0-Hz-IF FSK/AM SUB-CARRIER DEMODULATOR on a 6U-VME-CARD

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ABSTRACT

Aerospace Report No. TOR-0059(6110-01)-3, section 1.3.3 outlines the design and performance requirements of SGLS (Space Ground Link Subsystem) services. GDP Space Systems has developed a single card slot FSK (Frequency Shift Keying)/AM (Amplitude Modulation) demodulator. An application of this service is the US Air Force Satellite Command and Ranging System. The SGLS signal is tri-tone-FSK, amplitude modulated by a modified triangle wave at half the data rate.

First generation FSK/AM demodulators had poor noise performance because the signal tones were filtered and processed at IF frequencies (65, 76 and 95 kHz). Second generation demodulators suffer from "threshold" due to non-linear devices in the signal path before the primary noise filtering. The GDP Space Systems demodulator uses a 0-Hz-IF topology and avoids both of these shortcomings. In this approach, the signal is first non-coherently down converted to baseband by linear devices, then it is filtered and processed. This paper will discuss the GDP 0-Hz-IF FSK/AM (SGLS) demodulator.

KEY WORDS

SGLS, FSK/AM, Tri-Tone FSK, Non-Coherent Down-Conversion, 0-Hz-IF, Threshold

INTRODUCTION

SGLS has been around since the early 1960's. The first generation SGLS demodulators (Figure 1) did all their filtering and signal processing at IF. This approach – minimal in hardware – exhibited poor performance even in moderate noise. The second generation SGLS demodulators (Figure 2) used non-linear devices (typically rectifiers) for down converting the signal from IF to baseband where the signal was filtered and processed. These non-linear devices worked well in moderate noise but exhibited threshold (and the resultant poor performance) in noisy (Eb/No < 10 dB) environments. The 0-Hz-IF

