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Discrete Synchronous Detection for QuAM & ISB

Before we proceed into how this is done a good history of DSB + Carrier, Envelope Detectors, C-QuAM, and C-ISB QuAM, ISB, and Synchronous Detectors is in order and why these four two channel modes are similar and what makes this easy to do with the MC13020 chip. While the other Motorola AM Stereo chips are capable of the C-ISB mode they are not really adaptable to operate in synchronous ISB mode.

§ DSB + Carrier – Double Sideband with Carrier. This is the simplest method of modulation only requiring rectification with a simple diode to convert the envelope of the signal to the audio baseband. It is very inefficient in that only up to half of the transmitted energy is in the sidebands while the other half is always dedicated to the carrier. This is the case for Mono AM, C-QuAM, and C-ISB. Instead of developing the BFO in the receiver to demodulate the signal and in which all the transmitted energy would be in the sidebands the carrier is sent along with the modulation to be the BFO robbing the transmitter of half its power, power that could be dedicated to the modulated signal for abetter signal to noise ratio and much less power usage.

§ Envelope Detector – This type of detector is the simplest of all detectors but requires that all information be in the in-phase channel and none in the quadrature channel for distortion free reception. This also requires that the information in the upper and lower sidebands maintain their proper phase and amplitude characteristics in relation to the carrier. During skywave, selective fading, and multi-path conditions these relationships can become distorted and the coherent information that once was only contained in the in-phase channel can now be found in the quadrature channel. Interfering signals from other stations and natural sources are incoherent in relation to the desired carrier and produce information in both the in-phase and quadrature channels of the desired signal and will produce the same kind of distortion known as “quadrature distortion.” This is the inherent weakness in the signal's envelope in which this detector demodulates.

