Filter Summary
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ABSTRACT: DRAFT Undergoing Revision – Still subject to corrections

This paper is in two parts. The first part contains a description of the conventional filter, which includes all filters in general use today. These filters when utilizing LC components date from the turn of the century, and from the 1920s using crystals, with improved analyses made later by Nyquist and others. There are some facts stated here which are not generally considered in all texts, or probably overlooked in communications work. The second part covers the special zero rise time and near zero group delay RF filters for Ultra Narrow Band methods.

HOW CONVENTIONAL FILTERS WORK:
Conventional filters cannot be used with Ultra Narrow Band Method, which are absolutely dependent upon Negative or Zero group delay filters.

THE OPTIMUM FILTER:
The integrating Filter:
The optimum filter is described as "the filter that passes the most signal power with the least noise power". The integrating filter used as part of a “correlator”** is considered to be a matched filter.

Fig. 1.
The integrating filter used after a detector, or as a bandpass filter, is shown in Fig. 1. The circuit** consists of a preceding mixer plus the integrator and a sample and hold circuit following the integrator to separate the noise and data. In this case, the integrator RC time is optimized for the group delay = Ts. (Symbol rise time, which is equal to the bit period for 2 level modulation).

** Correlator circuit notes pages 2 and 39. **
Using the post detection amplitude data pattern of ones and zeros at (A) as an input, the integrator charges positively as shown in (B) until it is sampled by S2 at its peak (C). The capacitor is then discharged by S1, to be recharged anew by the input signal at the end of a bit or symbol period, or simply left to obtain a new level from the incoming signal. The amplifier can be a sample and hold circuit as shown, or merely a clipping amplifier with a rectangular output obtained when the signal rises or falls above or below the center line seen in (B).

A conventional crystal or LC filter has the same integrating effect as the RC integrator in Fig. 1. Instead of the RC rise time responding to level, there is an equivalent group delay time, or phase slew rate. The incoming signal is a sine wave at the resonant frequency, which is varying in amplitude or phase. The filter response preserves the sine waves (ringing), while reducing the effects of noise disturbances due to the time delay.

![Figure 1A. The rise time of a filter with positive group delay responding to a burst pulse. This filter has a rise time of 200 nanoseconds and a Nyquist bandwidth B of 5 MHz.](image)

The group delay (rise time) for conventional filters is traditionally calculated to be:

\[ T_g = \frac{\Delta \phi}{(2\pi \Delta f)} \]  
Eq. 1. Derived from \( \omega t = \Phi \).

For LC or Gaussian filters, this is:

\[ T_g = \frac{1}{(\Delta f)} \quad \text{or} \quad T_g = \frac{Q}{[IF]} \]

Where IF is the filter center freq.

\[ T_g = \frac{Q}{[4IF]} \]

is the phase slew time for 90 degrees. For 180 degrees it is \( T_g = \frac{Q}{[2IF]} \). The basic formula is for \( 2\pi \) radians, where \( T_g = \frac{Q}{[IF]} \).

Obviously, a very narrow \( \Delta f \) or high Q bandwidth filter has a very large group delay unless \( \Delta \phi = 0 \), or \( f \) is infinite. A network analyzer will show \( T_g = \frac{Q}{[IF]} \).

\[ \Delta f = BW = B \]

\[ (\Delta f)t = \frac{\Delta \phi}{2\pi} = BT \]

\[ BT = \frac{2\pi}{2\pi} = 1 \]

There is an associated equation for the equivalent RC rise time of the conventional LC or crystal filter: \( T_r = 0.7/B \), where B is the 3 dB bandwidth \( \Delta f \) of the filter.
This is the time from 10% to 90% on the RC curve. Bandwidth, rise time and sampling rate are mathematically linked. The general custom in analysis is to assume $T_r = 1/B$ and that there is an associated slew rate of 360 degrees during $T_r = 1/B$. ($BT_r = 1$).

For the amplitude relationship and noise bandwidth, the crystal $Q$ is $= \text{Freq}/(3\text{dB BW})$, which is from 20,000 to 50,000, therefore $B = IF/Q$ and $BT_r = 1$, so $1/T = IF/Q$, or $T = Q/IF$. $T$ is normal group delay, not the transient vector sum delay.

The rise time and sampling period are related --- $T_r = 1/B$, where $B$ becomes the necessary sampling rate, as in $B = 1/T_r$, which is optimized to match the signal peaks. Most engineers associate $B$ with the filter bandwidth $\Delta f$, and use it as such in the Shannon Channel Capacity equation. This can lead to serious errors, since it is not the bandwidth of the RF filter used, except in the optimum case for the correlator with integration (matched filter). This must be the Nyquist bandwidth $B$, since it relates to Nyquist's sampling theorem:

"You must sample at the symbol rate, or at frequency higher than the symbol rate ".

All conventional digital communication takes place in the form of amplitude, frequency or phase changing pulses, usually rectangular pulses, which are altered by filtering. As seen in the integrating filter of Fig. 1, each pulse has a rise time using conventional filters, having a duration '$\tau$' ($T_r$), and an associated optimum repetition rate $B$, optimized at $B = 1/\tau$. Conventional filters are integrators. The rise time is also associated with a fixed filter bandwidth $= 1/\tau$, which is optimized at 1 pulse or bit period. B is referred to as the Nyquist bandwidth, which must also equal to the sampling rate.

**The correlator, consisting of a detector plus an integrating filter, is considered to be an “optimum filter” in the presence of white noise. The maximum signal power is obtained by integrating the incoming signal pulse in the filter. The noise is white and has a long term integrated output level at 0 volts. The short term signal information will have a positive or negative integrated value.

The input signal pulse here is considered to be rectangular, but other pulse shapes apply as well if the sample time is properly chosen.

The matched filter is best described as the best filter that is usable for the modulation method employed. It may or may not be the optimum filter, since the optimum filter could mask some modulation details. Generally it is the filter that results in the best SNR.

All conventional filters, LC, Crystal and SAW function on a similar principal. It is all a matter of phase shift $\Delta \phi$ and rise time $T_r$ through the circuit bandwidth.

A conventional filter with rise time = bit period is a form of integrating filter. Normally, the signals are sampled at the minimum Nyquist sampling rate, which is equal to two samples per bit at the baseband frequency $f_m$, which is $1/2$ of the
data bit rate $f_b$. Thus the minimum Nyquist sampling rate $W$, and the minimum bandwidth $B$ with conventional filters, are equal to the bit rate $= 1/\tau$, and both are tied to the filter rise time $\tau$.

The optimum filter is described as "the filter that passes the most signal power with the least noise power". Obviously a filter with a group delay longer than the bit period does not fit this definition, since low frequency noise when $T_g > T_b$, comes through at full strength, while the signal is reduced by the ratio $T_b / T_g$.

The closest approximation to a brick wall filter, (which does not exist in practice) is a Nyquist filter (Raised Cosine) with $\alpha = 0.0$. Simpler filters such as the 2 pole crystal filter seen in Fig.3, yield crude approximations. A series resonant crystal at the top in Fig. 3 is shunted by a parallel resonant crystal 'B'. The two operate at slightly different frequencies to yield the frequency response seen at the right. The phase shift from edge to edge is 180 degrees, so there is a resemblance to the ideal filter. The group delay is calculated in the same manner from Eq. 1. The bandpass is determined by the Q of the crystals.

Fig. 2. Ideal, or Brick Wall Filter. The ideal, or brick wall filter, is shown in Fig. 2. This filter has a phase shift $\Delta \Phi$ from edge to edge of 180 degrees, or $\pi$ radians over a frequency shift $\Delta f$. Normally, $\Delta \Phi$ and $\Delta f$ cover one bit period. If the group delay is longer than the bit period, as shown at the bottom, the output level decreases to a level $= \text{Bit Period/Group delay. } (T_b / T_g)$.

Fig. 3. Typical Crystal Filter.
Figure 4. The slow phase shift in the signal (slew rate) that occurs with a conventional filter having a large group delay, or finite rise time.

The maximum data rate possible depends on this phase slew rate, which is related to filter group delay $T_g$. This is a Continuous Phase Frequency Shift Keying (CPFSK) method. Sampling after completing the maximum phase shift is required. A 180 degree shift is shown. $T_g = Q/[2IF]$ is the required filter group delay.

The typical crystal filter shown in Fig. 3 has two crystals, the pass through crystal in the series mode and the shunting crystal in the parallel mode. Often the crystals are weight loaded to broaden the bandpass (lower Q). The crystals operate at different frequencies to result in a filter that approximates the ‘Ideal’ filter. There is a 180 degree phase shift over the bandpass of the filter. The group delay can be calculated from Eq. 1.

Near Zero Group Delay Filters

Abstract:

Used as shown in Fig. 3, the filter cannot respond instantly to a phase change. The phase changes slowly as seen in Fig. 4. Used independently as mono crystal filters, the filter can be made to respond almost without group delay (fast slew rate = zero rise time) at a single frequency. There is no suitable broadband response. The ‘half lattice’ and ‘ladder’ filters using crystals are a filter group that can offer near zero rise time and group delay ($T_g$) to sudden changes in phase, or to an amplitude burst input at a single frequency.

