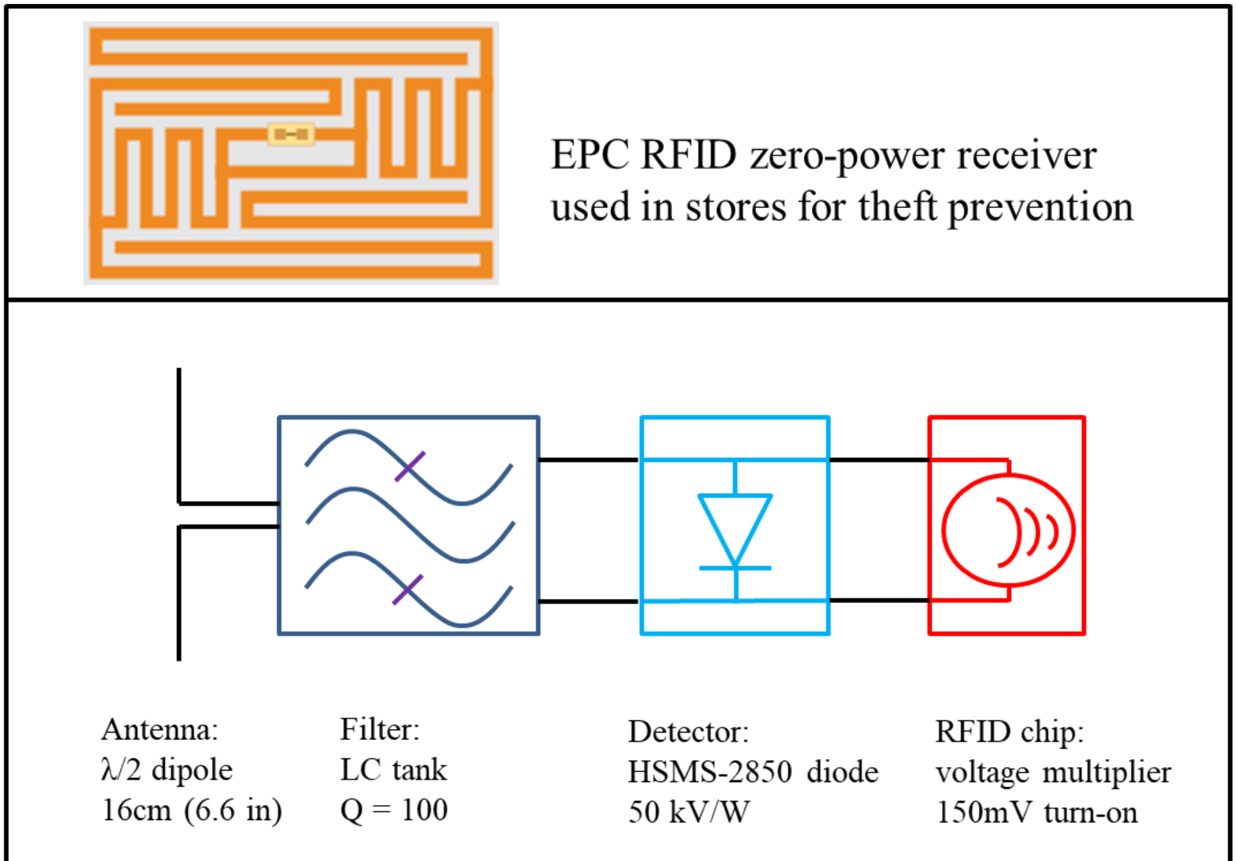

Figure 12: Improved RFID zero-power receiver, ISM band at 900MHz with a receiver range FOM of 0.019 km/Watt$^{1/2}$.

The composite RFID tag (Fig. 12) uses commercially available technology to build a zero-power receiver. The antenna is an omnidirectional dipole with gain of 2.1dBi and impedance of about 73Ω. When stretched out, it is 6.6 inches long, but it can be folded around a credit card-sized plane. The input filter is an LC tank with a Q of about 100, and it provides both input filtering and some matching between the 73Ω antenna and the detector. The detector diode is a Hewlett-Packard HSMS-2850, a high performance zero-bias microwave detector diode for use in the 900MHz -5.8GHz frequency range. The HSMS-2850 has a high responsivity, γ_1 = 50 kV/W, according to its data sheet [9]. The detector output is connected to a 100kΩ load, in accordance with the design specifications. The output is then processed by the EPC RFID chip, which is similar to the one described in [17].