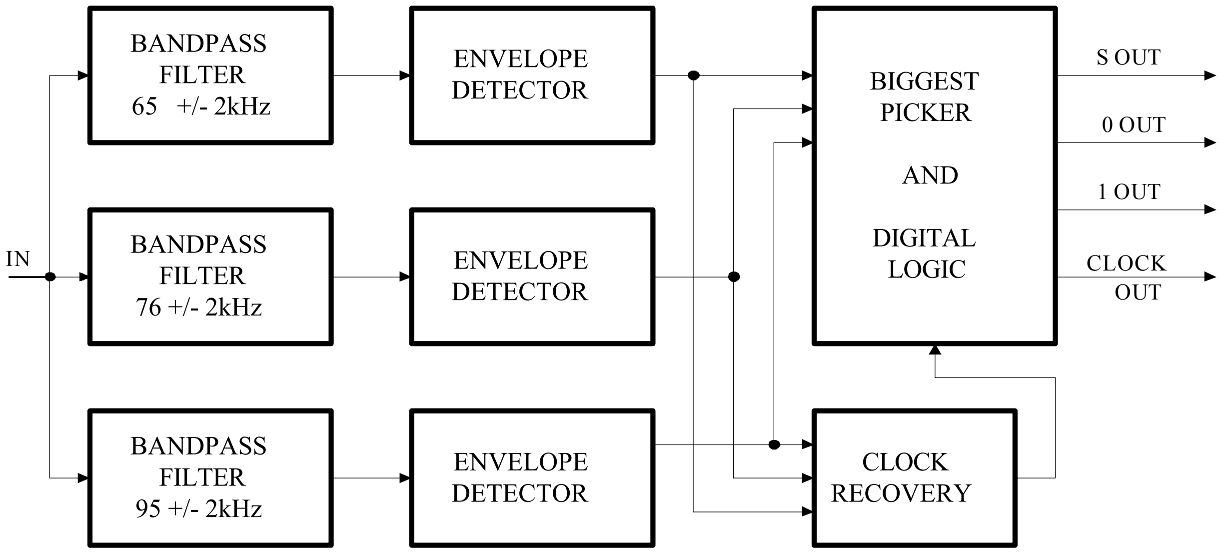

FIGURE 1 - FIRST GENERATION DEMODULATOR

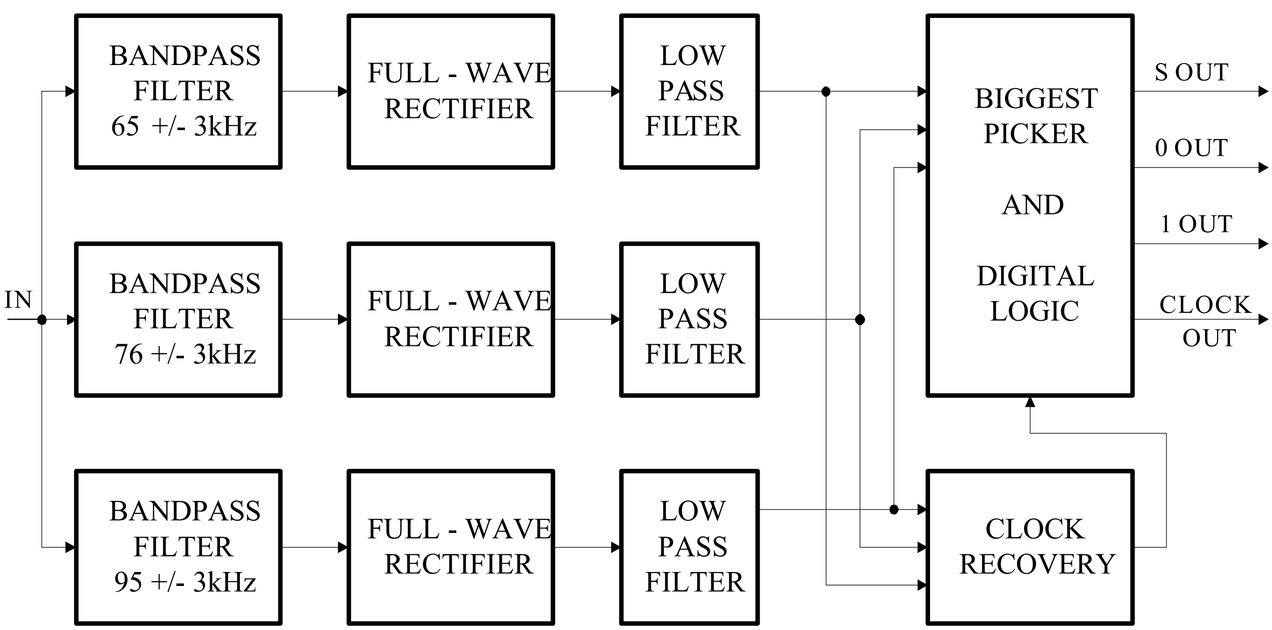

FIGURE 2 - SECOND GENERATION DEMODULATOR

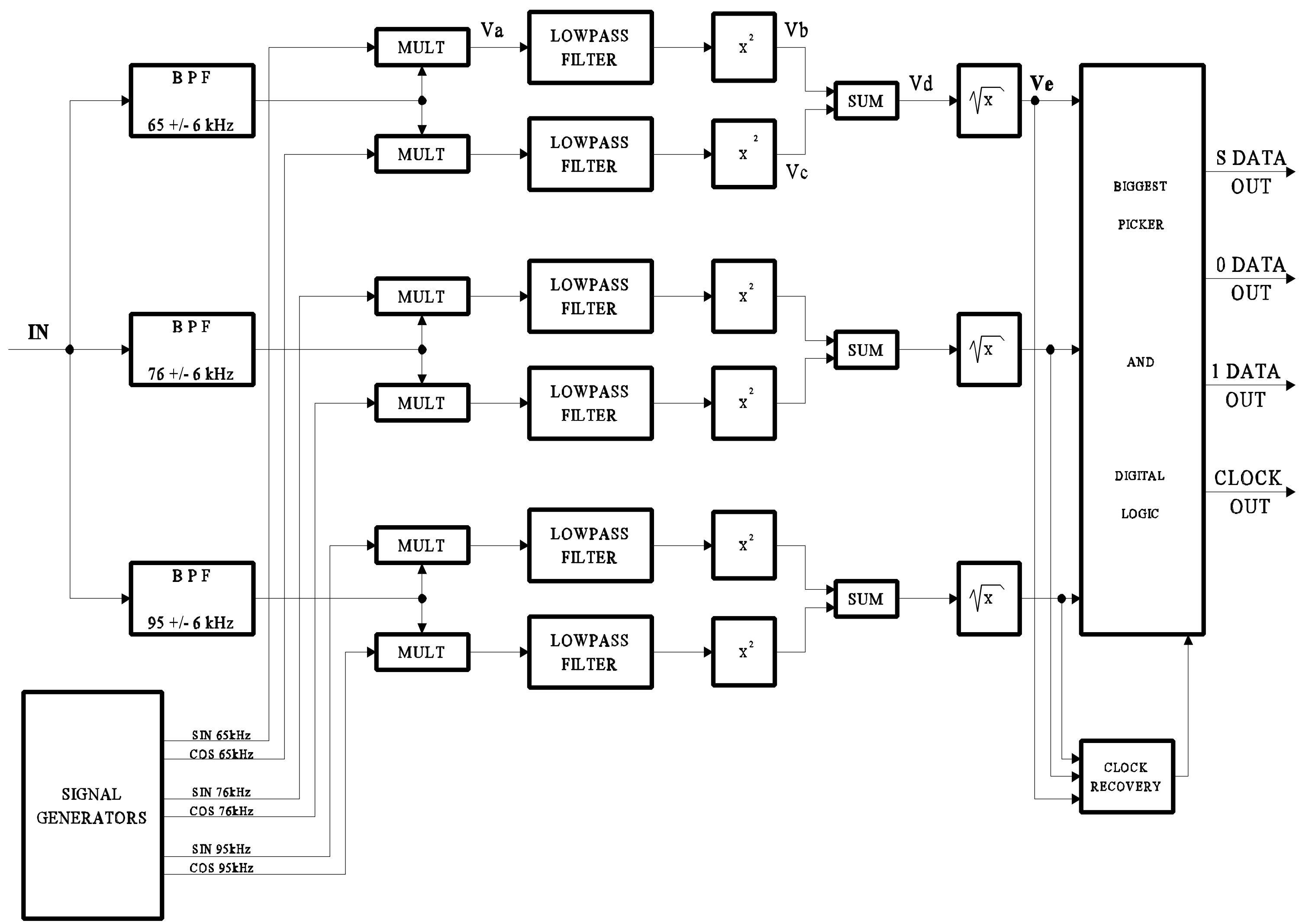

FIGURE 3 0 - Hz - IF DEMODULATOR

demodulator (Figure 3), described in this paper, uses linear devices for down conversion; therefore, it does not suffer from threshold and the resultant degradation of performance at moderate to low Eb/No's. However, this demodulator requires somewhat more hardware than either of the two earlier demodulators. The 0-Hz-IF topology, similar to one currently used in FSK paging demodulators, was considered to be too complex until the advent of integrated circuits possessing a high level of functionality. A requirement to process two data rates lessens the difference in hardware requirements between the 0-Hz-IF and second generation demodulators.

GENERAL DEMODULATOR THEORY

The performance of any demodulator is heavily dependant on its noise filtering. In a sampled system, which most digital data systems are, the optimal noise filter has a frequency transfer response that matches the input signal's frequency spectrum. When the signal of interest is run through this optimal filter, the time domain response is an "integrate and dump" waveform – the filter's output takes exactly the entire bit period to ramp to its final value. NRZ data's frequency spectrum is sin(x)/x and the optimal filter must have that frequency response.

An ideal demodulator can filter its input signal at either IF or baseband with equal levels of performance. However, the filter at IF requires twice the hardware of a baseband filter. The sin(x)/x frequency spectrum has a peak at dc, nulls spaced at odd integer multiples of the data rate and peaks of decreasing amplitude at even integers multiples of the data rate. When a baseband sin(x)/x signal is used to modulate an IF carrier (F_c), this same frequency spectrum is now translated up by F_c Hz. In addition, a mirror image of the spectrum also starts at Fc but continues downward toward 0 Hz. The complexity of the IF signal is twice that of its non-modulated baseband signal. For sin(x)/x signals, the baseband filter requires imaginary axis zeroes at odd integer multiples of the bit rate, while the IF filter requires its zeros at the carrier frequency plus AND MINUS odd integer multiples of the bit rate – an equivalent IF filter requires twice the number of zeroes (and attendant hardware) are required. Another disadvantage of IF filtering is that component tolerances affect the filter response as a percentage of carrier frequency, while at baseband the error is a percentage of the bit rate – a much smaller value.