There is a group of filters that have near zero group delay response to a pulse, or which add stored energy to the signal by vector adding to obtain a near zero group delay response. The vector adding filters are all basically filters that depend on the shunting effect of the crystal where the shunting impedance varies and changes the amplitude accordingly. This impedance can be controlled by using a transformer to vary the shunting impedance load.

The transformer reflected shunt filter (TRS) has near zero transient group delay with optimized shoulder reduction. Conventional integrating filters with a delay response as seen in Fig. 4 cannot respond to short bursts unless they have very broad bandwidths i.e. (low Q - from $T_g = Q/[IF]$).
Any filter used with UNB methods must respond to a burst of 1 IF cycle, meaning near zero group delay. These near zero group delay filters function only at a single frequency, or extremely narrow band of frequencies, and are capable of responding to a single RF cycle. All filters used in known communications systems are subject to the BT=1 rule, including these zero group delay filters. Thus the Nyquist BW of a UNB filter is extremely broad, ( equal to the IF ) but the noise bandwidth can be very narrow.

This abrupt phase change characteristic is required for all Ultra Narrow Band modulation methods, such as VPSK, VMSK, MSB, and MCM. These modulation methods are characterized by a single frequency, altered by an abrupt change in amplitude, or phase that lasts only one to three RF cycles for phase reversal keying, or a full bit period with NRZ-MSB. Ordinary very narrow bandwidth filters cannot be used, since they all have a rise time and slew rate characteristic ( group delay = $T_g = [\Delta \Phi / (2\pi \Delta f)]$) which is too large and which will not pass the burst amplitude, or abrupt phase change modulation, without loss. A filter with near zero rise time is required. This is accomplished by means of vector addition.

A filter without rise time or group delay loss can be achieved if the filter is equivalent to a high pass filter, or differentiator, as in D in Fig. 5. The desired filter functions approximately in this manner, or as a voltage divider. Figure 5E is also a differentiator. F in Figure 5 is a differentiator if the coil has a very low Q ( less than 1 ), otherwise it is similar to a high pass filter.

The filters described in Chapter 4 of the Textbook have positive group delay, which smoothes over the missing cycles that are required for optimum Ultra Narrow Band methods. ( See Fig. 1.10 ). A filter is required that does not have appreciable group delay, or has negative group delay, yet has acceptable shoulder reduction. This filter needs to have this effect for only a single frequency – the carrier, or one transmitted sideband in the case of VMSK – since the other Fourier sideband frequencies are irrelevant and removable. The group delay of the filter is determined from $T_g = [\Delta \Phi / (2\pi \Delta f)]$. A zero group delay filter ($T_g = 0$) requires $\Delta \Phi = 0$. There is a group of filters based on the ‘half lattice’ filter that can exhibit this characteristic using vector addition. Also the crystal ‘ladder’ filter can be tuned to exhibit zero group delay using vector addition.
Figure 6. Characteristics of the Crystal Resonator. The phase change with frequency depends upon the change of impedance with frequency.

Figure 7 shows the well known bridge or half lattice filter. In Figure 7A, the filter is used in a bridge circuit that permits the crystal to be used in regions 'b', 'c' or 'd' of Fig. 6. The trimming capacitor Cp adjusts the phase of the circuit while canceling or adding to the crystal shunt capacity. This trimmer enables the circuit to be used at points ‘b’, ‘c’, or ‘d’ of Fig. 6. Point ‘c’ is an inflection point where $\Delta \Phi = \text{max}$ and $T_g = \text{max}$. This frequency varies with tuning.

Note that the circuit is merely a variation of the well known phase shifter circuits shown in Fig. 7 B, C, and D, where the crystal can be resonated on the capacitive or inductive side by Cp, or tuned to be a pure resistance, as in 7B. If the crystal at resonance is considered a pure resistance, then 7C applies. Off tuned to be an inductance, 7B or 7D applies.

Figure 7. The Half Lattice Bridge Circuit and Walker Shunt Filters.

The circuit of Figure 7A divides the signal into two paths, each 180 degrees out of phase. The crystal in the upper path forms a varying impedance load that alters the level of the signal passed through the capacitor Cp. The signal with the phase changes intact passes through the capacitor in the lower path. Figure 7E is the ‘Walker Shunt’ filter, where the transformer is not required, since the circuit can be driven through the phasing capacitor alone, while still using the varying crystal impedance as a shunt load. Moving the ground to the opposite end of the transformer results in a ‘ladder’ filter.
Fig. 8. Phase shift (yellow) and group delay (blue) of a typical half lattice filter **balance tuned** (Fig. 9). The group delay is approximately 500 microseconds. Phase shift approaches ± 90 degrees.

The phase shift through the half lattice filter (Fig. 8) can be observed using most of the circuits shown below. The phase shift is linear around the crystal impedance inflection point as in Figs. 8, 9 and 11.

Figure 9. The swept response of the Bridge and TRS filter when balance tuned.
Figure 9a. Swept response of the 3rd overtone TRS filter (Fig. 19). The shoulders are narrower than for the fundamental crystals. This filter has very little tuning range and a series inductor may be needed to increase this range to the desired frequency.

Unlike the fundamental crystals, this filter can be operated closer to point ‘c’ while the fundamental crystals usually must be operated at points ‘b’ or ‘d’. The Overtone TRS filter has almost no stored reference signal energy.

Figure 9b. Swept Response of all half lattice filters when the phasing capacitor Cp is too large or too small. The dip can occur at the right or left side depending upon the value of Cp.
In Fig. 9b, the magnitude peak ‘c’ occurs when the phase shift rate is at a maximum. The frequency points ‘b’ and ‘d’ in Fig. 6 are marked. Using the more common half lattice filters with fundamental crystals the filter must be tuned to operate at ‘b’ or ‘d’.

Figure 9c. To test the filters with the bandpass response of Fig. 9, a UNB signal on phase 1 for 35% of the time and on phase 2 for 65% of the time was used with a modulation angle of 120 degrees. The data pattern was 0101010101. The test signal spectrum – prefilter – is shown above. The sidebands are almost at the same level as the carrier. After one stage of Overtone TRS filtering, the spectrum seen in Figure 9d is obtained.

Figure 9d. Post bipolar Overtone TRS filter spectrum after 1 stage. The sideband shoulders are reduced nearly 30 dB. As seen below in Fig. 9e there is almost zero phase loss.
Figure 9e. Detected signal before and after one stage of the bandpass filter. Almost no phase loss is measurable. See Figure 38 for the results using cascaded filter stages.

Fig.10 shows speculation on the effect of the phasing capacitor $C_p$ (Fig. 7) in the shunt filter circuits. These plots show how the shoulders shift the dip below or above the center. Balance tuning as in Figs 9a and 9b appears to eliminate the dip.

The normal $X_L$ and $X_C$ relationship is seen in A. This response from a filter indicates the crystal is being operated at the peak point ‘c’ from Fig. 5. Operation at ‘b’ is shown in E. Operation at ‘d’ in D.

If the series capacitance is less than the crystal shunt capacity, the $X_L$ is altered as in B to yield a response curve like that seen in D. The zero crossing is seen as a series dip on the right side.
Figure 11. A network analyzer plot showing a typical bridge or shunt filter response when $C_p$ is large compared to balance tuned. The amplitude plot is in yellow, the phase in blue.

The Swept Response of the Bridge and Shunt Filters is shown in Figure 11. Pointer 1 shows the frequency of the peak. Pointers 2 and 3 (points ‘b’ and ‘d’) show the operating point for the TRS and Shunt filters where the phase loss is a minimum. (for fundamental crystals).

Figure 11 shows the measured phase shift with frequency of the regular bridge and shunt filters using fundamental crystals. It is seen here that there is a very large phase shift $\Delta \Phi$ in a very narrow frequency range at the series and parallel resonant frequencies. The filter cannot be operated properly at those frequencies with fundamental crystals, one of which is at the desired peak amplitude response. However, the filter can be operated slightly above or below this frequency at the points marked ‘b’ and ‘d’. The optimum operating frequency point may suffer an amplitude loss of about 4-5 dB using the ordinary bridge, Walker Shunt and TRS filters with fundamental crystals.

Phase Loss:

The crystal in the half lattice circuits is caused to resonate with the incoming signal at an average phase between phases 1 and 2. This stores energy in the crystal, which then forms a reference to be vector added to the incoming signal to result in the detected phase.

In the Bridge and Shunt filters, this reference level is relatively large and the vector sum for a +90 degree phase modulated signal suffers a phase loss of approximately 50% as seen on the left. Of Figure 12 A. This assumes the filter is passing a vector sum to give a resultant output.

The Transformer Reflected Shunt (TRS) filter has a different effect if the turns ratio between primary and secondary is varied. The reference has a lower value and the sum has less phase loss,
as seen on the right. A ratio of 3 turns primary to 5 turns secondary is effective in this regard. This gives the TRS filter an advantage over the Shunt and Bridge in terms of phase loss. Losses of 10-20% in phase per stage are being obtained.