Commercial RFID zero-power receiver example:

$P_T := 1$ Watt transmitter power to calculate figure-of-merit

$G_T := 1.64$ transmitter antenna gain, dipole, 2.1dBi

$G_R := 1.64$ receiver antenna gain, dipole, 2.1 dBi

$f_o := 900 \cdot 10^6$ $\lambda := \dfrac{3 \cdot 10^8}{f_o}$ transmit frequency, 900MHz ISM band

$C_j := 0.15 \cdot 10^{-12}$ $R_s := 55$ HSMS-2850 diode junction capacitance, series resistance

$\gamma_1 := 50000$ V/W diode detector responsitivity, given in data sheet

$R_L := 100 \cdot 10^3$ Ohms, detector load resistance at output, given in data sheet

$R_V := 8000$ Ohms, diode video resistance, given in data sheet

$\gamma_2 := \dfrac{\gamma_1 \cdot R_L}{R_L + R_V}$ $\gamma_2 = 4.63 \times 10^4$ V/W detector responsivity, w/ output load

$Z_d := R_s + \dfrac{R_V}{1 + \left(2 \cdot \pi \cdot f_o\right)^2 \cdot C_j^{\,2} \cdot R_V^{\,2}}$ $Z_d = 225.04$ Ohms, diode impedance at fo

$\rho := \dfrac{Z_d - 50}{Z_d + 50}$ mismatch between input circuit and the detector diode

$\gamma_3 := \gamma_2 \cdot \left(1 - \rho^2\right)$ $\gamma_3 = 2.755 \times 10^4$ V/W detector responsitivity, w/ mismatch

$V_d := 0.15$ minimum detector output voltage to turn on RFID voltage multipliers

$P_i := \dfrac{V_d}{\gamma_3}$ $P_i = 5.446 \times 10^{-6}$ Watts, received to produce min detectable signa

$m2km := 10^{-3}$ range conversion factor for meters to kilometers

$$r := \frac{\lambda}{4 \cdot \pi} \cdot \left(\frac{P_T \cdot G_T \cdot G_R}{P_i}\right)^{0.5} \cdot m2km$$

$r = 0.019$ km/W^0.5, range-per-root Watt figure of merit for receiver

The output of the HSMS-2850 detector diode needs to put out enough voltage to turn on the voltage multipliers in the RFID chip. The Schottky diodes in a high-performance RFID chip typically have a threshold of about 100mV, but the voltage multipliers need at least 150mV to be able to begin to function [17]. Consequently, the HSMS-2850 must put out at least 150mV for the detector chip to accurately receive an input signal. Taking the minimum detector voltage output of 150mV gives a receiver sensitivity figure-of-merit of 0.019 km/Watt$^{1/2}$. The figure-of-merit of 0.019 km/Watt$^{1/2}$ means that a 1 Watt transmitter is range limited at 19 meters for this best possible commercial RFID. The significant range advantage which the receiver of Example 1 has over the receiver of Example 2 is due to the large antenna, the low frequency, and the advantage which the human ear has over the RFID chip.

C: Example 3, Pyroelectric demodulator based zero-power receiver

In this example, we examine the performance of a zero-power radio receiver based on a pyroelectric demodulator. The radio receiver described in this example is a narrowband, low data-rate device used solely to wake up a wider bandwidth receiver. Such a device could be useful in a cellular telephone, in wireless sensor networks, or in other low power applications where a device needs to remain dormant until signaled by an external communications device.

A pyroelectric device generates charge, Q (Coulombs), in proportion to its pyroelectric coefficient, p (Coulombs/m^2 ºK), area, A (m^2), and temperature, T (ºK), according to:

$$Q = p\,A\,T \tag{11}$$

The voltage across the output of the pyroelectric device, V (Volts), is determined by the charge, Q (Coulombs), and capacitance, C (Farads) according to:

$$Q = C\,V \tag{12}$$

Here the capacitance, C (Farads), is approximately calculated from the permittivity of free space, ε_o (8.85 x 10^{-12} Farads/m), the relative permittivity of the pyroelectric material, ε_r (unitless), area, A (m^2), and thickness, d (m), according to the parallel plate capacitor equation:

$$C = \frac{\varepsilon_o\,\varepsilon_r\,A}{d} \tag{13}$$

Combining these equations gives a relation for the voltage across the output of the pyroelectric device, V (Volts), as:

$$V = \frac{p\,d\,T}{\varepsilon_o\,\varepsilon_r} \tag{14}$$

In this equation, the temperature, T (ºK), is determined from the input power, P_{in} (Watts), the thermal conductivity, G_{th} (W/m ºK), of the pyroelectric material to its surroundings, the volumetric heat capacity of the pyroelectric material, C_v (J/m^3 ºK), the area of the pyroelectric material, A (m^2), and its thickness, d (m). The temperature as a function of time, T(t, z), is then governed by the heat flow equation:

$$\frac{\partial\,T(t,z)}{\partial t} = \frac{G_{th}}{C_v}\,\frac{\partial^2 T(t,z)}{\partial z^2} \tag{15}$$

