IEEE P802.15
Wireless Personal Area Networks

Project	IEEE P802.15 Working Group for Wireless Personal Area Networks (WPANs)
Title	**PHY Proposal Using Dual Independent Single Sideband, Non-coherent AM and Defined Unit Pulse**
Date Submitted	March 3, 2003
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Re:	**15.3a CFP Response presentation**
Abstract	This proposal is a combination of: 1) a baseband waveform with desirable properties for a bandpass medium, and 2) where the unit pulse period is several bit durations and overlapped to double the information transfer rate by creating three amplitude levels, and 3) linear translation of the data bearing waveform to a single radio sideband, and 4) two instances of that translation such that the homodyne image of one sideband falls inverted in the passband of the other. This contribution contains: a) mathematical simulations of the baseband data waveform properties, and b) implementation description in support of asserted execution simplicity, and c) results available from recent work on system simulation (continuing).
Purpose	To show evidence of advantage of analog-intensive approach to high speed data transmission with particular emphasis on optimization criteria and definition of propagation environment.
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Proposal–SSB and DSB AM Modulation for High-rate Data Microwave Radio Systems

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Proposal–SSB and DSB AM Modulation for High-rate Data Microwave Radio Systems

Summary

The application addressed is high-rate microwave radio embeddable in wearable and portable devices requiring high digital transfer rates over relatively short distance. It is desirable to realize as much range and first-try transfer success as can be obtained from relatively simple and low power equipment.

Problem Definition

The problem addressed is transmission of high speed data at microwave radio frequencies with greater tolerance to the distortion from inevitable multipath radio propagation and like-signal interference. The first premise is:

> The data information must be entirely amplitude coded so that the phase of the local oscillator that recovers the modulation does not have to be locked to the phase of the remote transmitter. The phase of the received signal relative to that transmitted will be variable as received over a radio path making recovery more prone to error.

Many widely used modulations do use phase and amplitude modulation together resulting in a constellation of phase–amplitude possibilities. These may be called "vector modulations." Those using A/D converters and fast digital signal processors can deduce the demodulating phase by mathematical methods operating at speeds much greater than the data rate. As the number of points in the constellation is increased to provide higher data rates in a fixed bandwidth, the required signal-to-noise ratio for low error rate increases. At first, the interference is assumed to be white noise, but in fact it may be internal as a result of the modulation choice.

> Internal noise with constellation modulation can come from propagation-caused crosstalk between the I and Q phases each carrying independent information. For a pure binary system about 11-13 dB of SNR is required. Each time the constellation is changed to double the information rate in the same bandwidth, about 6 dB higher SNR is required.

Assuming a flawless optical path, a second ray, delayed by 5 nanoseconds or more at 20 Mbps, will be off quadrature phase by 36°. This second ray will be attenuated only about 6 dB down degrading the internal SNR accordingly. This is far less than the isolation necessary for accurate data demodulation.

The above difficulties can be mitigated by avoiding orthogonal use of two data streams in the same spectrum, and by using a modulation and demodulation where the data is defined only by the absolute amplitude and not by relative phase.

This proposal is about an analog-leaning alternative not requiring A/D or DSP. For low order modulations, the mathematical steps do not require the precision that is necessary for modulations using high order constellations.

Power Utilization

There are regulatory limits on power and power-density primarily to limit interference to incumbent users of the same spectrum for satellite and other services. For this and other reasons, maximizing the communication yield from a milliWatt of transmit power is essential. Inherently, this results from use of low order modulations using whatever bandwidth is consistent with the data rate.

There is also a common belief that more-channels-are-better leading to minimization of channel by using higher bits/Hz modulations. In fact, narrow channels with higher rates results in less capacity in Mbps per unit area. The degradation from lesser interference resistance and greater spacing between possible reuse areas defeats the gain from higher rate in the channels.

Optimization Choice

Within the scope of the technology now described, there are two criteria that give different tradeoffs:

- simplest hardware implementation, and
- better spectrum utilization by 2X with moderately increased hardware

The latter option is the subject of this proposal. There are many other criteria often invoked most of which are well met by both of these choices, specifically: power drain, out-of-band emissions and cost of implementation.

The system criteria which is infrequently invoked, but which is most important to overall system design is:

- resistance to degradation from like-signal and foreign interference

This is the optimization which minimizes susceptibility to distortion from multipath propagation, and which allows frequency reuse at minimal distance. Both of the radio solutions identified above will have low susceptibility to interference of all types.

Propagation Driven Design Considerations

The radio design is greatly influenced by the perception of multipath and time dispersion. An early experiment at 5 GHz and 10 Mbps showed that, with low directivity antennas, signals were received from all directions at a movable test point. The problem was not absence of signal but far too many of them. At 5 GHz with 1” square patch antenna, a useful amount of gain and directivity can be obtained, enough to greatly reduce the number of paths with significant energy and compensate lost level from an indirect path.

It was also found that path diversity brought missed packet rates down by a useful amount. It is well worth considering coverage from more than one access point. With more than one access point on a common channel, the coverage area may be more accurately defined with directive antenna, and the consequences of multipath greatly mitigated.

The effectiveness of a good radio can be greatly aided by a planned environment in which antenna directivity is fully utilized.

Wideband signals are usually resistant to diminished transmission in a part of the power spectrum. If most of the spectrum is transmitted, data can be received. If there is a deep fade and it occurs at a carrier frequency or in the middle of the power spectrum, there can be loss of detectability. The deep fade usually indicates a phase reversal between the energy on either side of the notch in either time or frequency. This phase reversal is disruptive to modulations in which data is phase encoded, and can be a cause of hit errors.

Avoidance of dependence on the phase of the radio signal is a property that may greatly reduce error probability. Stated positively, the information is better imposed by amplitude provided that the amplitude reference is continuously refreshed in the time span of ten's of bits.