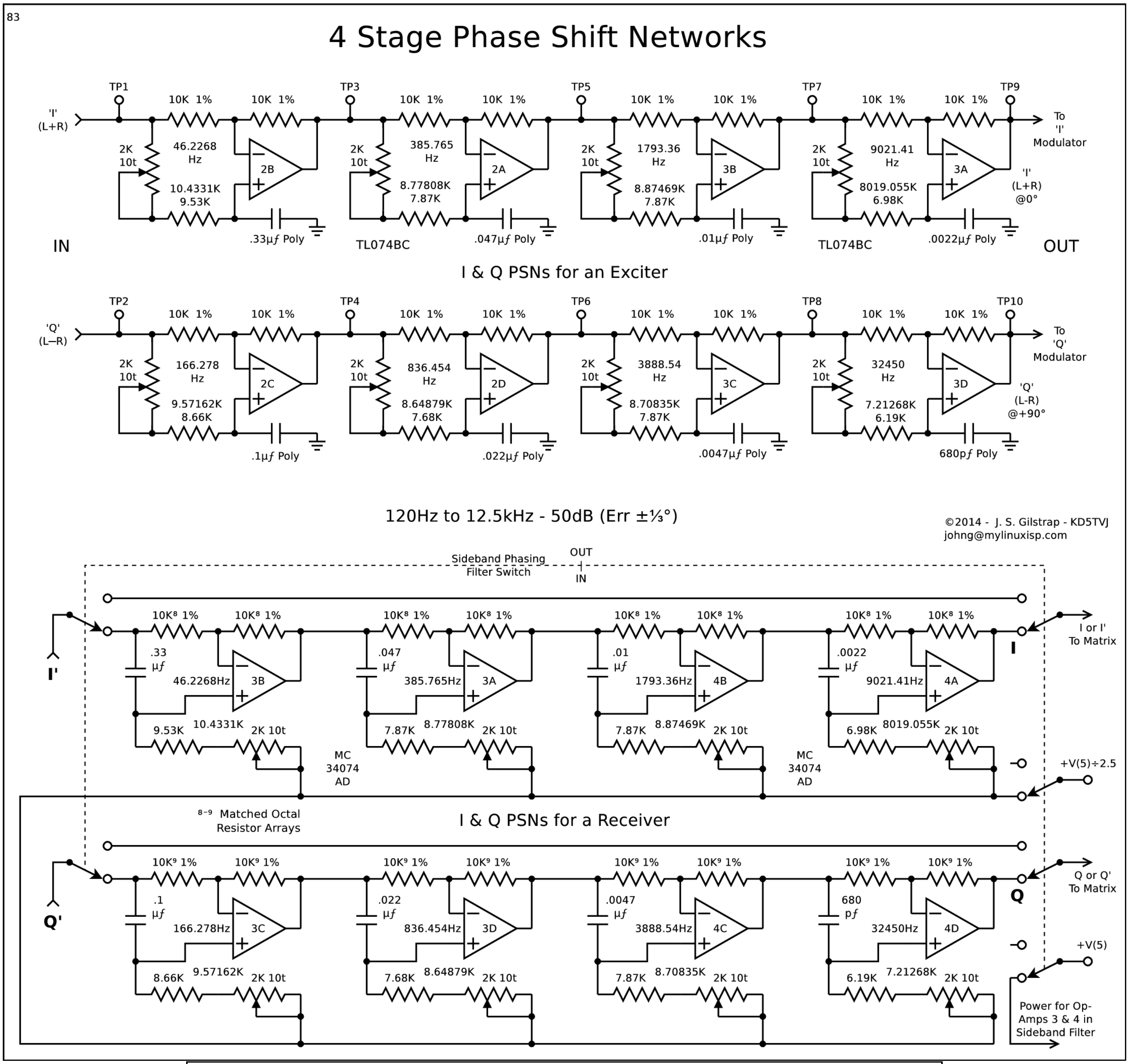
§ Synchronous Detector – This type of detector does not require a carrier other than the purpose of having the BFO that controls it is frequency locked to the carrier. This is what makes it synchronous because the phase angle of the BFO is synchronized at a certain angle with the carrier. This detector only demodulates the signal for the channel that the BFO driving it and is phase-synced to, be it the in-phase, quadrature, or somewhere in between and completely eliminates the channel 90° away from it. Since it ignores the channel 90° away from it it is immune to quadrature distortion and this also allows for two channels to be transmitted on one carrier known as QuAM while only occupying the same bandwidth as a DSB mono transmission.

§ QuAM – Quadrature Amplitude Modulation is an amplitude modulation mode that allows the transmission of 2 different signals on the same carrier. One of its widespread uses is the transmission of the color information along with the black & white information of an analog color television signal for both the NTSC and PAL and their numerous variations. This was done in suppressed carrier mode. There is no carrier at all to demodulate the signal but a short burst of a carrier signal without modulation during retrace for the PLL that controls the BFO to lock onto. These are two carriers on the same transmission frequency but are separated by 90° as in Sine & Cosine i.e. when sine is 0 then cosine is 1 and vice-versa. This allows the two signals to be separated and demodulated with synchronous detectors that are driven by a BFO that is frequency and phased locked to the reference signal via a PLL circuit. In the case of color TV this is the color burst signal and for AM radio it is the carrier. For QuAM detection very little carrier is need to allow enough for the PLL to lock onto. As little as ⅛ the power required for envelope detection is sufficient for PLL locking.