![Diagram](image)

**Figure 12a.** Speculation Regarding Phase Loss in Filters. **Showing VECTOR ADDITION.**

The loss can be minimized if the transmitted signal has a variable amplitude level for the phases. This is particularly true in the transmitter. The receiver has a limiter so that amplitude variations due to the filter tuning do not pass to the phase detector. Overtone crystals store almost no energy so the reference level is lower. Using quadrature modulation does not allow any build up of a reference, so there is no phase loss.

The phase can be skewed by tuning to lie closer to one or the other input phases so that the vector sum will vary in amplitude. In the case of ± 90 degree modulation, as in 3PRK, the signal can null on one phase and be augmented on the other.

All of the shunt circuits above using fundamental crystals will introduce a phase shift loss per stage if two phase modulation is used. Typically this can be 50% per stage using fundamental crystals, but may be less if the reference level and tuning phase are changed. Optimized fundamental circuits show .75 to .9 phase angle retained with a 120 degree signal phase shift. This becomes less as the angle is decreased. Overtone crystals store less...
energy and there is little or no phase loss as seen in Fig. 9e if some shoulder sacrifice is accepted. **Four phase modulation (quadrature) stores no energy so there is no phase loss.**

![Diagram](image1)

**Figure 12C.** Skewed Reference.

The phase modulation can also be 90 degrees instead of the phase reversing 180 degrees. This method is used with NRZ-MSB modulation. The 180 degree modulation is used with 3PRK. Adjusting the reference phase and phase 2 levels gives better control of the vector sum and results in less phase loss. A phase difference of 120 degrees at the modulator and controlled reference level makes it possible to retain 90 degrees after filtering with acceptable magnitude loss.

![Diagram](image2)

**Figure 12c.**

**Figure 12D.** Filter Phasing to Create Missing Cycles from 180 Degree Phase Modulation. This shows clearly the vector adding principle using signal and stored energy. The filter should never be operated with this phasing. It removes the opposite phase and leaves nothing for the following stages to work with. Tune the filter so that about 120 degrees of the original 180 remain for following stages. All of the near zero delay filters will do this with excellent shoulder reduction.

Phase loss in the transmitters and receivers can be largely eliminated by using frequency multiplication, as in the Armstrong method. 30 degrees of remaining phase shift after three stages
following a 90 degree start can be restored to 90 degrees after frequency tripling. A much better concept is to use quadrature modulation where there is no stored reference energy.

The shoulder level relative to the carrier frequency peak is determined from the driving impedance and shunting impedance. These form a voltage divider. The TRS filter has better shoulders since the shunt impedance and reference level can be adjusted by means of transformer impedance matching, and the response can be balance tuned. (See Figure 9).

![Transformer Coupled Shunt Effect](image)

Figure 13. Transformer Coupled Shunt Effect.

Another variation of the ‘shunt effect’ filter is the “Transformer Reflected Shunt” filter (TRS) shown in Fig. 19. The shoulder rejection of the bridge and shunt filters is shown in Fig. 9. This reflected shunt filter places the resonating element in a transformer secondary, as in Figs. 13 and 19, and couples the shunting impedance back into the primary. The shunt impedance is thereby optimized for best shoulder reduction.

Figure 13 illustrates the transformer coupled (TRS) principle. At the inflection point, which is the point of maximum phase change, the crystal has near infinite impedance. At any other frequency there is a large capacity shunt load that reduces amplitude response, since the circuit is normally driven by a high impedance source. In this circuit the impedance is merely reflected as a shunt load to the source. If an oscillator is substituted for the crystal, the circuit is that of the well known synchrodyne, or autodyne, receiver circuit.

When using 180 degree modulation (3PRK), the filter can be tuned to be phased with the predominant signal phase. In that case, phase one is peaked and phase two is cancelled. This is shown in Fig. 12c. This may offer some advantages in the receiver. Phasing along one phase as in Fig. 12c results in ON/OFF keying, or missing cycles at the output. The smoothing over effect shown in Figure 1.10 of the Textbook does not occur since the group delay is near zero.

Unfortunately, the signal phased as in Fig. 12c is converted to an AM signal and cannot then be used with a limiter.

NOTE: The HP5714, 5753 and E5071 Network analyzers can give false or misleading results regarding the resulting group delay when used with these filters if falsely interpreted. Vector adding is ignored. The TRS and other half lattice based filters will show a residual group delay as high as 500 microseconds as seen in Figure 8, which would make them useless for UNB modulation. This is due to reference energy stored in the crystal. A Q of 15,000 at 32 MHz will show a residual group delay of 35 microseconds. Negative group delay must be assumed.
The crystal at resonance is a very high resistance, which combined with the driving capacitor makes an RC differentiator. An RC differentiator has no group delay and zero envelope response (rise) time. Off resonance, the shunting impedance of the crystal is a capacitor (Fig. 6) and the circuit performs as an integrator with rise time and envelope delay. If there were no energy stored in the crystal to be vector added to the incoming signal, the circuit would show no phase loss.

The optimum filter stores the least reference energy in the crystal while maintaining the largest shoulder reduction. Overtone crystals store very little reference energy.

The phase loss can be measured using a phase detector that is linear with phase change. The NE602 circuit shown in Figure 33 has this characteristic. Some results using the Overtone TRS filter are also given in Figure 9 to show almost zero phase loss.

The burst response for TRS filters is closer to 30 nanoseconds (for 32 MHz IF) as seen in Fig. 15 below, since the filter operates by vector addition of the reference and the incoming signal as explained above. This is a near zero group delay response where the rise time = group delay = 1 IF cycle.

![Figure 14](image)

Figure 14. The Swept Response of 3 Cascaded half lattice Filter Stages (Fig 8). This is applicable to the bridge, shunt and TRS circuits, particularly to Fig. 17.

It should be noted that these filters as used resolve each individual IF cycle without group delay. The Nyquist Bandwidth from BT = 1 is equal to the Intermediate Frequency. The noise bandwidth however, is as seen in Figures 9 and 12. These filters are analogous to phase locked loops where the noise bandwidth is narrow, as determined by the loop filter time constant, and the
phase slew rate is also set by this time constant. However, the phase change at the output of the phase detector ahead of the loop filter is near instantaneous. (See Textbook Fig. 4.7).

With a conventional filter (Textbook Chapter 4), the noise bandwidth and Nyquist bandwidth are close to the same. The raised cosine filter (Figs. 2.3 and 4.4) with $\alpha = .5$ has a noise bandwidth twice that called for at the Nyquist minimum, as in the ‘Ideal’ filter, where $\alpha = .0$

The filters in Figs. 9 and 12 have a noise bandwidth of approximately 1 kHz at the 3 dB points. The Nyquist Bandwidth from $BT = 1$ is the intermediate frequency (crystal resonant frequency). Frequencies as high as 96 MHz have been used. The rise time, or group delay is 1 IF cycle period. Overtone crystals do not have the tuning range that is available with fundamental mode crystals. See page 32. A series inductor must be added.

That these are truly near zero group delay filters can be shown with a burst test (Fig.15). A burst of 4-5 cycles will show the 1 cycle rise and decay times. (Fig. 15).

Ultra Narrow Band filters must have a burst response = 1 IF cycle. Hence the group delay must be = $1/IF = T$. The Nyquist BW from $BT = 1$ is = IF. The noise bandwidth (figures 9 and 12) depends on the crystal Q.

![Picture](image)

Figure 15. Burst or Impulse Response of the TRS Filter to Multiple IF Cycles. The top trace is the input signal, while the lower trace is the filter output for one stage. This photo shows near zero group delay, since there is no rise time or after impulse ringing. There is no stored energy in the crystal when the bursts are not close together. Overtone crystals have little or no stored energy.

**All sinx/x (sideband) frequencies are rejected due to the capacitive load on the filter at all frequencies off resonance.** (Fig. 10). Figures 5.4, 6.7 and 6.8 in the textbook show how the sinx/x products are reduced without having any effect on the detected phase modulation.
Figure 16. The Response to 3PRK Modulation Showing a 1 cycle Phase Reversal. The lower trace in Fig. 16 is the recovered data clock. Missing cycles at the phase change edges are clearly seen.

Figure 17. Half Lattice Walker Shunt Filter Variations. The series inductor may be necessary to improve tuning range, especially with overtone crystals.
Adding a series capacitor or inductor as in A or B will alter the crystal resonant frequency. It is usually necessary to add a series inductance to an overtone crystal to increase the tuning range. See Figs. 17 and 18.

Do not use general purpose transistors such as the 2N2222 in these circuits. Use only RF transistors such as the BF240. There is capacity feedback from collector to base in the shunt and bridge filters that lowers the gain with frequency. Use a low gain per stage unless this characteristic is desired.

Figure 18. Bridge (left), Walker Shunt (center) and TRS (right) filters. An inductance is often added to lower the frequency and increase the tuning range. This is very necessary with overtone crystals. Do not use general purpose transistors. Use only HF RF transistors such as the BF240.