Technology Proposed

Much of the technology is well known art, though much of it requires experience, special skills and a number of physical and software resources to execute. A few of the technologies used and described below are either novel or unusual.

There is a clean partition and interface between the radio and the associated signal processing. The processing of the data results in a video waveform of defined spectral energy density. The radio linearly translates the data-bearing baseband video waveform to the operating frequency, and back at the receiver.

Summary Description of the Modulation

There are two primary technology selections made as follows:

- Use of three-level sym-pulse baseband data shaping (to be described), and
- Use of AM suppressed carrier with ISSB (independent dual single sideband) radio modulation

The sym-pulse baseband waveform power density spectrum is positioned with the high frequency edge near 80% of the bit rate, and the low frequency edge near 10% of the high edge. The reduced level and utilization of low frequency energy is a large implementation benefit. The mathematical derivation of this pulse shape will be described in a later attachment.

The SSB signal consists of two independent data streams about a virtual carrier, and provides twice the bits/Hz as the DSB The SSB is arranged so that ***the suppressed image of one data channel is in the passband of the other.*** The phasing cancellation employed doesn't need to provide more than 20 dB suppression of the image relative to the desired signal for correct operation.

Phasing Method of Generating Single Sideband

Thr legacy phasing method requires one IQ mixer, one narrow band phase shifter and one video band phase shifter. The video band phase shifter can be produced using the Hilbert transform (experimentally confirmed). It is difficult to generate this function with the accuracy necessary for more than 20 dB worst case image rejection.

The method now employed uses two (vector) IQ mixers, and two narrow band phase shifters some what similar to the Weaver method. The sum or difference of the I and Q mixers must be combined selecting either an upper or lower sideband depending on the choice. This can be a bilateral function.

Both the radio frequency and video mixers can be shared by the two data channels, but the sum and difference combiners and all following circuits are separate data paths.

Phase Independent Amplitude Detection

A feature of the implementation is phase-independent amplitude detection for both SSB and DSB. Given the propagation-induced time-dispersion of the received signal, this method inherently provides better first-time accuracy than methods decoding both phase and amplitude to recover the data information.

The operating principal is based on the identity: $\cos^2\theta + \sin^2\theta = 1$. With a pair of quadrature phased mixers, the output will be a function of the amplitude of the input and independent of the phase of the carrier supplied to the mixers. This feature is an important part of the overall design.

Properties of the Video Waveform

The transmit wave form is generated by assembling a shift register with 5X clocking into which the binary NRZ data stream is applied. The weight of each tap is determined by resistor values. The current amplitude is the sum of all taps and associated weighting resistors. The sym-pulse wave shape is created by using the appropriate values for the weighting resistors. This will be recognized as FIR filter with a particular mathematical weighting function.

The advantages of the sym-pulse three-level random data stream is the combination of the following characteristics:

- First null near 80 % of the bit rate (rather than 100%)
- All sidelobes beyond the first null are more an 30 dB down—worst case
- Low energy content at frequencies below 10% of the first null frequency
- SSB spectral utilization of more than 1.3 bits / Hz

Because there are only three amplitude levels and no crosstalk from a quadrature phase, this modulation will be more robust than 4-QAM/QPSK and much more robust than higher order constellation modulations.

Shown below is a simulation binary data pattern and the resulting video waveform. At sampling time, the amplitude is either +1 or –1 for data 1 or approximately zero (several dB) lower for data zero.

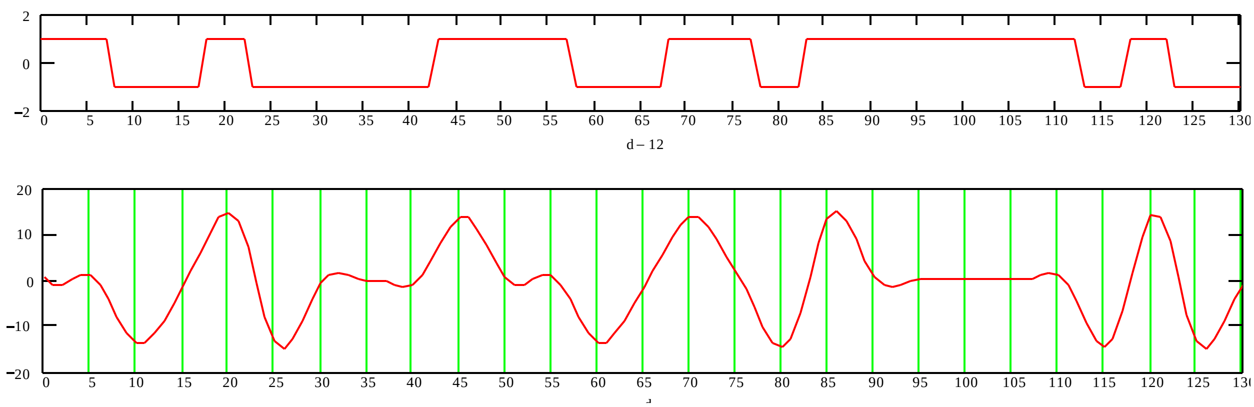


Figure 1 Data stream and associated 3-level analog waveform

Below, the associated 'eye diagram obtained by mathematical simulation is shown. The sampling instant for data amplitude determination is at the point of maximum eye opening or width.