§ ISB – Independent Sideband is a mode in which the two signals transmitted are separated by frequency and not by phase as it is in a QuAM signal. QuAM modulation is used to generate an ISB signal but the audio is phase encoded in a way to cause one of the sidebands to be nulled per channel. After the audio is matrixed from L & R into L + R & L – R it is audio phase shifted to produce a 90° differential phase shift between the matrixed channels. This produces sidebands generated from (L+R) w/f@0° & (L – R) w/f@+90° modulating the I & Q modulators and after summing they produce the distinct L & R channels in the lower and upper sidebands respectively. Through trigonometric proofs this can be demonstrated. In its simplest definition: Let A=audio and C=carrier, then USB = Cos(A) × Cos(C) + Sin(A) × Sin(C) & LSB = Cos(A) × Cos(C) – Sin(A) × Sin(C). During good signal conditions via ground wave there isn't much difference other than a 3dB reduction in noise and as background noise increases this provides a slight benefit. Where ISB really excels is when the transmission path via skywave and/or multi-path distorts the signal by introducing non-linear phase and amplitude effects within the passband of the signal and ISB has the best immunity out of all the other modes. With a non-ISB QuAM type of signal when the natural amplitude and phase relationship of the two modulations as they are distributed in both sidebands is compromised channel separation is reduced and under more severe skywave effects this produces noticeable audio phasing effects and even signal nulling. With co-channel interference where the interfering signal causes the PLL to mis-track a QuAM signal will suffer from channel mixing and platform motion where the dominant signal content will bounce between the two channels at a frequency rate that is the beat frequency of the desired and un-desired carrier. With ISB this PLL mis-tracking error can occur but since the signals are separated by frequency and not by phase, each channel is contained within its own sideband, and there is no special amplitude and phase relationship between both channels and sidebands as they are in a QuAM signal so the signal can't be compromised in this manner. For ISB during PLL mis-tracking and selective fading during skywave phasing effects play themselves out in a much less severe form and in most cases phase distortion within the audio passband of the signal caused by moderate PLL phase mis-tracking is not readily noticeable to the human ear. It is the differential phase distortion between upper and lower sidebands that produces signal nulling in a DSB or QuAM signal similar to a comb filter. A perfect example of this is a guitar phaser effect. Within these circuits the guitar signal is separated into two paths where one path is phase shifted in which the phase is also slowly modulated and mixed back in with the original signal to produce a differential phase mixing.

§ Audio Phase Shift Networks (PSNs)– In the next drawing are the phasing filters for ISB. These are 4 stage PSNs that have a ±⅓° phase deviation from 90° and a theoretical -50dB opposite sideband suppression capability. These are best used for High Fidelity ISB Stereo with a 50Hz to 15kHz audio response. Its -50dB attenuation covers a range of 120Hz to 12.5kHz and at 100Hz and 15kHz it still has a fairly respectable -35dB suppression for stereo programs. For the exciter the PSNs are phase lagging while the ones for the receiver are phase leading. If the exact same frequencies used in the exciter are used in the receiver then the ±⅓° phase deviation from 90° is also negated for a 0° phase error correction thus returning the modulating signals to their original phasing and separation before transmission. To produce a 90° phase difference between the I & Q modulation signals in analog mode a running phase shift is needed. At the lower and upper end of the range ~59°/oct. is produced while in the middle it is ~73°/oct. This rolling phase shift has the benefit of shifting harmonic peaks off their fundamentals reducing p-p values for a given signal level thus allowing a louder signal to be modulated for the same p-p level. Whether to use phase leading in the exciter and lagging in the receiver contrary to what is previously stated would depend on which would produce the greatest reduction in p-p value for a given signal level on most program material during transmission. In the digital domain just applying a Hilbert transform to the 'Q' channel only would not produce this effect but might be detrimental to 'Q' channel peak levels e.g. a square wave. For best results high quality metalized film resistors should be used and polystyrene capacitors if available or polypropylene. The resistors that form the unity gain inverted signal in the op-amp should perfectly matched pairs whether it be 10K, 15K, 22K, etc... the values don't have to be the same from op-amp to op-amp but each pair for each op-amp should be matched.