Figure 18(a). The floating bridge filter, which combines features of the Walker Shunt and bridge filters, has good shoulder reduction with little phase loss. There will be a 30-40 percent phase loss. The circuit may be amplitude and crystal sensitive. Third overtones work well. Note the resemblance to Fig. 7E. Replacing the 3 pf capacitor with a ground connection results in a ‘ladder’ filter.
Useful versions of the parallel Transformer Reflected Shunt filter are shown above. When using overtone crystals it may be necessary to use an inductance in series with the crystal instead of the 47pf shunt capacitor. The shunt capacitor is used to force the operating point to move to resemble a series circuit, with the response seen in Fig. 18 instead of that seen in Fig. 12. Adding the inductance lowers the frequency and increases the tuning range. Overtone crystals have low tuning ranges. See page 38. It was necessary to add a .33uH series inductor for some 60 MHz 3rd OT crystals.

Fig. 19a. The TRS filter which can be near balance tuned for best shoulder reduction. The variation at the right has almost no phase loss using third overtone crystals. It is less effective with fundamental crystals.
Overtone TRS filter with large voltage output and very small phase loss.

The sidebands in the nulling filter are combined 180 degrees out of phase, hence can be level adjusted between the 180 degree phased inputs to cancel in the bridge circuit. In practice, the balancing pot sets the level of sideband (shoulder) reduction.

Some measurements may lead experimenters to believe the phase shift obtained with UNB modulation is a result of the sidebands being in a quadrature relationship to the carrier as in the Armstrong method for PM (Fig. 2.5). This is not the case as seen with VMSK (Chapter 5), where there is no carrier, and as evidenced by the phase shift between carrier and sidebands obtainable by off tuning the sideband nulling filter. This phase shift difference between carrier and sidebands can rotate through 360 degrees with minimal effect on the detected signal, thus proving the phase differences in the signal come from the switched input and not from a relationship between sidebands and carrier as in Fig. 2.5. See Fig. 7.16. 1 KHz off tune can rotate the sidebands relative to the carrier by 90 degrees and there is little or no observed loss of detected phase.
Figure 19D. The series emitter filter.

Attempting to obtain too much sideband reduction per stage causes increased phase loss. 0 to 12 dB sideband reduction per stage is obtained with minimal phase loss. Tuning for maximum sideband reduction possible per stage can cause the phase loss to approach the calculated line in Figure 6.13.

Figure 19E. Amplitude vs Phase for the sideband nulling filter. This is an extreme case. See Figure 9a for the typical case.

*The overtone TRS filter has been found to be the filter of preference, since it has very little phase loss and acceptable sideband rejection. In some cases the series emitter filter has performed better.*

In bipolar transistor circuits *Taking the signal from the collector is preferred to connections to the emitter. Emitter load connections often feedback to detune the filter and cause a loss in shoulder depth.*

All half lattice based filters have a long rise time (group delay) at the peak amplitude, but near zero group delay response time to instantaneous phase changes slightly off resonance (Fig. 9). *BT is*
always = 1. For a single frequency, this means a near infinite Nyquist bandwidth for phase changes, but a very narrow noise bandwidth, which establishes the reference, determined by the Q of the crystal.

\[ B = IF \text{ Freq.} \quad T = 1/IF, \text{ for the phase relationship} \]

For the amplitude relationship and noise bandwidth, the crystal Q is \( \text{Freq}/3\text{dB BW} \), which is from 20,000 to 50,000, so \( B = IF/Q \) and \( BT = 1 \), so \( 1/T = IF/Q \), or \( T = Q/IF \).

Example: Using a filter at 60 MHz in a Cable TV modem with Ultra Narrow Band modulation. The signal is a single frequency at 60 MHz. The signal contains phase modulation which is detectable, but no AM. There are no usable sidebands. (Figure A2.14).

The phase response group delay of this filter is equal to \( T_g = 1/f = 1/60\text{MHz} = 16 \text{ ns} = 1 \text{ IF cycle} \). The amplitude response peak group delay (crystal Q based, see Eq. 1.) is approximately \( T_g = Q/IF = 20,000/4(60\text{MHz}) = 333 \text{ microseconds} \). Values over 500 microseconds have been measured at peak resonance.

Fig. 9 is the noise bandwidth of the TRS filter balance tuned. The scale in Fig. 9 is 5dB per vertical division. The phase is non linear and does not conform to the linear LC or Gaussian curve in Fig. 25. The response is also obtained when the drive capacitor in the shunt filter is large, otherwise Figs. 9 and 12 apply.

The filters cannot be used in this balanced condition. A dip at the right or left side is necessary to retain a phase shift. See Figs. 9, 12 and 22.

Figure 20. The response of the TRS filter to 1 phase reversed cycle. This phase reversed cycle is easily detected by the phase detector. The indicated group delay is 20 nanoseconds for a 48 MHz IF. For ‘Ultra Narrow Band’ applications it is essentially a zero group delay filter. Note that the modulation is not lost as in Fig. 1.10. See also Figure 13.
Figure 21. The Crystal being used in the series mode. Ladder circuit on left..

Figure 21a.

In Figure 21a the filter is shown using a series mode crystal transformer coupled driving a critical load resistance. At series resonance, the circuit is a simple voltage divider. The capacitance added by a load, although very small, has some effect. The critical resistor must be trimmed to obtain maximum shoulder reduction. Fundamental crystals perform much better than overtone crystals, which lack tuning range. This is not the best filter for Ultra Narrow band use.

Fig. 22. Swept Response of the Series mode filters.

Near zero group delay is also available when the crystal is used in the series mode. The circuit must be tuned slightly above or below series resonance to operate in an optimum phase shift
region. The TRS filter can be tuned to have the parallel or series response by increasing the shunting capacitor from crystal to ground (Fig. 16).

Tuned at exactly the peak resonant point the circuit has considerable phase loss which destroys the modulation. Refer to the zero group delay regions in Fig. 8. However, this circuit tuned to peak response where there is excessive group delay can be used to obtain a steady reference for phase detectors.

Figure 23. The ZQM filter. There are three possible variations in the feedback. The ZQM (Zero Group Delay Q Multiplier) filter is an unusual regenerative circuit in that it can be operated as a Q multiplier, or locked oscillator, by adjusting the variable resistor. When used as a Q multiplier, just before it breaks into oscillation, it will have shoulders of 15-18 dB. The feedback circuit determines the bandpass. The signal never passes through, or connects to, the resonator. Frequencies outside the resonant frequency of the crystal (such as the sinx/x products) are returned as negative feedback, reducing the output level.

Figure 23A shows the measured group delay of the ZQM filter when the Q is approximately 50.

Figure 23A. Network Analyzer plot of ZQM filter showing zero group delay (blue).
At 32 MHz, the cycle period $T$ is 33 nanoseconds. At the 3 dB point on the response curve the measured delay is 4.9 nanoseconds. Beyond that the group delay is negative. This filter could be operated closer to the peak without serious group delay loss. The magnitude is shown in yellow. The group delay of 1.2 microseconds at the peak is too large for most UNB uses, but the group delay off peak is negative and could have UNB uses.

This Feedback filter shows good results with $Q = 10-12$ It can be used with NRZ-MSB or 3PRK as a pre filter. Did well with 2 MHz 3 dB bandwidth at 32 MHz using 3PRK and MCM. Shoulder reduction was -16 dB with minimal phase loss. Be careful of second harmonic or spurious higher frequency response. This was the only LC filter found that did not adhere to the $T_g = Q/[IF]$ rule. Bandwidth is adjustable over a wide range. Unfortunately the filter is level sensitive. (try with FET)

The near zero group delay filters are not the same as the raised cosine filters, or conventional LC filters. The transient group delay due to vector adding can be as low as one IF cycle period, hence there is no upper frequency bound as long as the intermediate frequency can be raised. The limitation would appear to be the resonator frequency. Crystals as of Jan. 2010 appear to be limited in frequency to about 300 MHz. There is some comment in the literature that SAW resonators might be usable at higher frequencies.

The crystal impedance plots and formulas for $T_g$ are also published by numerous crystal and SAW filter manufacturers. The group delay of the filter is $T_g = [\Delta \Phi/(2\pi \Delta f)]$. The group delay in this discussion is not the differential group delay normally associated with SAW filters or multi-pole filters.

Automatic frequency control using a PLL can be obtained from the filter shown in Fig 7.4. At the peak there is a large phase shift with frequency. A phase detector used with a PLL will hold the frequency at this cross over point. If the desired zero group delay operating point is slightly above this frequency all that is required is a filter after the limiter to be used with the PLL phase detector to obtain a reference. Assuming the filter crystals all have the same frequency drift with temperature, the PLL circuit can correct for temperature drift.
A reference phase for use with phase detectors can be obtained by tuning the reference to the peak where the group delay is large and the modulation phase changes are no longer obtainable. Either a series mode or parallel crystal circuit can be used to obtain this reference.

The zero group delay filters do not have the same noise bandwidth as Nyquist criteria filters. The noise bandwidth instead is related to the crystal circuit Q, which can be 10,000 or higher. Cascading raises the apparent Q with the number of cascaded stages so that a bandwidth of 1 kHz at 72 MHz is obtainable. This can be seen in Fig. 12 where a 3dB bandwidth less than 1 kHz is seen.