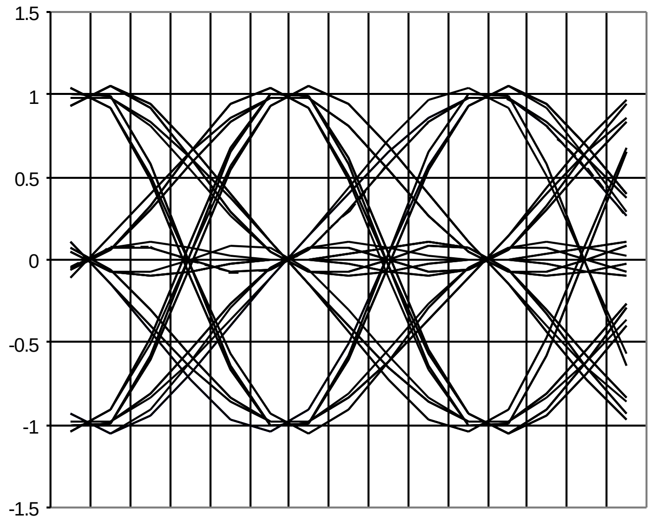


Figure 2 Simulation "eye" diagram for 3-level modulation

The power density spectrum at baseband (video) that results from this signal is shown on the following page. The first null is at 20.5 MHz with the 25 Mbps data rate. The transform is performed for a sequence of 4095 pseudo-random bits. The normal null at the bit rate also appears at 25 MHz. The out-of-band level is more than 40 dB down. The actual implementation with errors from inexact tap weights and area approximation, is not this good. Nonetheless, the potential of the method when fully exploited is shown.

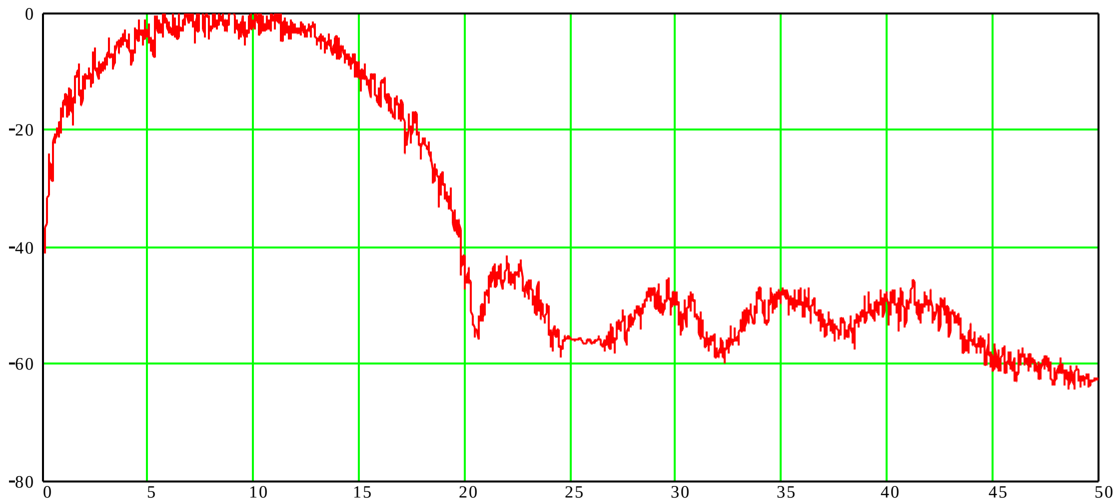


Figure 3 Calculated and smoothed spectrum for waveform of previous Fig. 1

The scale of the spectrum can be evaluated in the context of the spectrum power density masks for 802.11 and 802.15.3 as shown below. The scale of the calculated sym-pulse spectrum is that necessary for 25 Mbps. A considerable advantage in out-of band emissions can be seen particularly for the adjacent channel. ***This has been achieved not by filtering, but by generating the right pulse shape from the start.***

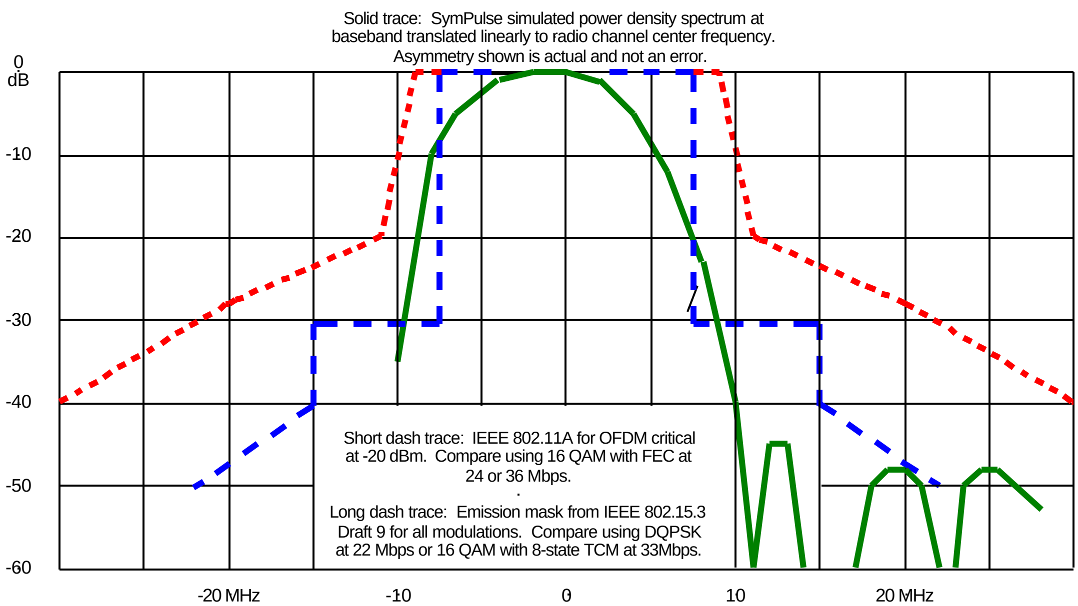


Figure 4. Spectrum of Fig.3 shown against the IEEE P802.11b and 802.15.3 masks

The solid line trace is the simulation value for the sym-pulse power density spectrum translated to the operating frequency by linear single sideband. The asymmetry shown is actual and not an error. The short dash trace is taken from IEEE 802.11A for OFDM with critical points at –20 dB down. The modulation used with this mask is 16 QAM with FEC at 24 or 36 Mbps. The long dash trace is taken for IEEE 802.15.3 Draft 9 for all modulations. The *.3 modulation is single carrier with a variety of complex modulations. This mask is used with DQPSK at 22 Mbps or16 QAM with 8-state trellis coded modulation at 33 Mbps. These are references against which sym-pulse may be compared.