There are test points TP1-TP10 in the exciter PSN section. These test points are there to calibrate each phasor within the array at its defined frequency. Using a function sine wave generator with an optimum 6 digit frequency accuracy set it to the desired frequency for the phaser to be calibrated. Inject the signal into the array at the input and place a quadrature phase detector on the test points associated with the input and output of the phaser section to be calibrated. Adjust the phaser until the output of the phase detector is zero. Repeat this for the remaining nine sections using their defined frequency. By connecting the phase detector to the final output of the I & Q PSNs and injecting the signal into both inputs the frequency test points at the ±⅓° peaks are: 156Hz, 283Hz, 606Hz, 1.225kHz, 2.575kHz, 5.315kHz, & 9.65kHz can used for further checking. If equipment is available running a full frequency sweep with the phase detector is a good way to check for full accuracy and perform fine tuning if necessary. Test points are not present in the PSNs for the receiver but this same process can be used to calibrate them also. In the drawing after the PSNs there is a Quadrature Phase Detector to calibrate the PSNs.


83
4 Stage Phase Shift Networks
I & Q PSNs for an Exciter
'I' (L+R)
'Q' (L—R)
IN
OUT
TP1 TP2 TP3 TP4 TP5 TP6 TP7 TP8 TP9 TP10
10K 1%
2K 10t
46.2268 Hz
10.4331K
9.53K
.33µƒ Poly
385.765 Hz
8.77808K
7.87K
.047µƒ Poly
1793.36 Hz
8.87469K
7.87K
.01µƒ Poly
9021.41 Hz
8019.055K
6.98K
.0022µƒ Poly
166.278 Hz
9.57162K
8.66K
.1µƒ Poly
836.454 Hz
8.64879K
7.68K
.022µƒ Poly
3888.54 Hz
8.70835K
7.87K
.0047µƒ Poly
32450 Hz
7.21268K
6.19K
680pƒ Poly
TL074BC
To 'I' Modulator
'I' (L+R) @0°
To 'Q' Modulator
'Q' (L-R) @+90°
120Hz to 12.5kHz - 50dB (Err ±⅓°)
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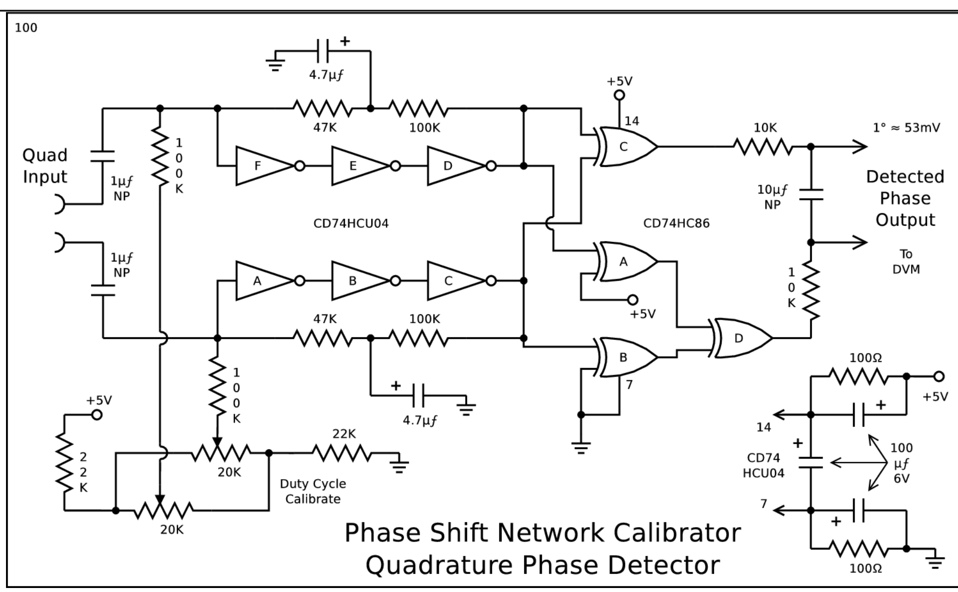
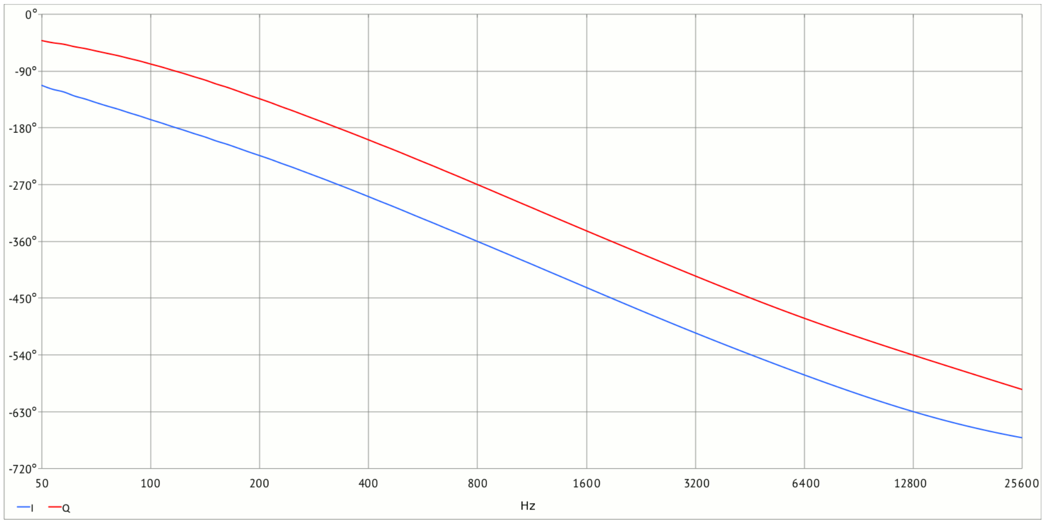
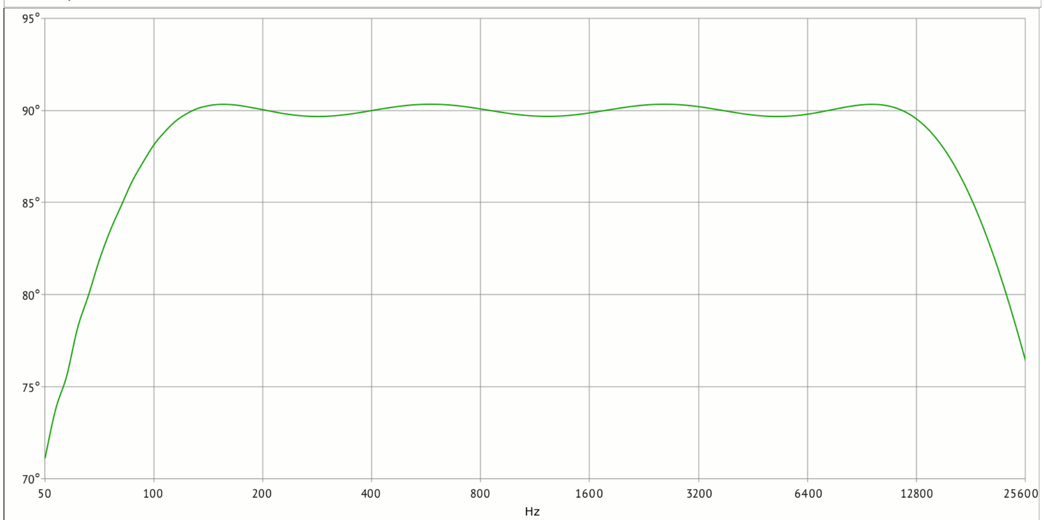
johng@mylinuxisp.com
Sideband Phasing Filter Switch
OUT
IN
10K⁸ 1%
10K⁹ 1%
I'
Q'
I
Q
I or I' To Matrix
Q or Q' To Matrix
MC 34074 AD
+V(5)÷2.5
+V(5)
⁸⁻⁹ Matched Octal Resistor Arrays
I & Q PSNs for a Receiver
Power for Op-Amps 3 & 4 in Sideband Filter