The filters normally used for CPFSK modulation methods (all methods presently in use other than Ultra Narrow Band methods, or methods with zero group delay filters) are integrators, as seen in Fig. 1. Fig 5 A, B, and C are examples of this type of integrating filter. Ultra Narrow Band methods such as VMSK and MSB require filters with as little group delay as possible in order to detect the phase or amplitude change of a single IF cycle. D, E, and F are the equivalent circuits. (Note BT = 1, T must be as small as possible to pass the abrupt phase change in one cycle. This means of course that B must be large and T must be small). B is the Nyquist bandwidth and not the actual filter bandwidth, which can be much narrower.

Overtone crystals can be used. Caution however, a 3rd overtone crystal will also show a response at the fundamental and vice versa. The loading resistors and transformers can have an effect on the response. The higher the loading impedance the better.

Single pole LC resonators and crystals generally have about 16-17 dB drop off at the shoulders per stage. (see Crystal Response curve Fig.6 and Universal Resonance Curve - Fig. 24). Cascading will improve this. For example, 3 poles cascaded will have shoulders down 45-50 dB. One should not be surprised however if the shoulders are only 10-12 dB below peak for a single stage when the components are not optimized.

Better shoulders than this are available from both the shunt and autodyne filters by controlling the transformer and/or inductance values and turns ratios. The TRS filter shoulders generally are about -15 dB, but may have better shoulders with fundamental crystals and lower shoulders with overtone crystals.

Overtone crystals lack tuning range. It is usually necessary to add a series inductance to an overtone crystal to obtain the necessary frequency adjustment. The inductance should be near that which resonates with the nominal shunt load capacity of the crystal. See “Note on tunability” below (pp39).

These filters are analogous to phase locked loops where the noise bandwidth is narrow as determined by the loop filter time constant, and the phase slew rate is also set by this time constant. However, the phase change at the output of the phase detector ahead of the loop filter is near instantaneous.

Crystals do not have a response similar to the Universal Resonance Curve (Fig. 24). The LC curve in Fig. 24, which is also close to that of the Gaussian filter curve, has no inflection point. (See Fig. 6). Ordinary LC filters cannot have near zero group delay. Parallel mode LC (ΔΦ) is shown in Figure 24. The phase change, as plotted, reverses for the series mode. Note the (ΔΦ/Δf) change for a crystal in Fig. 10.
A filter having a response curve similar to the LC filter is the Gaussian Filter, which has less phase distortion. The transfer functions for both are given. The Gaussian filter is easily obtained using digital filters. When (Bandwidth) x (Rise Time) = (BT) = 1, the filter has a 3 dB bandwidth $\Delta f$ and $T_g = \left[ \frac{1}{\Delta f} \right] = IF$ (Page 2).

All ultra narrow band filters have very broad shoulder bandwidths outside the narrow bandwidth shown by the swept pattern. Also, overtone crystals have more than one response peak. Some additional filtering is required to confine the wide open input bandwidth to a narrower bandwidth ahead of these filters to prevent noise overloading, or input at the overtones. This is the subject of another section below covering pre-filters. (Filter Overload).

![Diagram](image)

Figure 24. The Universal Response Curve for an LC. $\alpha = Q$ (cycles off resonance / resonant freq).

**The group delay of this filter is determined from $T_g = Q/[IF]$** IF is the filter freq. This is not a true representation of crystal phase vs frequency using the Bridge, TRS or Shunt filters. Refer to Figs. 9 and 11.

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**Inductors:**

The load and coupling inductors shown in the above circuits are low Q ferrites. The table below shows the parallel resistance as measured on a bridge for a single wire through the center of EMI snubber ferrites. The resonant capacity for any frequency is essentially 0 pf $-$ 1 pf, so that any circuit capacity causes the circuit to operate at a frequency below LC self resonance. The impedance rises with the number of loops through the bead. 4 leads through the center raises the impedance by a factor of 12. As an example, at 85 MHz, the measured impedance using a 9 mm bead with 4 wires through the center was 1.45 k Ohm.
Table 1. The impedance for a single wire through the ferrite core. Not looped. (EMI snubber beads).

<table>
<thead>
<tr>
<th>Size</th>
<th>9mm</th>
<th>5mm</th>
<th>3mm (Panasonic beads)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Freq.</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>24 MHz</td>
<td>70</td>
<td>40</td>
<td>25 Ohms</td>
</tr>
<tr>
<td>50 MHz</td>
<td>80</td>
<td>45</td>
<td>30 Ohms</td>
</tr>
<tr>
<td>100 MHz</td>
<td>92</td>
<td>50</td>
<td>32 Ohms</td>
</tr>
</tbody>
</table>

The FairRite .236 inch bead has a Z = 60 Ohms at 100 MHz.

When winding transformers and coils using beads or donuts, try to keep the turns separated to reduce internal capacity and increase the self resonant frequency.

Half Lattice Filter Transfer Functions:

Amplitude:

\[ G(j\omega) = A(j\omega) = \frac{1}{1 + jQ(2\Delta f / f_o)} \]

The amplitude response G(f) or A(f) of the half lattice group of filters is shown in Figures 7.4 and 7.10. Note that the Nyquist Bandwidth of this filter is Intermediate Freq. = Sampling Rate, while the noise BW is approximately 1-2 kHz. \( \Delta f \) is the frequency off resonance in this equation. This equation is valid for the TRS filter balance tuned (Fig. 9 above).

The UNB signal is:

\[ H(t) = [K_1 \sin(\omega_0 t + \theta_1)] \] for pulse duration \( t \) (3PRK), or phase 1 NRZMSB.

\[ H(t) = [K_1 \sin(\omega_0 t + \theta_2)] \] for remainder of bit period \( T-t \) and phase 2.

The frequency response is:

\[ H(j\omega) = [\sin(\omega_0 t + \theta - \phi_{\text{rees}})] \], which is a single carrier frequency \( \omega_0 \) shifting in phase \( \phi_{\text{rees}} \).

There is no correlation or mixing of the two \( H(t) \) components.

The phase response is:

\[ H(\phi) = [0] \] For frequencies below resonance in pass band

\[ H(\phi) = [-160 \text{deg} \text{rees}] \] For frequencies above resonance peak.

The group delay of the filter is \( T_g = \frac{\Delta\Phi}{(2\pi \Delta f)} \). During the periods \( t \) and \( T-t \), or between phase one and phase 2, \( \Delta\Phi \) is zero, hence the group delay is zero. (Chapter 6).

Signal Interference

In order for an ultra narrow band signal (such as VMSK or MSB) to be detected, the level must be high enough to trigger a CMOS XOR gate or D flip flop phase detector to a positive on/off state. For purposes of comparison, this is normally assumed to be a rail to rail input of 5 volts.
The actual gate window using a 74HC, or AC chip is typically 0.1 volt located somewhere between the specified 2.2 V\text{IL} to 2.8 volts V\text{IH}. A 5 volt peak to peak input (after a good limiter), crosses this window with a wide margin and overcomes any noise on the input signal as well as chip and power supply noise. Some limiter chips which can be used have built in quadrature detectors so the above comment does not apply.

The interference most likely to cause a problem is an interferer having an amplitude voltage level large enough that, when added vectorially to the desired signal, will cause a large drop in the phasor sum beyond the ability of the limiter to handle it. See Fig. 25.

It can be seen from the Figure 25 below that this AM level is dependent upon the filter and limiter. The limiter reduces the amplitude of the noise vector and the equivalent vector from any adjacent channels. In practice this limiter margin is usually less than 2 dB.

Fig.26. Signal to noise relationship.

It is well known that the total interference level after a bandpass filter must be below the signal level. That is C/N must be above 0 dB. (Shannon’s Limit for a 2 level system).

AM and FM noise interference are present ahead of the limiter at the output side of the IF filter, which is presumed to have a very narrow bandwidth. Interference differing in frequency by a wide margin will presumably not be passed. Interference close in to the filter center will pass at a level depending on the closeness to the filter center.

Standard practice requires that all filtering be done ahead of a limiter. This is because a negative limiter margin will leave holes in the modulation due to the excess AM level. However, if the total noise level does not exceed the signal level, some additional filtering can be used after the limiter. (A positive Limiter Margin.)

**Filter Overload:**

This section is intended to provide guidelines to obtain the optimum C/N ratio. It does not imply that the UNB methods cannot be used without pre-filtering, but that the C/N will be improved if a pre-filter is used. In any case, given a signal that is strong enough to overcome all noise sources, the signal can be detected, as indicated in Fig. 25 with a positive limiter margin.
\[ SNR = \left( \sin m \right)^2 \left( \frac{SignalPower}{NoisePower} \right) \]

\[ m = \beta = \pm 90 \text{ for 3PRK or } \pm 45 \text{ degrees for MSB.} \]

(Applicable to all 2 level phase modulation methods)

Signal Power/Noise Power = SNR = C/N for +90 degree or missing cycle modulation. For 3PRK and MCM, \((\sin m) = 1.0\) and C/N is equal to \(E_b/n\).

Noise power bandwidth is the major concern. There are two noise sources to contend with. Source 1 (\(BW_1\)) is the noise and bandwidth associated with the ultra narrow band filter (1 kHz) and source 2 (\(BW_2\)) is the noise source that comes in from the shoulders outside the filter bandpass. All bandpass filters have broad shoulders extending from 0 to infinity. This can be seen in Fig.12.