Observed Properties

Some Figures showing measured results. The signaling rate was 25 Mbps for the next three figures. Shown below in Figure 5 is the measured baseband spectrum of the video baseband signal taken from the spectrum analyzer screen. Also shown is the eye diagram for that signal.

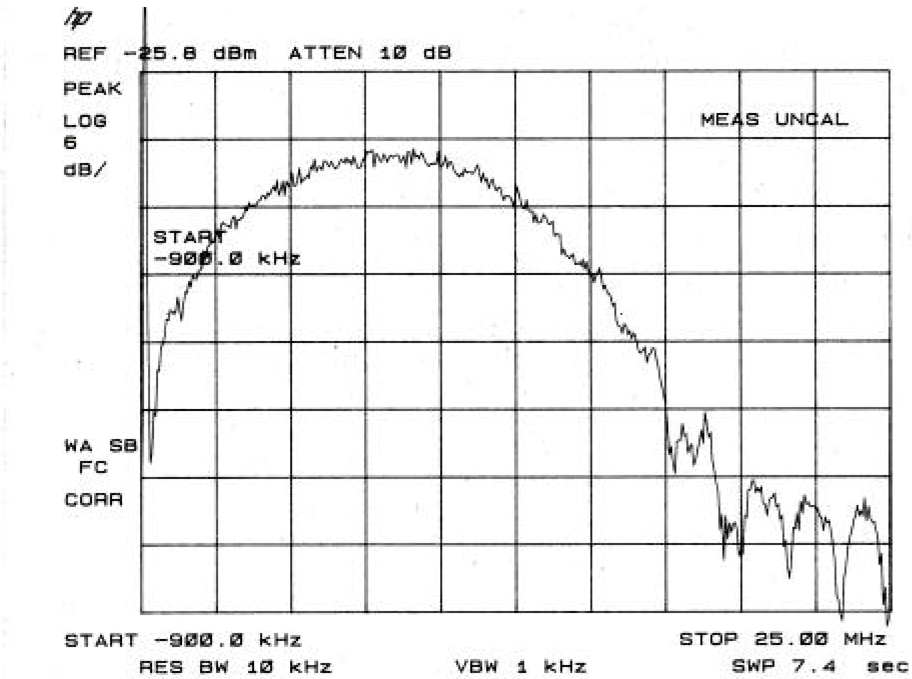


Figure 5 Measured basband spectrum for a random data stream with sym-pulse.

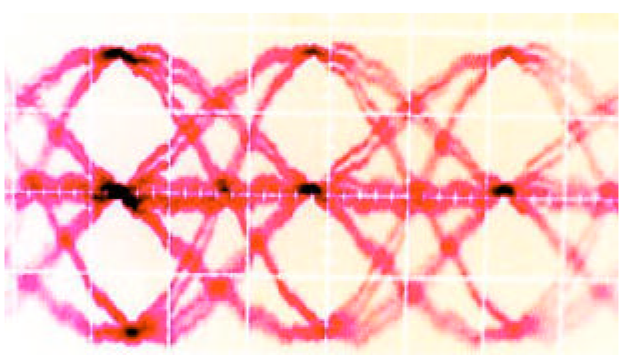


Figure 6 Measured eye diagram at baseband

A realized 5 GHz spectrum is shown in **Figure 7** below translating the 25 Mbp using a "Weaver" phasing circuit. This was an early demonstration that the linear translation could be achieved. In this figure, the image is within the signal at 30 dB down. In this proposal, the image is on the opposite side of the reference carrier but equally suppressed.