The SGLS signal has an inherent disadvantage in noise performance. Its SNR (Signal to Noise Ratio) varies from one bit period to the next. This is due to the timing relation between the AM-half-clock-rate-triangle-wave and the FSK data. The data (and FSK frequency) transition occurs one-tenth of a bit period before the triangle wave's maximum or minimum. The energy (and SNR) in a data period that started before an AM minimum is about two dB less than a data period that stated before a maximum; however, the noise power is equal in both bit periods. Therefore, SGLS demodulators start with a noise performance disadvantage even with optimum filtering.

HISTORY

First generation SGLS demodulators used minimal hardware and offered minimal noise performance. They used three bandpass filters as the principle noise filters with envelope detectors at IF for data and clock recovery (Figure 1, 1 kbps data). In theory, the bandpass filters could be made close to optimum for best noise performance; however, optimal filtering at IF requires complex hardware and a labor-intensive alignment. In practice, the bandpass filters were usually just good enough to eliminate grossly out-of-band noise while not excessively corrupting the data. Filter bandwidths were typically two to four times bit rate. These demodulators started off with about a five dB disadvantage (from optimum) due to wide bandwidth filters and the resultant excess noise in the bit detection/decision process. Because of the miserable noise performance, almost all first generation SGLS demodulators have been relegated to the scrap heap.

Second generation SGLS demodulators (Figure 2, 1 kbps data) first down convert the incoming signal to base band and then filter it. They use non-linear devices (typically full or half wave rectifiers) for down conversion. The output of each rectifier is a baseband signal which is then low pass filtered. It is these low pass filters that serve as the principal noise filters. While this approach allows better filtering with less filter hardware than an optimum bandpass filtered demodulator, it does have a significant short coming. The rectifiers (as in all non-linear devices) exhibit threshold. Threshold occurs when, for a given decrease in SNR at the input of a device, the SNR at the output decreases by a significantly larger amount. Threshold in rectifiers occurs at an SNR of approximately five dB [1] and in a noisy environment, this demodulator would have poor performance. Typically, the bandwidth of the bandpass filters is four to six times the bit rate (a larger bandwidth than the first generation demodulator) to avoid corrupting the data prior to down conversion and low pass filtering. With the bandpass filtering that wide, threshold in the rectifiers begins to occur at an Eb/No of between ten and fifteen dB. An obvious solution is to use narrower bandpass filters, but there is a limit and more accurate (and complex) bandpass filters would be required.

The 0-Hz-IF demodulator uses multipliers, not rectifiers, for non-coherent down conversion and does not suffer from threshold. The 0-Hz-IF as well as the first and second generation demodulators use similar "biggest pickers" (for data bit decision) and clock recovery circuits. The biggest pickers and clock recovery circuits are not discussed in this paper.

0-HZ-IF SGLS DEMODULATOR THEORY OF OPERATION

The General Data Products FDM001 is a SGLS FSK/AM Demodulator on a 6U-VME card. It auto-rate-detects between two data rates, one or two kbps. For the discussion of signal flow, refer to Figure 3. See the appendix for the mathematical derivation of the down conversion.

The input signal is first run through three bandpass filters. These filters serve only to limit the out-of-band noise in the incoming signal. This ensures that the multipliers are operating in their linear region. In the FDM001, these filters have a bandwidth of 12 kHz which is large enough to pass the main lobe and most of the first side lobes of a two kbps signal. The bandpass filter’s bandwidth is not critical and needs only to be wide enough to pass the signal of interest without distortion, yet narrow enough to limit out-of-band noise.

Each of the three bandpass filtered tone signals is non-coherently down converted to baseband. This is done by splitting each tone in two and multiplying by two quadrature oscillators running at that tone's nominal frequency. The result of each multiplication is low pass filtered to eliminate both the noise and the sum-frequency term. The filtered terms are squared, summed and then square-rooted. This final square-rooted result is the reconstructed baseband signal.