For highest accuracy, this equation would be solved in three dimensions. However, the predominant direction of heat flow for a pyroelectric demodulator, as shown in figure 4, will be in the vertical direction. The RF signal introduced to the demodulator supplies a power, P_{in} (W), which serves as a driver of heat into the device. The equation (13) then becomes:

$$\frac{P_{in}}{A\,d} = C_v \frac{\partial\, T(t,z)}{\partial t} - G_{th} \frac{\partial^2 T(t,z)}{\partial z^2} \tag{16}$$

This equation can be simplified by considering the single dimensional heat flow across the boundary between the pyroelectric material and the substrate that it is mounted on:

$$P_{in} = C_v A\, d \frac{dT(t)}{dt} + G_{th} \frac{A}{d}\, T(t) \tag{17}$$

The input power to the demodulator, P_{in}, can be taken as a sinusoidal signal at frequency f_o with peak level P_o and can be written as:

$$P_{in} = P_o\, sin(2\,\pi\, f_o\, t) \tag{18}$$

Using the Laplace transform to convert this and ignoring initial conditions gives:

$$\frac{P_o\, \omega_o}{s^2 + \omega_o{}^2} = C_v A\, d\;\; sT(s) + G_{th} \frac{A}{d}\, T(s) \tag{19}$$

This has a general solution given by:

$$T(t) = \frac{P_o\, \omega_o}{C_v A\, d} \frac{1}{(\frac{G_{th}}{C_v\, d^2})^2 + \omega_o{}^2} \left[-\cos(\omega_o\, t) + \frac{G_{th}}{C_v\, d^2\, \omega_o} \sin(\omega_o\, t) + e^{-\frac{G_{th}}{C_v\, d^2} t} \right] \tag{20}$$

This solution can be simplified by taking the steady state solution ($t >> 1$) and by considering that $G_{th}/C_v\, d^2 >> \omega_o$ and $G_{th}/C_v\, d^2 >> 1$. Then the temperature is given by

$$T(t) = \frac{P_o\, d}{G_{th} A} \sin(\omega_o\, t) \tag{21}$$

and the device peak voltage output is given by

$$V = \frac{p}{\varepsilon_o\, \varepsilon_r} \frac{P_o\, d^2}{G_{th} A} \tag{22}$$

Using (13) for the pyroelectric element capacitance, the responsivity (Volts/Watt) is then given by

$$R_v = \frac{V}{P_o} = \frac{p\, d}{C\, G_{th}} \tag{23}$$

For lithium tantalate, the pyroelectric coefficient, p, is 23 nC /(cm^2 °K), the relative permittivity, ε_r, is 43, and the thermal conductivity, G_{th}, of the device fabricated on a layer of SiO_2 with a 10nm interface layer is 0.2 W/(m °K). A 0.5 x 0.5 µm device with a 300-nm thickness fabricated on lithium tantalate has a capacitance of 0.3 fF. Such a device will have a responsivity of

$$R_v(lithium\ tantalate, 0.5x0.5um) = \ 1.1\ x\ 10^6\ V/W \tag{24}$$

Pyroelectric demodulator, single stage zero-power receiver example:

$P_T := 1$ Watt transmitter power to calculate figure-of-merit

$G_T := 1.64$ transmitter antenna gain, dipole, 2.1 dBi

$G_R := 1.64$ receiver antenna gain, dipole, 2.1 dBi

$G_{R3} := 0.001$ receiver antenna gain, dipole, -30 dBi

$f_{o1} := 900 \cdot 10^6$ $\lambda_{1.} := \frac{3 \cdot 10^8}{f_{o1}}$ transmit frequency, 900 MHz ISM band

$f_{o2} := 9.7 \cdot 10^6$ $\lambda_2 := \frac{3 \cdot 10^8}{f_{o2}}$ transmit frequency, 9.7 MHz 31m band

$f_{o3} := 60 \cdot 10^3$ $\lambda_3 := \frac{3 \cdot 10^8}{f_{o3}}$ transmit frequency, 60 kHz WWVB signal

$R_{in} := 73$ Ohms, pyroelectric demodulator input resistance

$\gamma_1 := 1.15 \cdot 10^6$ V/W lithium tantalate, pyroelectric demodulator responsivity

$R_L := 10^{12}$ Ohms, N-channel FET input resistance

$R_v := 1000$ Ohms, pyroelectric demodulator video resistance

$\gamma_2 := \frac{\gamma_1 \cdot R_L}{R_L + R_v}$ $\gamma_2 = 1.15 \times 10^6$ V/W detector responsivity, w/ output load

$\rho := \frac{R_{in} - 73}{R_{in} + 73}$ mismatch between antenna and the pyroelectric demodulator

$\gamma_3 := \gamma_2 \cdot \left(1 - \rho^2\right)$ $\gamma_3 = 1.15 \times 10^6$ V/W detector responsitivity, w/ mismatch