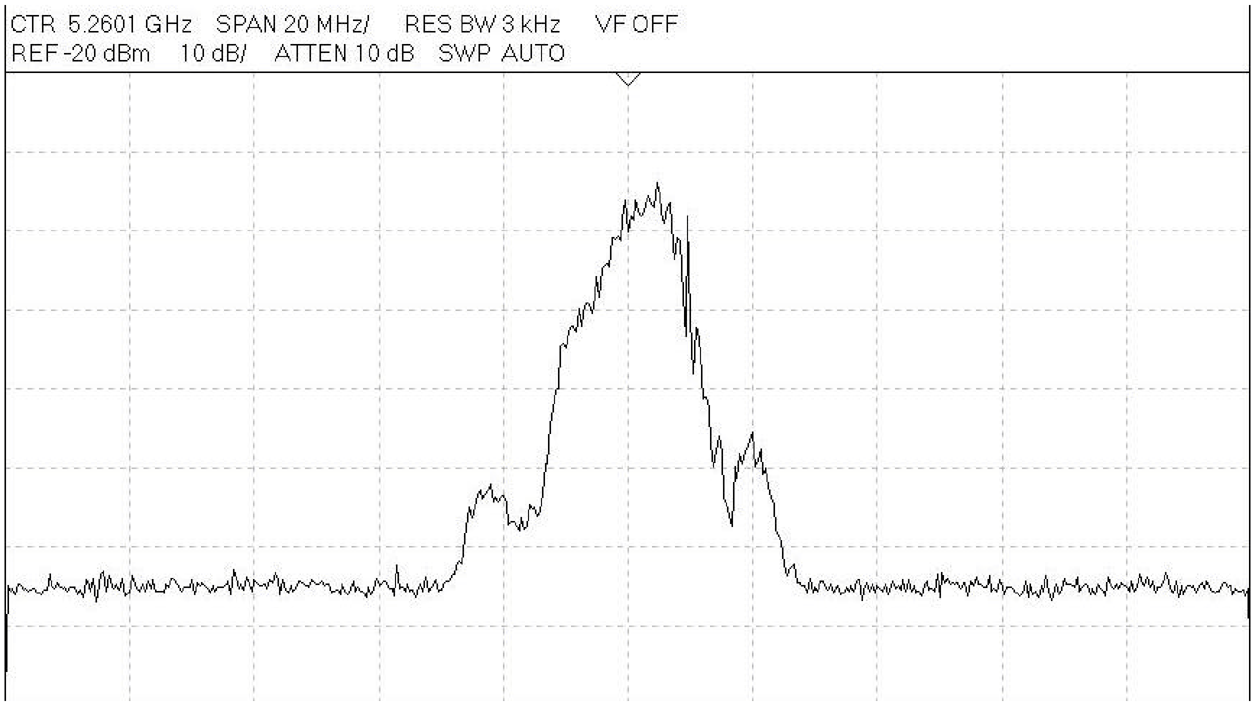


Figure 7 Sym-pulse power density spectrum after translation to 5.26 GHz

The data shown in the preceding figures is taken from earlier work, based for 25 Mbps in a 20 MHz band. Since rate is fully scalable, we have not needed to repeat many of these measurements and simulations for the newer 32 Mbps rate. We are currently developing much of this data from system rather than component simulations.

Comments and Recommendations

Development Status

Most of the work so far has been directed toward the hardware elements necessary. All of the 5 GHz amplifiers and synthesizers are available for working models. For the last few months, the focus has been on achieving system level simulation for a complete radio link. This is progressing but not complete. Within this simulation, the sym-pulse waveform generator and a few other key circuits are working. Also, integration of S-parameter defined passive components has been accomplished.

Simulation work in Progress

It is hoped that the radio simulation will have come to a point where the propagation models can be introduced, and results compared with other methods already in the simulation software. **Attachment C** is a place holder for such results at a later date.

From the present skill and knowledge base, the construction of models can go relatively quickly once the simulation has confirmed the methods used.

Need for Collaboration

A one-person or one-company proposal rarely gets as far as the short list of candidates. Provided that this work is seen as sufficiently attractive for adoption in whole or in part, this Contributor (and partner) would like to see this technology absorbed within the effort of a stronger sponsorship. There is no possibility of carrying this proposal forward without finding a stronger sponsor.

Recommendations to 802.15.3

Recognize as a class modulations which have the following properties which:

a) use a single virtual or actual radio frequency carrier, and
b) use amplitude coding to carry data information, and
c) use radio frequency phase-independent means of detecting the data carried, and
d) use a small number of levels for information coding.
e) Achieve the desired rate within a bandwidth of one set of regulatory constaints

Note: Other classes might use vector modulations of high order vector modulation, OFDM or UWB

For modulations of this class use a virtual or physical interface between video data waveform bearing information and the radio translation from baseband.

Declare a need for a baseband data waveform which has at the output of the generating means the following properties:

a) an acceptable relationship between bit rate-carried and occupied bandwidth
b) a sufficient suppression of side-lobes so that additional filter is not required

Declare a need for a radio design which is a linear translator between the video and radio frequency without material increase in spectral width.

The material presented above is intended to indicate that these are all feasible objectives whether or not either the radio or the modulation described is actually the best available solution. The reasons behind these consideration can be developed and shown in later presentation.

Acknowledgments

This work is the result of the combined efforts of Bob Ritter (partner), John Arminini and the Author. All had an indispensable and highly valuable part in this work.

Attachment A Implementation of Dual independent SSB

A possible implementation is described to support the assertion of simplicity, lower power drain and small size.

The design described in this paper is optimized for higher spectrum utilization. Another similar design using double sideband would be the choice for the simplest radio. This possibility is separately described.

The basic architectural considerations are:

- The radio band chosen for demonstration is at 5 GHz in part because of the feasibility of small directional antennas that are considered imperative for reduced effect from multipath radio propagation.
- The radio is a linear translator of a baseband video waveform bearing the digital information to and from an operating radio frequency.
- The chosen modulating signal will be generated with the necessary constraints on out-of-band energy content reducing the need for flat-time delay filtering in the radio, and with shaping suited to a bandpass medium.
- The video data signal generation and its demodulation can be accomplished entirely in high integration operating at lower frequencies.

Radio Transmit Function

The bandpass requirement has an upper limit now defined for sym-pulse at 80% of the bit rate. The lower limit is about 5-10% of the upper limit. The low frequency limit is tied to the maximum time at which the envelope does not change amplitude to an alternative state. The defined service passband is also the constrained by a limitation in variation of group delay which is now assumed to be about 20% (peak) of a bit period, and about half that value for an average across the passband. This is a requirement for handling of spectrum efficient video wave-shapes.

Our current best implementation is block diagrammed in **Figure A-2,** on a later page. The baseband waveform is converted to the form shown in **Figure 1** (preceeding section) and applied to each of two vector mixers (video frequency). A separate vector mixer is used for each sideband. The two circuits differ only in that the input signals are equal for one mixer and opposite in phase for the other. The outputs of these two mixers are combined so that each data source is coupled to both the I and Q inputs of the following shared vector mixer.

The local oscillator for the video band mixers is at the first null in the baseband spectrum. For a 30 or 32 Mbps channel this is 24-25 MHz. The difference product of this mixer causes the baseband spectrum to be inverted at the output. The main energy falls between 2 and 23 MHz.

The colors designate circuit function: transmit = red, receive = green and shared = blue.

The purpose of this seemingly avoidable set of mixers is to enable two quadrature phase shifters to be narrow-band. The alternatives available without this set of mixers require either a broadband quadrature network or a *Hilbert* transform network. What is shown is more direct than either of these alternatives.