The noise portion of the equation above becomes:

\[(\text{Noise Power/BW})_1 + (\text{Noise Power/BW})_2\]

\[(\text{Noise Power/BW})_2\] must be kept small, that is - less than \((\text{NP})_1\), or the filter will be overloaded. Assume there is a pre- filter, or post filter, that is less than 1 MHz wide and the shoulders are -30dB down. Then \((\text{Attenuation} \times \text{Noise Power/BW})_2\) is the factor that must now be used.

The NP1 bandwidth assumed is 1 kHz, so the NP2 bandwidth could be 30 dB (1,000 times) wider. Noise power is proportional to bandwidth, so the pre-filter could be 1 MHz wide at the 3 dB points. To allow a margin of safety, 6 dB is arbitrarily subtracted so the pre-filter bandwidth allowable is only 250 kHz.

Zero group delay pre-filters with bandwidths this narrow are available in the ZQM filter. If used, the lowest (best) C/N is obtainable. This has been measured to be better than for BPSK, where \(SNR = C/N\), but the noise bandwidth is many times greater for BPSK than it is for MSB ultra narrow band methods. (For BPSK, \(BW = \text{Bit Rate}\), for MSB = \(1.4/ N\) kHz per stage)

(Note: The RF bandpass filter in a superheterodyne receiver is a pre-filter - wider than the IF bandpass). It is also subject to the group delay rules. The RF filter ahead of the mixer must not be too narrow.

Unless a proper bandwidth limiting filter is used, C/N and SNR measurements will be falsely too high due to interference coming from the shoulders.

Given an ideal filter with zero shoulder comeback, or, an out of filter bandpass region with infinite attenuation, there is no problem with using a broadband noise source in testing, or with a strong adjacent channel interferer. If the filter has only a limited level reduction at the sides, it can cause a problem if the noise bandwidth used for the C/N measurements is too wide. Also the signal plus noise power level can cause the filter circuit to go into limiting. This overload problem can occur if there are many strong interfering adjacent channels as well.

This is one of the reasons RF filtering is used ahead of the IF filtering in Superheterodynes.

Assume a single mono-pole crystal filter with a 3dB noise BW of 3 kHz, but only 16 dB of shoulder rejection at the sides, and a CW error breakpoint at the detector of -6 dB. The C/N will be accurate as long as the noise bandwidth from the noise test set is less than 30 kHz. If the noise bandwidth is 300 kHz, the noise power will overload the filter by 10 dB and the measurement
will be off by 10 dB. A CW interfering signal at +10 dB located at either side, will just cause the signal to reach the break point. Noise power varies linearly with bandwidth.

![Diagram of noise levels and shoulders](image)

**Fig. 27.**

This points out the necessity for the narrowest possible noise bandwidth at the input to the ultra narrow narrow band filter, and of course, the greatest possible off center rejection in the narrow bandwidth filter itself. ( Or cascaded filters ). ( Pre- Xtal bandpass filtering).

It is almost impossible to make an exact calculation of the signal and noise relationships vs bandwidth, since there are so many variables. About all that can be achieved in practice is to use the known available mathematical relationships to improve the system overall.

Given this assumed bandwidth, the shoulders as shown in Figure 27 should be approximately 35 dB below the peak. This shoulder value is easily obtainable with 2 or more stages of the zero group delay filters cascaded. An assumption is made that the noise at the filter input is balanced between that passing through the filter and that passing around the filter at the filter output.

This consideration is based on noise only. If there is a strong interfering signal outside the narrow bandpass, this must be considered as well. Assume an interfering signal that is 35 dB stronger than the desired signal. This signal would pass through and add vectorially to the desired signal, with the result that the limiter could have a blank-out periodically, as seen in Fig. 26. The filter should have shoulders lower than the expected peak interference. Generally this is greater than 35 dB in practice. Most UNB filters in use have shoulders that are at least 40 dB and sometimes 50 dB below peak.

When a strong interfering signal is present, consideration must be given to the IP3 values associated with the amplifying devices in the signal chain. ( Due to Inter-modulation and Cross Modulation ). A MMIC device can become a very good mixer when one of the signals approaches 0 dBm input level. See references [12] and [18]. If the interference always occurs at a known location, it is best to trap it out before allowing it to reach a portion of the circuit where it can cause inter-modulation, or limiter blank-out.

Filter overload can occur for a different reason other than white noise. Suppose there are 5 channels of equal levels at the input. Then the total peak vector will be 5 times as large as the single desired carrier vector. Most of the filters shown above have an active element - a bipolar transistor or FET. These devices have input limits. Beyond that limit they either saturate or cut off. Assume the device can accept a voltage swing of 1 volt pp. This is fine for one channel at that level, but once added channels increase the vector length, the device saturates and has a
cutoff that destroys the desired data. The input level must be lowered to 1/5 the per channel level (14 dB) to prevent the effect of a negative limiter margin. (Fig. 25).

**Do not overdrive any active device in the filter chain ahead of the limiter.**

**Limiting Noise Bandwidth:**

Assuming 100 cycles per bit, each cycle has a period of .01 microseconds for a 100 MHz IF frequency. For a 1 Mb/s data rate, the bit period is 1 microsecond. If 3 cycles are needed for the impulse, the group delay of the filter should not exceed .03 microseconds. A normal Nyquist, or Ideal filter, would require a 1 MHz bandwidth and a group delay of $T_g = [1/2 \Delta f] = .5$ microseconds. The 3 dB bandwidth required for a Gaussian filter to pass .03 microsecond (30 ns) impulses is $T_g = [1/4 \Delta f] ---- .03 = [1/4 \Delta f]$ or $0.12 = [1/4 \Delta f]$ , so $\Delta f = 8.5$ MHz. It is desired to pass a 1-3 cycle impulse through a much narrower noise bandwidth, so narrow band Gaussian or LC filters cannot be used, since they introduce a considerable group delay increase.

**Conventional filters have too much group delay.**

For 3PRK modulation, a ZQM filter is the only practical pre-filter. For other ultra narrow band methods, some group delay may be tolerated. In general, a conventional pre-filter with Q less than 4 is required.

![Figure 28](image_url)

Figure 28 shows the burst response time of a crystal used in the series mode where the burst (upper) passes through the crystal. The crystal is caused to ring by the bursts (lower), with a rise to maximum level depending on the Q of the crystal. The calculated response is pure textbook (normal group delay). The rise time is typically several hundred microseconds. In this test, the crystal was being pulsed with 5 cycle bursts at a 266 kb/s rate. This characteristic is used in narrow band techniques to obtain a reference, which is not affected by the abrupt short burst phase change modulation. See Fig. 16.
Figure 13 above shows the Burst or Impulse Response of the Shunt or TRS Filter to Multiple IF Cycles. The top trace is the input signal, the lower trace is the filter output for one stage. This photo shows near zero group delay, since there is no after impulse ringing or rise time at the start.

Fig. 30. Reverse Burst Response of a filter with some group delay. This is with 2 cycles phase reversed in the IF stream. A slew rate- \( \Delta \Phi/\Delta t \) of 0.8\( \pi/\Delta t \) - can be calculated from this response. As long as the slew rate exceeds 1/10 radian per RF cycle, the signal can be detected with a very low BER. The filter in Fig.30 requires more than 3 IF cycles to complete a phase reversal.

Figure 13 shows the amplitude response of both the bridge filter and the shunt filter tuned to use the crystal in the parallel mode. A 4 cycle burst at 24 MHz results in 4 cycles being passed as seen in Fig. 13, with full amplitude being reached on the first cycle. Allowing for the loading of the scope probes etc., the group delay is about 30 nanoseconds, but can be less than 20 ns. It is not the calculated 62.5 microseconds according to the \( T_g \) equation ( Eq. 1 ). A Network analyzer will show 62.5 microseconds.

The calculated group delay for a conventional LC or Gaussian filters is traditionally:

\[
T_g = \left[ \frac{1}{\Delta f} \right] / \text{??? obviously, a very narrow bandwidth conventional filter has a very large group delay. It is also related to } Q. \quad T_g = Q[4\text{IF}] \quad \text{IF = Int. Frequency.}
\]

As noted above, there is an associated equation for the rise time of the conventional filter: \( T_r = 0.7/B \), where B is the 3 dB bandwidth of the filter. This is the time from 10% to 90% on the RC curve. In practice it is considered to be \( T_r = 1/B \). A conventional filter with \( 'B' = 2 \) kHz has a rise time of 350 microseconds. If the group delay is too large, (bandwidth too narrow), Ultra Narrow Band modulation bursts will not pass through the filter. The object of the burst tests of VMSK/MSB filters is to find filters that do not obey this rule. They must have a very narrow noise bandwidth and a very fast rise time.
The shunt filters and bridge shunt filters do not obey this rise time rule as seen from Fig. 29, whereas the series mode crystal as in Figure 16, or a tuned LC, does obey this rule.

A slew rate can be calculated from the rise time. A 180 degree shift in the ideal filter is considered to be 100% of rise time. A change from 10-90% is 80% of the 180 degree change, so there is a slew rate of 144 degrees/time interval. A slew rate $\Delta \Phi/\Delta t$ of $0.8\pi/\Delta t$. This slew rate is very important in filters for ultra narrow bandwidth methods which must recognize a change in amplitude or phase lasting only 1-2 cycles, hence must have a very rapid slew rate, or fast rise time, or near zero group delay. In most analyses, it is assumed to be $\pi/\Delta t$. Refer to Figure 4.