$V_{th} := 0.15$ Volts, detector FET threshold voltage

$P_i := \frac{V_{th}}{\gamma_3}$ $P_i = 1.304 \times 10^{-7}$ Watts, received to produce min detectable signal

$m2km := 10^{-3}$ range conversion factor for meters to kilometers

Pyroelectric demodulator, single stage zero-power receiver example, continued:

$$r_1 := \frac{\lambda_{1.}}{4 \cdot \pi} \cdot \left(\frac{P_T \cdot G_T \cdot G_R}{P_i} \right)^{0.5} \cdot m2km \qquad r_2 := \frac{\lambda_2}{4 \cdot \pi} \cdot \left(\frac{P_T \cdot G_T \cdot G_R}{P_i} \right)^{0.5} \cdot m2km$$

$$r_3 := \frac{\lambda_3}{4 \cdot \pi} \cdot \left(\frac{P_T \cdot G_T \cdot G_{R3}}{P_i} \right)^{0.5} \cdot m2km$$

$r_1 = 0.12$ km/W ^0.5, range figure of merit for receiver at 900 MHz, ISM band

$r_2 = 11.176$ km/W ^0.5, range figure of merit for receiver at 9.7 MHz, 31m band

$r_3 = 44.615$ km/W ^0.5, range figure of merit for receiver at 60 kHz, WWVB signal

Using the calculated responsivity for the pyroelectric demodulator in the same equations used for the previous two examples produces figures-of-merit for a pyroelectric demodulator based zero-power receiver operating under various conditions. These calculations are shown in the boxes labeled, Pyroelectric demodulator, single stage zero-power receiver example. Three different range calculations are included. The r_1 range calculations were created using the same parameters that were used for Example 2, the best possible zero-power receiver using commercially available electronics. This is for a transmitter operating in the 900 MHz ISM band with dipole antennas for both the transmitter and receiver. For this comparison, the pyroelectric demodulator based receiver will have a range of 120 m per $W^{1/2}$ versus a range of 19 m per $W^{1/2}$ for the best possible receiver using commercially available components. This is more than a 6-fold improvement in range, and it suggests that many applications which use the ISM bands could make effective use of the pyroelectric demodulator based zero-power receiver technology.

The r_2 range calculations were created using the same parameters that were used for Example 1, the crystal-set receiver, except that the pyroelectric demodulator is used in place of the 1N34A detector. This is for a transmitter operating on the shortwave band at 9.7 MHz with dipole antennas for both the transmitter and receiver. For this comparison, the pyroelectric demodulator based receiver will have a range of over 11 km per $W^{1/2}$ versus a range of about 1 km per $W^{1/2}$ for the crystal-set receiver. Crystal-set receivers are no longer generally used, and this 11-fold improvement factor is just included for reference purposes. The pyroelectric demodulator not only provides a large improvement in range, but it is also capable of operating over a very wide range of carrier frequencies. The HSMS-2850 diode, described in example 2, and other high-performance microwave detector diodes typically have a minimum operating carrier frequency of 10 MHz.

Finally, an r_3 range calculation is included using parameters for reception of the time-synchronization signal transmitted by the WWVB operated by the National Institute of Standards and Technology (NIST). NIST operates a 70-kW transmitter with an amplitude modulated 60 kHz signal. This signal can be easily received over the entire North American continent using a pyroelectric demodulator based zero-power receiver with a small, very low gain antenna having the r_3 range figure-of-merit of 44.6 km per $W^{1/2}$. The detector already described, with a small but poorly performing -30dBi receiving antenna will be able to receive the station at a maximum effective range of 11800 km from the station near Fort Collins, Colorado. Since the WWVB signal travels predominantly as a ground wave, the signal can propagate for very long distances following the curvature of the earth [18]. A receiver of adequate sensitivity, as described here, could operate at up to the maximum effective range.

9. CONCLUSIONS

In this work, an overview of zero-power radio receivers was presented, along with a description of the basic elements that go into making a zero-power radio. The various types of pre-existing zero-power radios were described, including crystal-set receivers and current state-of-the-art, commercially available, passive RFID devices. A new type of zero-power radio receiver was then described, which is based on the pyroelectric demodulator. Design examples for three different zero-power radio receivers were presented, including one that uses the pyroelectric demodulator. Numerical computations were presented to demonstrate that the pyroelectric demodulator based receiver can receive signals at 6x-11x the range of the other high-performance zero-power receivers discussed. Several potential commercial applications for this receiver were discussed. The values used in numerical computations for the pyroelectric demodulator-based zero-power receiver are representative of a medium level of development of the pyroelectric demodulator. Ultimate performance could be much better than is represented by the range figures-of-merit presented here, if the technology is developed sufficiently to attain a high level of performance.

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