100
Quad Input
1µƒ NP
100K
4.7µƒ
47K
100K
+5V
CD74HCU04
CD74HC86
10K
1° ≈ 53mV
10µƒ NP
Detected Phase Output
To DVM
22K
20K
22K
Duty Cycle Calibrate
100Ω
100 µƒ 6V
CD74 HCU04
Phase Shift Network Calibrator
Quadrature Phase Detector

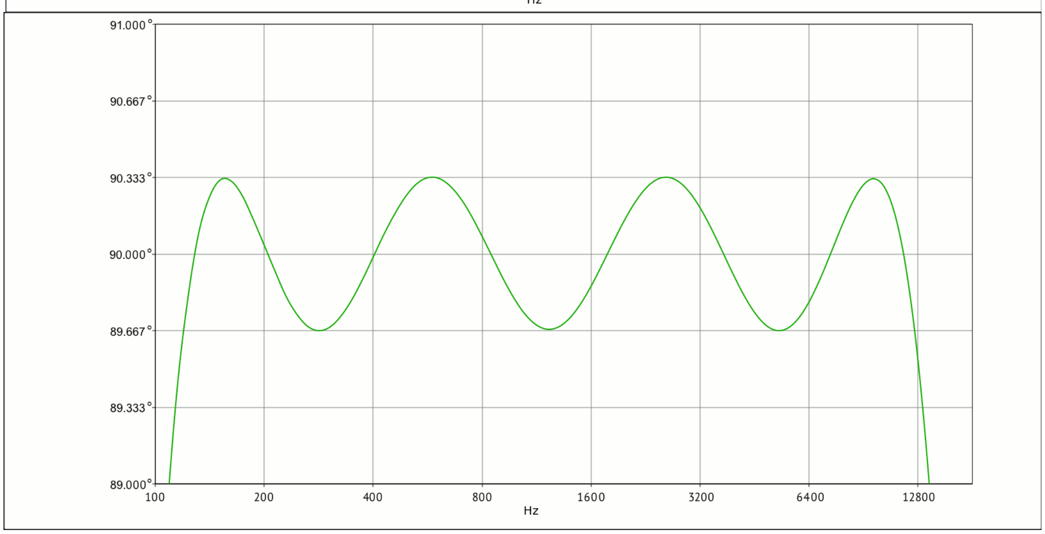
Rolling phase shift for both I & Q channels showing Q leading I by 90°.



Phase differential from 50Hz to 25.6kHz.



Phase differential close up showing ±⅓° accuracy.