The 0-Hz-IF topology is only moderately demanding in its hardware implementation. The quadrature oscillators need not be running at exactly the same frequency as the incoming tone, but the frequency offset must be small compared to the data rate. The crudest crystal oscillators (100 ppm) easily meet this offset requirement which also includes allowances for the input signal's frequency uncertainty (100 ppm). The multipliers are running in their linear region and do not suffer from threshold. The multipliers only need to have sufficient signal overhead to stay in their linear region when high levels of noise are present. The squaring and square-rooting devices after the low pass filters are non-linear and subject to threshold; however, at this point the signal has been maximally filtered yielding the lowest noise level possible, so threshold is not a problem.

The low pass filters are the principal noise filters. While an ordinary low pass filter would suffice, the FDM001 card uses filters spectrally “matched” to the data. These matched filters' time domain response is an integrate and dump waveform which offers a performance gain of about one dB compared to equivalent non-matched low pass filters. A matched filter is much easier to implement at baseband than at IF for the reasons previously stated.

CONCLUSIONS

The 0-Hz-IF demodulator has several advantages over previous generations of FSK demodulators. Primary noise filtering is done at baseband which yields filtering closer to optimal and with less hardware than filtering done at IF. This gives a significant improvement over first generation demodulators. The lack of threshold (and resultant better performance in heavy noise) is the primary advantage of a 0-Hz-IF demodulator over a second generation demodulator. The cost of the improved performance is an increase in hardware complexity. Hardware complexity has less of an impact now than in the past due to the availability of higher functionality IC's. The second generation demodulator's threshold problem requires a good bit of care and complexity in the bandpass filter design for just adequate noise performance. A second-generation, dual data rate demodulator with reasonable noise performance requires two sets of moderately complex bandpass filters to avoid threshold in the rectifiers. The FDM001 (a 0-Hz-IF, dual data rate demodulator) requires only one set of crude bandpass filters, somewhat mitigating the increase in hardware compared to a second generation demodulator, yet has superior noise performance.

REFERENCE

1. Taub, Herbert and Schilling, Donald, "SQUARE-LAW DEMODULATOR," Principles of Communication Systems, Magraw-Hill, 1971, p.287.

The proof in this reference is for a squaring device. A squaring device can be thought of as a rectifier working with only the fundamental term as its inputs. The threshold level of a rectifier is within one dB of the value given for a squaring device.

APPENDIX

Mathematical Proof of Non-Coherent Down Conversion

Each of the input FSK tones can be represented as $M_n(t)\sin(\omega_n t)$. $M_n(t)$ is an on-off switching function (having a value of either 1 or 0) and signifies whether the given tone is present or not. ω_n is the radian frequency of the specific tone. The subscript n has a value of 1, 2 or 3, and respectively refers to the 65, 76 or 95 kHz tone ($\times 2\pi$ is implied for the rest of the appendix). Only one M_n can be 1 at any time – there is only one tone present at any given time.

Refer to Figure 3. The signal path ending with V_e is used. The input signal is 65 kHz + Δf (M_1 =1). The n subscript 1 is implied and not shown for both M and ω. ω_1 is equal to 65 kHz. The signal at the output of the multiplier, V_a, is:

$$\underbrace{M(t)\sin(\omega t + \Delta f)}_{\text{Input signal}} \times \underbrace{\sin(\omega t)}_{\text{Local Osc.}} = M(t)\{-\cos(2\,\omega t + \Delta f) + \cos(-\Delta f)\}.$$

The low pass filter eliminates the sum-frequency term $(\cos(2\,\omega t + \Delta f)$. M(t) is modified by the low pass filter but for now will still be referred to as M(t). Δf is small and the low pass filter has no effect on $\cos(-\Delta f)$. The output of the low pass filter, $M(t)\cos(-\Delta f)$, is squared and at V_b yields:

$$[M(t)]^2 \cos^2 (\Delta f).$$

By similar logic, the signal at V_c is:

$$[M(t)]^2 \sin^2 (\Delta f).$$

The sum of the two squared terms, V_d, is:

$$[M(t)]^2 \cos^2 (\Delta f) + [M(t)]^2 \sin^2 (\Delta f) \qquad = \qquad [M(t)]^2.$$

The square root of V_d is M(t), which has a value of either 0 or 1 depending on whether the tone is present or not. This is a recreation of the original data.

Modification of M(t) by the low pass filter was mentioned earlier in the appendix. The M(t) which occurs at V_e is equivalent to NRZ data that has been run through low-pass/matched filters. The signal at V_e is an integrate and dump waveform and is optimal for noise performance in a sampled data system.