The conversion plan may be better understood by referring to **Figure A-1** below. At the input a video signal with a power spectrum symbolized by a trapezoid (T1) is applied to a video mixer with a 25 MHz LO. The output of this mixer, shown in mid-figure (T2), is double sideband replica extending from 2-48 MHz. The lower sideband is desired, and the upper sideband is to be stopped by the 25 MHz low pass filters preceding the microwave mixer. Failure of this stop function causes an undesired product (shown dotted) to appear at the 5 GHz output at 25-50 MHz from the microwave mixer LO frequency.

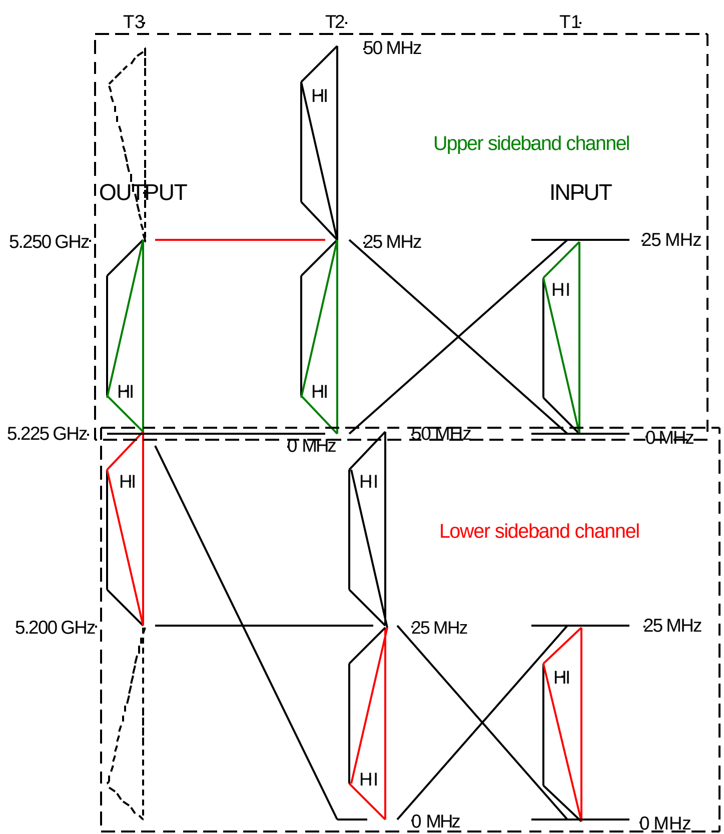


Figure A-1 Frequency relationship diagram for transmit conversion steps

The trajectory of the desired sideband is shown red color for the lower sideband and green for the upper sideband (relative to 5.25 GHz).

The filtered and inverted sideband from the video mixer is applied to the input of the microwave mixer. The single sideband selection causes one of the data signals to become the upper sideband and the other to become the lower sideband. As transmitted, the higher frequency part of the spectrum is closest to the microwave carrier.

There is not intentional radiation in the 4 MHz gap between the two active power spectrums.

Returning to **Figure A-2**, switches that are shown are used to select transmit and receive mode alternately. This diagram shows only the transmit mode position associated circuit functions.

The inverted data spectrum is passed through the same lowpass filters used for the receiver. In the transmitter these may be simples *Bessel* or linear phase filters, because the stop band is not required to reduce the out-of-band energy of the waveform. The spectrum is good enough before filtering at close-in frequencies and the filter does help considerable an octave above the LO.

The filtered and inverted data spectrum is applied to the microwave vector mixers with an LO at the radio operating frequency of 5.25 GHz. The sidebands are positioned on both sides of this reference frequency with small but useful gap in the middle. The two sidebands are carrying independent information. The image of each falls in desired sideband of the other. This image can be down by 40 dB, but as a design assumption, 26 dB is sufficient for negligible data degradation

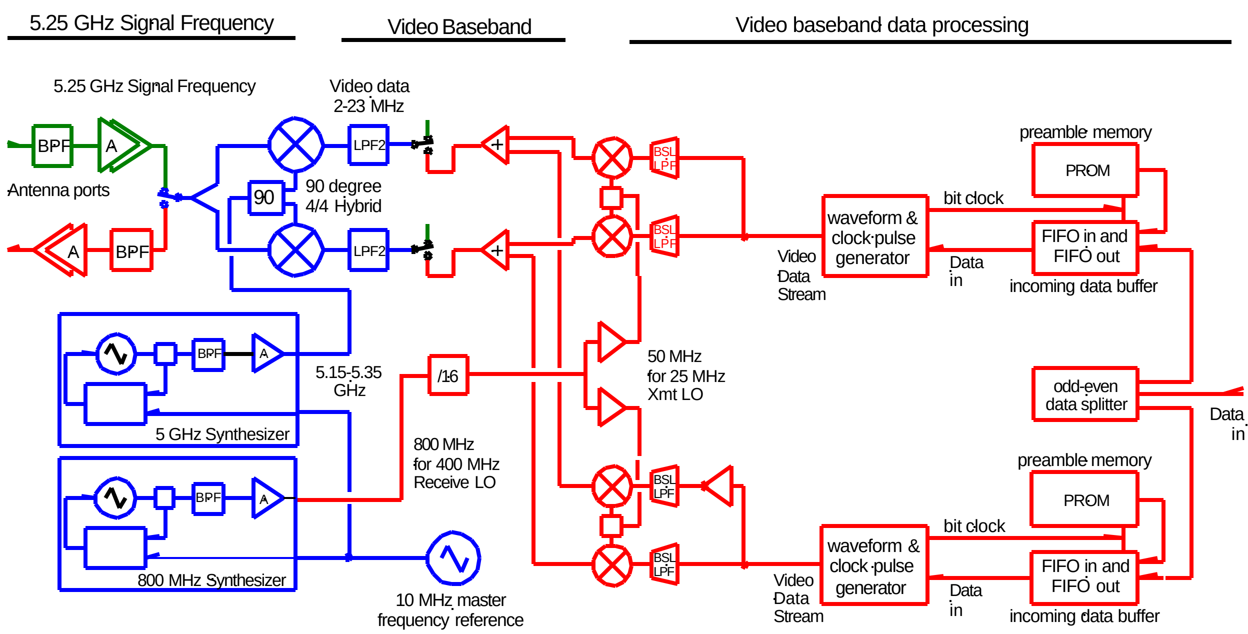


Figure A-2 5.25 GHz dual ISSB AM data radio transmitter block diagram (60 Mbps in 48 MHz bandwidth)

The 5 GHz amplification after that point is routine.