![Fig. 31](image)

**Fig. 31.** A: Zero Group Del. B: Very Low Group Del. C: Normal Group Del. D: Zero Rise Time E: Fast Rise Time F: Slow Rise Time

The desired rise time for the conventional (Optimum or integrating) filter is shown at the right. See figures 1 and 2. The desired zero rise time for the ultra narrow band filter is at the left. The center B shows what is being achieved (slight group delay). A straight wire is represented by A.

**When the crystal is used as a shunt load, it becomes a near infinite resistance at parallel resonance.** The result, when using a burst input, is as if the crystal were not there and the transfer function is like that of a straight wire (31A), or a differentiator, as seen in Figs. 5D and 15.

![Fig. 32](image)

**Fig. 32.** Phase Detector.

Figure 32 shows using the crystal in the series mode to obtain a near continuous phase reference which is not affected by short burst differences from ultra narrow band signals. The series mode crystal has a very large group delay as seen in Fig. 28.

The XOR gate is a 74HC86 or 74AC86 chip. The XOR output requires doing without the 45 degree phase shift (bypass it, or use a smaller shift). The D flip flop requires it.
phase at the clock inputs can be reversed using + - on the input to U4. This circuit is useful for most ultra narrow band modulation methods. Using the XOR gate is best without the LCR phase shift combination and directly coupled. Note results below. The XOR gate functions here as a quadrature detector. Using the D Flip flop as a phase detector requires the phase shift and a change in clock phase.

Fig. 33. The linear phase detector usable with all UNB methods. The output level is linear with the phase difference between phase 1 and phase 2.. The circuit is generally preceded by an amplitude limiter.

Many experimenters have fallen into the filter trap, if they use Eq. 1 and a conventional filter with MSB modulation. Any baseband analysis of an Ultra Narrow Band system will fail and all modulation can be lost due to group delay. Baseband analysis depends on the modulating waveform. At RF this is not the case.

A filter with a normal Q, ie for a narrow BW in $T_r = 1/BW$, has a long rise time and loses the modulation. (See correlators with integrating filters in section one).

$$T_g = Q[IF]$$

A burst of a number of cycles at a given frequency with a conventional filter will show a rise time according to Q. With a high Q, there is a long rise time and large group delay. The missing cycle, or a phase changed single cycle, will not pass a conventional filter with a large Q.

One must use a zero group delay filter ($T_g = 0$) responsive to a single frequency. Such a filter would have a very low Q in terms of rise time (applying the formula in reverse), but it has a very good noise bandwidth. With Q = 1, the rise time can approach one RF cycle. The Nyquist bandwidth then is = the IF frequency, but the noise bandwidth is 2-3 kHz or less.

Figure 13 shows the burst response of a near zero group delay, very fast rise time, filter. The rise time is approximately 1 to 2 IF cycles. Instead of a burst response
test, missing cycle modulation will give a clear picture of group delay as well. (Fig. 30). The difference relates only to the duty cycle.

There is no known filter with zero group delay for use at baseband. At Baseband, it is necessary to use conventional filters, with a Q of 1.0 to 1.5 at most. This is why a baseband analysis of VMSK or MSB using a narrow band conventional filter fails. The Q of the filter kills the modulation.

Phase locked loops can be used as filters and detectors. They can enable the quadrature detector used with MSB to have the best performance. See Best [7].

Detector Note:

The waveform of the original IF signal must be preserved if the method is to be usable. A synchronous detector is equivalent to sampling the IF waveform at the minimum sampling rate, which is the IF frequency. It is also the equivalent of a zero IF, or Direct Conversion receiver. Note: The sampling for ultra narrow band methods is not done at the bit rate, but at the IF Frequency.

When the signal is up converted, there must be enough cycles in the RF frequency to enable the IF waveform to be reconstructed in the down conversion. Probably, the RF frequency should never be below 6-8 times the IF frequency, otherwise the signal on down conversion may not truly match the original IF waveform.

Nyquist’s Sampling Theorem cannot be violated. Do not expect an up conversion or down conversion to properly represent the original IF modulation if there are an insufficient number of samples. An RF mixer is actually a sampling device.
Figure 34. Scope photo of XOR detector output after 2 stages of bridge filter, sampled at the Intermediate Frequency. Since the Ultra Narrow Band modulation is done on a cycle by cycle basis and the filter operates on a cycle by cycle basis, the XOR chip used as a detector has a cycle by cycle response. Figures 33 and 35 are the XOR outputs for 5 cycles phase altered out of a longer data stream. These cycles must be smoothed over to result in a data pattern recovery. The D flip flop flop and NE602 circuits have the cycles smoothed over.

The scale is 1 Volt per division. With no filter, the bleed through of the carrier at the top is less, but the spikes indicating the phase shift of 180 degrees are approximately the same. This photo shows some group delay (about 1 cycle). The first spike may not reliably reach a maximum peak, but the second will. To detect this signal, a voltage clipper is used to respond to the large spikes while rejecting the carrier spikes at the top left. A low pass filter used here will result in a much clearer and broader spike. The detector may exhibit a one IF cycle indeterminate time for the phase change start. The tapering off at the right is due to the series mode crystal as used in Fig. 22. The reference phase begins to track the input phase.

Adding a low pass filter and boost diode will clean up the waveform seen in Fig. 33. Values are for a 48 MHz IF.

Fig. 35. Post Detection Low Pass Filter and Boost Diode.
Figure 36.
Fig. 36. Detected waveform after two stages of near zero group delay filtering. This is the XOR detector output with the waveform passed through the low pass filter of Fig. 37, with boost diode. The first down going spike to have enough level change to trigger the decoder will result in a detected digital one. There are no spikes for a digital zero. (approximately 1 volt per division, hence a rail to rail output equivalent to 180 degrees of phase shift is seen).

Second and 3rd harmonics can be troublesome with zero group delay bandpass filters. The low pass filter shown below in figure 37 is effective in removing them. The 3dB roll off is set half way between the fundamental and the 2nd harmonic.

![Diagram](image)

Fig. 37.

**Summary Comments:**
Figure 38 shows the detected phase change level (top) vs the loss in phase (calculated) that would result if the Fourier \( \sin x/x \) sideband products resulting from ultra-narrow band modulation are assumed to cause the PM as Bessel products do. Special ultra-narrow bandpass filters as described above reduce the Fourier sidebands with very little loss in transmitted carrier phase shift. The assumed phase angle loss as calculated for Bessel products is compared to the actual measured loss with Fourier products. Obviously, Fourier amplitude \( \sin x/x \) sidebands do not have the same effect as Bessel sidebands and are not necessary to sustain the phase change in modulation. The Overtone TRS filter has less measured phase loss than the values shown for fundamental crystal filters. See Figure 9.

**PM theory** says the phase angle should be equal to \( 2J_1 = \sin \Phi \ (J_0 = 1.0) \) for small phase angles [Hund, 21]. The ultra-narrow band modulation methods result in a large detected \( \Delta \Phi \) that completely disregards any changes in AM sideband levels. There are no \( J_n \) Bessel, or equivalent, sideband products with MSB modulation, and Fourier \( \sin x/x \) products, which are amplitude products, do not change the detected phase angle by any measurable amount.

Zero group delay IF filter Designing for Ultra Narrow band methods can be very difficult (and annoying) work. Often the designer finds it is difficult to repeat the last experiment. The boards can oscillate due to feedback on the board, or leakage from an adjacent board. The circuits given here have been used repeatedly for some time now and any failure is due to poor layout, or impedance matching. Working at 80-90 MHz is much more difficult than at 10-25 MHz. Some circuits work well at 10 MHz, but fail at any higher frequency. Crystals are very complex devices, especially in the overtone
mode. Low pass filters should be used with all MSB modulation methods to remove 3rd and higher harmonics before filtering.

Bipolar transistors have a great deal of feedback between the collector and the base and emitter. They can detune the entire circuit readily with a small change in load (or level). Emitter followers are poison in that they pass the output load characteristics to the circuit ahead of the input at the base and can de-tune the filter. Coupling to the emitter will also raise the shoulder level. Bipolar transistors are also notorious for cross modulation and the production of harmonics. (See Rheinfelder [2] and Hausman [12])

One commonly encountered problem is obtaining crystals that can be tuned reliably to a single frequency within about 160-200 Hz. Frequency tolerance is a serious issue and most overtone crystals do not rubber very well. See note below on tunability. Adding series capacity will raise the crystal frequency and adding inductance will lower it (Fig. 15). This broadens the available tuning range. Using an inductance in series with the crystal usually allows adequate tuning range, but the crystal frequency may be as much as 8-10 kHz below the nominal, or cut, value for 20pf shunt capacity. In some cases, using a series cut crystal will come out on the correct parallel mode when a series inductor is used. The detected waveform can be distorted if both L and C are used at the same time.

These cascaded IF stages can end up with a 3dB bandwidth of 1 kHz, or less, so little or no drift can be tolerated. AFC circuits have been designed to hold the UNB signal within the filter bandwidth. Using capacitors that have temperature characteristics to match crystal drift with frequency could be used.