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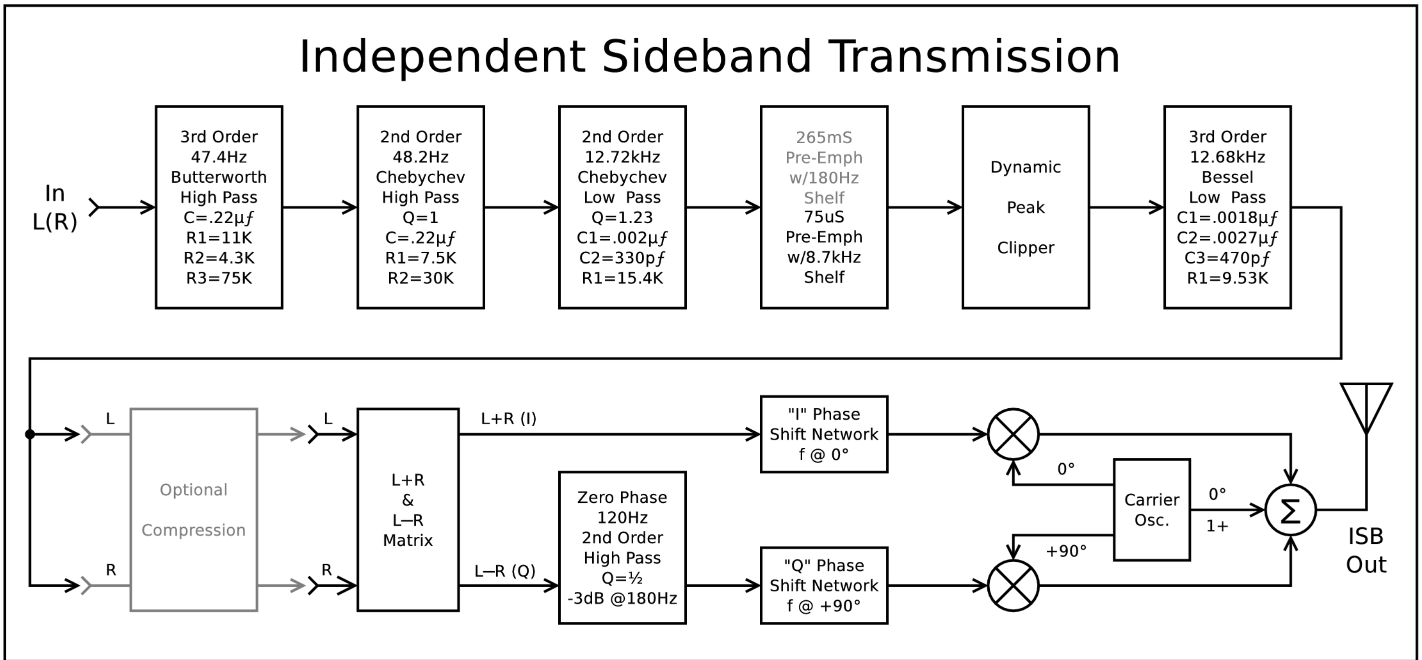
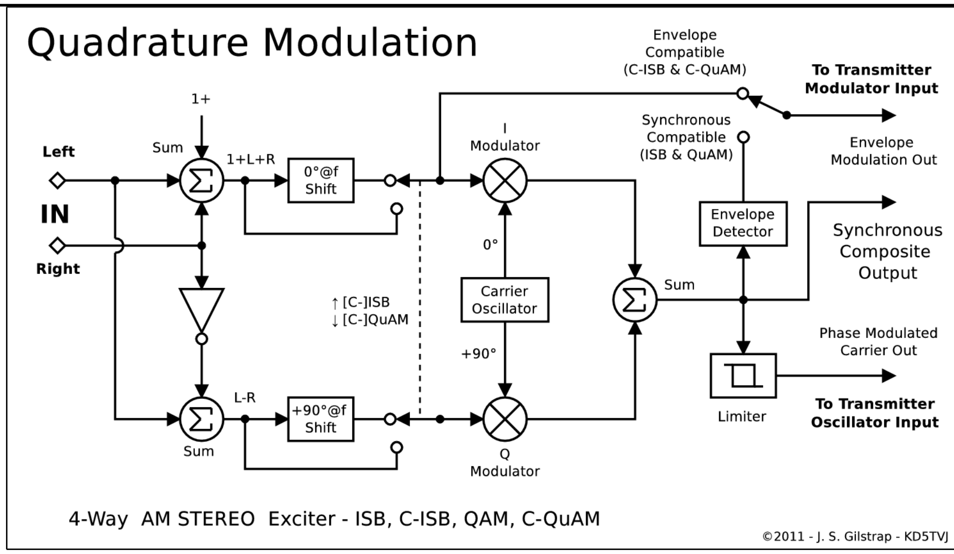
§ Exciter – This is a block layout and very simplified version of an exciter that could transmit four different systems. This article will not address 'Compatible' mode as C-QuAM & C-ISB but for the Harris 'Synchronous' mode and Harris made full 90° QuAM exciters and with the installation of the Sideband Phasing Filters will generate an ISB signal also.