This block diagram can be merged with that of **Figure A-3** to show a complete transceiver. The blue parts are common to both.

Radio Receiver Function

The radio receiver is complementary to and can share many of the same blocks as the radio transmitter.

Some special considerations apply to the radio receiver that result from positioning the channel selectivity as early as possible in the amplifier chain. The conversion scheme chosen keeps leaked LO frequencies from being within the data passband. As shown in the block diagram in **Figure A-3**, three sets of vector mixers are used. This enables the main amplification to take place at an intermediate frequency of 400 MHz.

The three mixers are designated as: 1) microwave, 2) IF (intermediate frequency) and 3) video. All of them in fact are involved with two of the three frequencies. The video frequency appears twice; first between the microwave and IF mixers and again at the output of video mixer.

Like the transmitter, the numeric values are for two channels of 30 or 32 Mbps in a bandwidth of 48 MHz.

Signal Frequency Circuits

The microwave mixer is critical for gain and phase balance of the injected LO. Fortunately, narrowband phase shift can be obtained with high accuracy from classic coupler design. Artful shielding, difficult to simulate, was found necessary to contain LO leakage into surrounding circuits. This part of the circuit is routine except for the precision execution required.

The gain and noise figure of the LNA is important. The LNA must have sufficient gain so that the following tandem circuit loss does not prevent the signal from being well above the local noise levels.

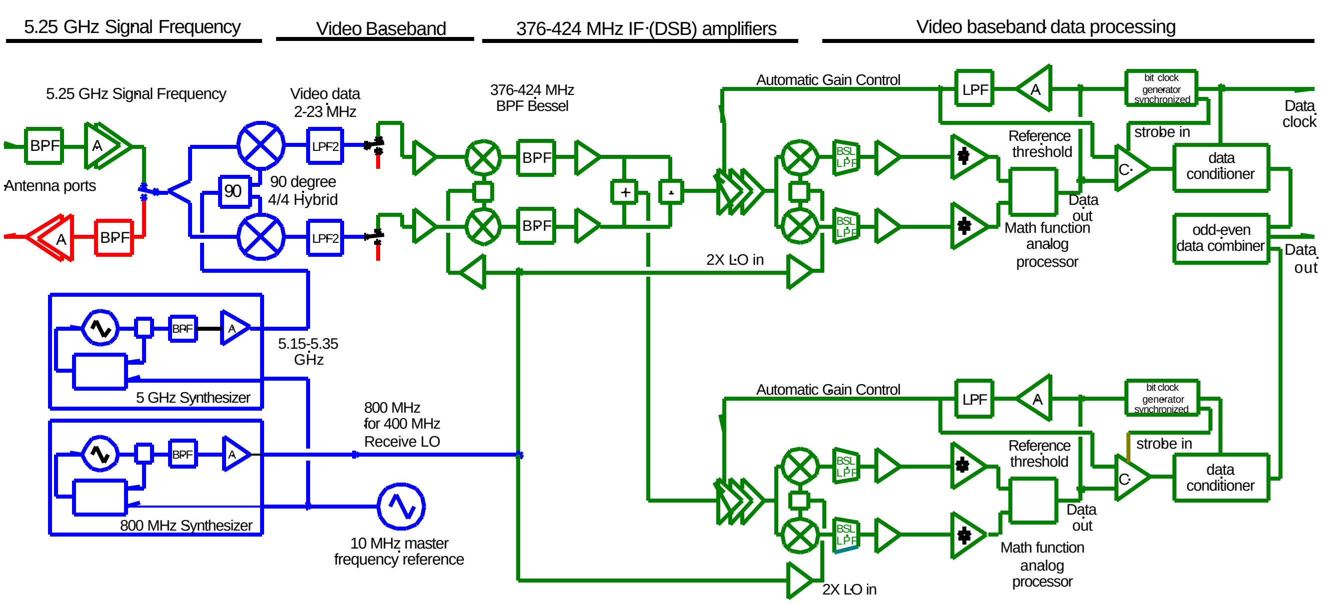


Figure A-3 5.25 GHz dual ISSB AM radio receiver block diagram

Low Level Video Baseband

The output of the microwave vector mixer is passed forward through lowpass filters. These filters determine the adjacent channel rejection. The actual design employed is a flat group delay lowpass, a bandstop and an equalizer in tandem. Fortunately, the requirements on group delay are less severe at the sides of the pass band because of the lower power densities there. The amplifier shown after the LPF will provides a small amount of gain, but it is equally important to provide an accurate termination for the filter.

One LO is used for both the IF mixer and the following video mixer. The LO frequency in this design is at 400 MHz. If the LO for the IF mixer was at 25 MHz, the original modulating signal would be recovered, but the level would be to low for demodulation. As shown, the same baseband signal mixed with 400 MHz provides a single DSB version of the baseband signal at the intermediate frequency.

Intermediate Frequency Circuits

Flat-delay bandpass filters are used to limit noise bandwidth and extraneous signals at the output of the IF mixer. A further amplifier is used to provide filter termination and a defined source impedance for the passive adder and difference circuits. In combination with the quadrature phase shifts in the two preceding LO’s, the adder and difference circuits select the upper and lower sideband of the 5.25 GHz signal.