Note that scope probe capacity across an output may also alter the group delay observed. This effect can be reduced by using low Q ferrite snubber beads as inductive loads and transformers

Remember SNR = $\beta^2 C/N$. In this equation it is the phase change $\beta$ that survives the filters relative to the reference that counts. The bit error rate is determined from:

$$P_e = \frac{1}{2} \text{erfc} \left[ \frac{SNR}{\sqrt{2}} \right]$$

This is the equation for 2 levels. For a single frequency as in VMSK or MSB, there are indications the formula should be $P_e = \frac{1}{2} \text{erfc} \left[ 2SNR \right]^{1/2}$. (Bellamy, [20]).

Reducing the shoulder level response of the filter, which reduces the level of any Fourier sidebands in the transmitted spectrum, does not necessarily reduce the detected phase angle.

General Data on Crystals:
Fig. 39. Equivalent Circuit of a Crystal Resonator.

\[ F_s = \frac{1}{2\pi} \sqrt{\frac{1}{C_s L_s}} \]
\[ \frac{\Delta f}{F_s} = \frac{C_s}{2(C_p + C_L)} \]
\[ Q = \frac{1}{2\pi R_s C_s} \]
\[ Q_L = \frac{2\pi F_s L_s}{R_{load}} \]
\[ \tau_1 = R_s C_s \approx 10^{-14} \text{ sec.} \]
\[ \frac{\Delta \phi}{\Delta f} = \frac{360}{\pi F_s} \frac{Q}{Q} \]
\[ \Phi \approx \frac{\omega L_s - \frac{1}{\omega C_s}}{R_s} = 0 \text{ at } F_a \text{ and } F_s \]

Some typical values are:

<table>
<thead>
<tr>
<th>Freq.</th>
<th>Mode</th>
<th>Rs</th>
<th>Cp (pf)</th>
<th>Cs (pf)</th>
<th>L (mH)</th>
<th>Q</th>
</tr>
</thead>
<tbody>
<tr>
<td>10MHz</td>
<td>Fund</td>
<td>8</td>
<td>3.5</td>
<td>.018</td>
<td>14</td>
<td>109,000</td>
</tr>
<tr>
<td>20</td>
<td>&quot;</td>
<td>15</td>
<td>4.5</td>
<td>.020</td>
<td>3.1</td>
<td>26,000</td>
</tr>
<tr>
<td>30</td>
<td>3rd</td>
<td>30</td>
<td>4.0</td>
<td>.002</td>
<td>14</td>
<td>87,000</td>
</tr>
<tr>
<td>75</td>
<td></td>
<td>25</td>
<td>4.0</td>
<td>.002</td>
<td>2.3</td>
<td>43,000</td>
</tr>
</tbody>
</table>
Note on Tunability. [Ref. 17] Frequency pullability (tuning the filter) (change in frequency per picofarad) is limited to less than ±0.2%, with typical values of 10 ppm to 500 ppm. This limitation is due to the very high Q values of the crystal, typically between 10,000 and 100,000. Crystals operating in the fundamental mode can be pulled typically 100 ppm with a capacitor used in series with the crystal. When a third overtone crystal is used, the pullability is reduced to 10 ppm or less. **If an inductor is placed in series with the crystal, pullability/tunability is increased considerably. In many cases, using a variable inductance is the only way to obtain the desired tuning range.** (See Figs 16 and 17.) The crystal is less tunable as the frequency is shifted toward series resonance. (almost no tunability is obtainable at series resonance $F_s$ (Fig. 40) or (Fig. 6 ‘a’) and maximum tunability at the anti-resonant point Fig. 40 $F_a$ (Fig. 6 ‘c’)).

![Diagram of Crystal Resonator Characteristics](image)

Figure 40. Crystal Resonator Characteristics. (Data above is from Piezo Crystal Company).

![Diagram of Cascaded Shunt Filters](image)

Figure 41. Cascaded Shunt filters once used with Cable TV and other transmitter units to obtain maximum shoulder rejection. The Inductor in series with crystal is necessary to obtain adequate tuning range. Typical value is the inductance required to resonate with 30 pf.
It has been found the TRS filter group is easier to use and has better shoulder reduction with less phase loss. The Cable TV unit operated at 60 MHz and therefore the crystals were necessarily 3rd overtone.

In earlier versions of this paper the integrating filter was incorrectly referred to as a correlator. Actually it is a following part, or addition to, the correlator circuit. Correlation occurs when two waveforms are multiplied (mixed). To obtain a detected output in practice, the sum and difference of the two signals are separated, usually preserving only the difference, which contains the desired signal information.

This combination of correlating multiplier and integrating filter, results in an optimum filter for CPFSK methods, which is identical in performance to a matched filter.

Figure 1 assumes the multiplication has already taken place, or that the input is a baseband waveform.

**Correlation** is used here in the sense that it comprises a multiplier (detecting mixer for \( v_1 \times v_2 \)) with the result of the signal plus local oscillator multiplication being integrated over a full bit period. (Ref. [8]).

\[
R_{12}(\tau) = \frac{1}{T_0} \int_{-T_0/2}^{T_0/2} v_1(t)v_2(t+\tau)dt,
\]

Is the general equation, which becomes:

\[
R = \int_0^T v_1(t)v_2(t+\tau)dt,
\]

when the time \( t = T_s \). \( T_s \) is both the bit period and symbol period, and \( t \) is the detected number of cycles being integrated. NRZMSB introduces a phase change for the entire bit/symbol period for a digital one (\( t = T_s \)).

Figure 42. NRZ-MSB Cable TV modulator for 90-120 degree modulation.

The balance pot. enables the level of phase 1 or Phase 2 to be varied to obtain the least phase angle loss. The TRS filter shown is adjustable for phase and frequency. The least phase angle loss results in an amplitude unbalance. This can be corrected in the second
stage by using the opposite phase. The filter should be tuned slightly above the
transmitted frequency for the least phase loss.

![Phase Detector Diagram]

Fig. 43. Phase Detector for use with the Modulator above. See also Fig. 32.

Based on experimental information, the inductor to be used in series with the crystal
in Fig. 14 to get the peak AM response and best group delay at the same frequency
is that inductance that resonates with the specified shunt capacity. This is usually
around 18-20 pf. 30 pf is also acceptable.
NOTE: Some chip inductors do not function at all. The reason is unknown, but it is
suspected the inductor must have a high Q.

**MODE CONVERSION:**

Some of the filters described above can be used to convert Ultra Narrow Band
modulation to a missing cycle modulation equivalent as part of the receiving process
( MCM ). **This is often desirable.** At the phase change edges, a timing gap is created
that looks like a missing cycle. The detectors look for this gap as well as the phase
change that follows, since the gap is equivalent to a 180 degree shift. **This is normal for
all modulation methods that do not use CPFSK ( BPSK, BFSK ). MCM is not normally
being used at the transmitter, but can be used advantageously when down
conversion is involved. It has been used successfully in an FM SCA transmitter.** The
phase detectors used can be made to respond to missing cycles, or to the phase change
gap, as well as to the phase modulation change that follows.

The vector relationship is seen in Fig. 12. If the reference vector is in phase with
( coherent to ) one of the signal vectors, as in Fig. 12, the output will be nulled, or
reduced, for the opposite signal phase, thus converting PRK or PSK from the transmitted
source to an MCM or AM equivalent.

The objection to this is that missing cycle modulation, which is AM, cannot be used with
a limiter and the BER will not be as good as it is for phase reversal or phase shift keying.
When passed through an off tuned filter, MCM can be converted to PM, which can be
limited. Limiting should be used after the bandpass filters.
The optimum BER is obtained when the total phase loss is kept below 50% (±45°) degrees with ±90 degrees input.). The maximum noise effect angle is $\arctan = 1/\text{SNR}$, so that a 1/1 SNR (0 dB) causes a 45 degree phase change. Transmitted MCM can be expected to be 6 dB worse.

One solution is to keep the phase changes intact ahead of the limiter, but use a post limiting filter that changes the mode.

The point of the above comment is that the waveform at the output of the limiter should be examined carefully for phase and timing gaps, then a phase detector type chosen accordingly. A true phase detector (Fig. 32) with an output proportional to phase may not be the best choice in a noisy environment, while a bang-bang type such as the D flip flop (Fig. 43) might be a better choice if the inputs are properly phased.

References:

It will be noticed that much of the reference data concerning LC filters is 30-40 years old or more. Some of it goes back to WW1. The industry has largely converted to crystal or SAW bandpass filters, or to DSP/FIR filters at baseband for other uses. Unfortunately, the group delay of these filters renders them useless for ultra narrow band methods. Low pass and high pass filters have near zero group delay in their flat bandpass regions.


[2] and [3] contain excellent sections on cascading LC amplifiers, group delay (envelope delay), rise time and phase shift (phase delay). They explain that group delay is NOT always according the conventional filter models.

This Section is mostly repetitive of traditional information.


- R effect, Sect 9.5, PLL Sect. 10.7, Narrow Band Noise Sect 7.5, --Correlator pp 454.


McGraw Hill. -- Rise time, BW etc. pp79.


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