The IF amplifier is similar to types commonly used in pocket telephones which mostly easily accommodate a 58 MHz wide signal. A variable gain control is necessary at this point.

Phase Independent Demodulation

The output of the IF amplifier is applied to video mixer using the same LO as the preceding IF mixer. The exact frequency of the 400 MHz LO is unimportant since it is added and then subtracted. This mixer step recovers the baseband signal as it exists coming out the IF mixer.

This last demodulation step recovers the baseband video waveform as an amplitude envelope. The effect of frequency mismatch between the local and received reference frequency is an amplitude precession at the output of both the I and Q video mixers.

The two precessing outputs are a sine and a cosine time-shaped amplitude with respect to the phase of the 400 MHz LO. The sum of the squares of each of these is constant dependent only on the amplitude of the signals and independent of the oscillator phase. The operation is based on the trigonometric identity of $\sin^2\theta + \cos^2\theta = 1$. This method enables recovery of amplitude of the double sideband signal (with or without supression of the carrier) without requiring the local oscillators to be phase or frequency locked to match the source LO. This method also works with one sideband. Once the amplitude envelope is recovered, the detection of the data message can start.

Data Detection

A good, but not optimum, method for recovering the data is to strobe at or near the center of the eye opening, and then to make the bit-value decision that the absolute amplitude is above or below a single threshold. This requires local bit-clock synchronization before demodulation can take place.

It is assumed that the rate of the bit-clock is known in advance, and only the phase need be adjusted upon acquisition. This is done by designing the burst preamble to use 1010… bit pattern which results in a near sine wave at a sub-multiple of the bit rate. The synchronization is accomplished by immediate initialization of a counter at 16 or 32 X the bit clock rate, and then refinement by dropping or adding a count. Once synchronized the clock runs open loop for the duration of the burst.

During the acquisition preamble the setting of the AGC takes place, and the peak value of incoming envelope amplitude is determined. The decision level of the strobed comparator is then set at a percentage of the peak level (nominally 80% or –4 dB). It is possible to have the threshold track the received level during the packet transfer or freeze it after acquisition.

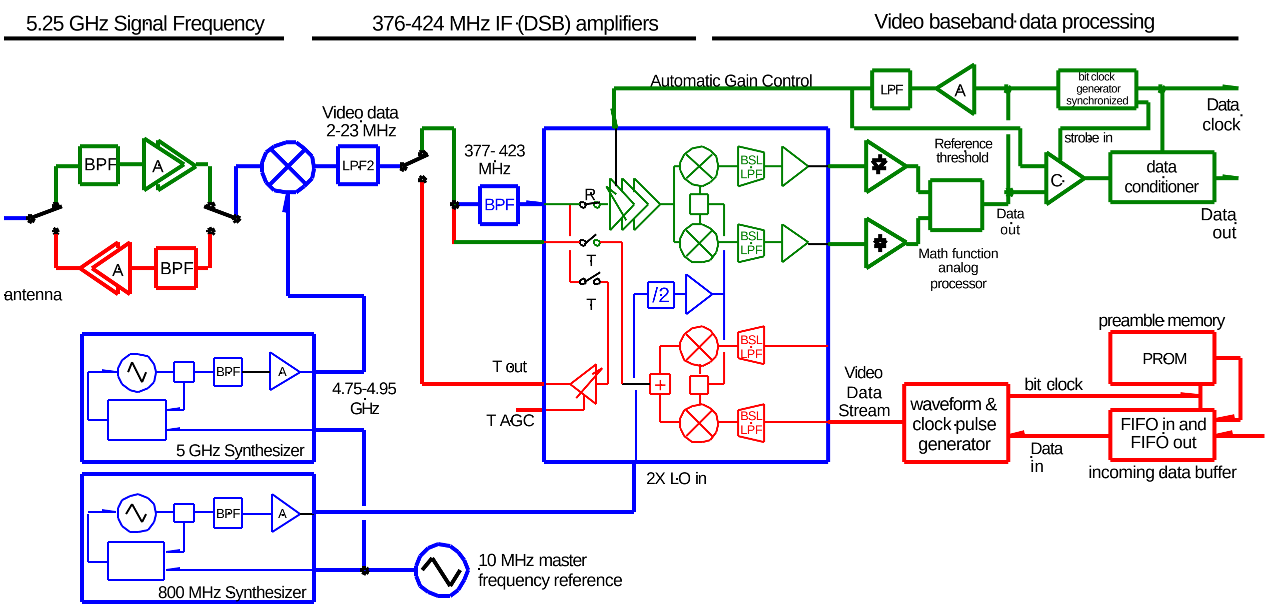
The received bit value and the clock can be delivered to the following data stream processors where delimiting and unscrambling is implemented.

The block diagram of **Figure 11** is an approximate representation of this process.

Attachment B Implementation of a DSBSC Data Radio

The simplest possible radio and data detection can be achieved using the same video baseband waveform as described in **Attachment A** with ***double sideband suppressed carrier amplitude modulation***. The power in the two sidebands is recovered in the phase-independent demodulation.

This design will provide 30-32 Mbps in 24-25 MHz or any other rate at linearly scaled bandwidth. Higher rates may stress our presently known methods of signal processing that would be used in demonstration models, but this would not be true for other development organizations with greater capability.



Shown below is a bock diagram of this type of radio.

Figure B-1 AM DSBSC Radio Modem Transceiver Block Diagram

It is believed that this modulation method will be particularly good at resisting all types of interference including that from the time dispersion of multipath propagation. It is also sufficiently simple that diversity dual channel versions could be small and cost effective.

Attachment C—System Simulation Results (placeholder)

When available, system simulation results will be placed here in future versions of this document.