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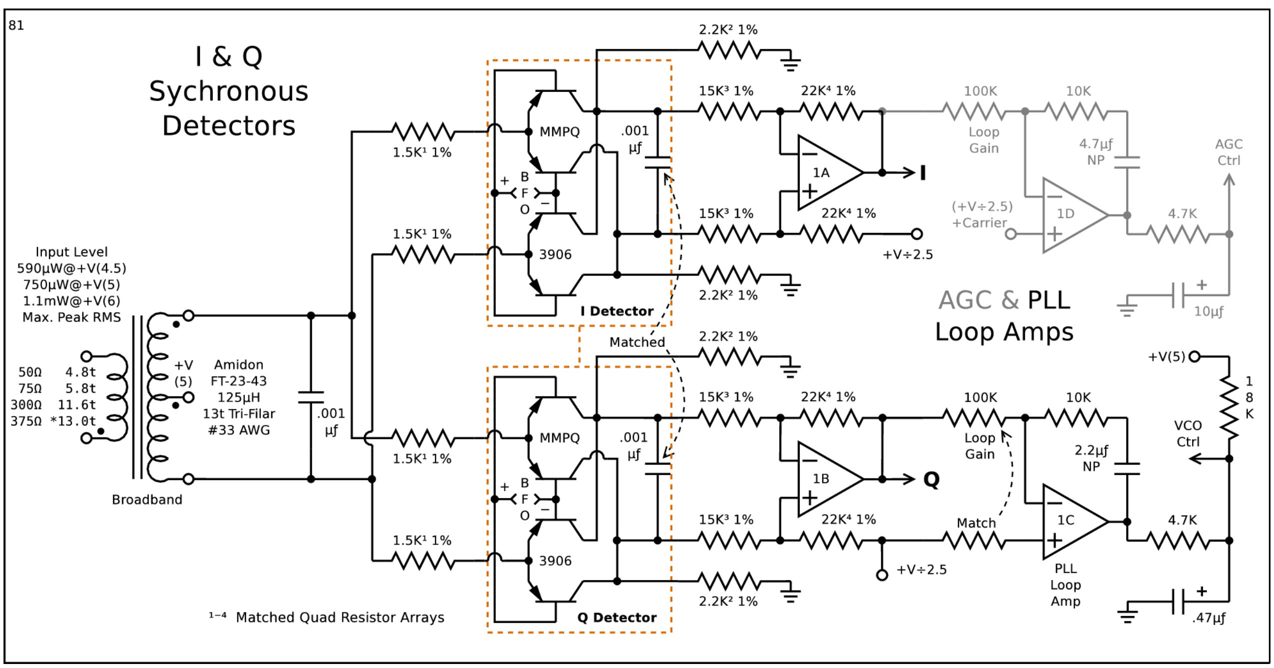
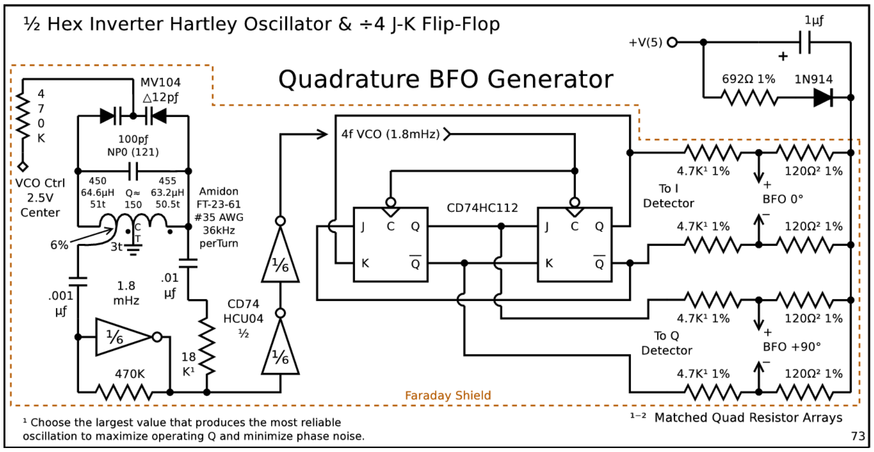
§ Discrete Component Detector – In designing a discrete component synchronous detector several goals are desired.

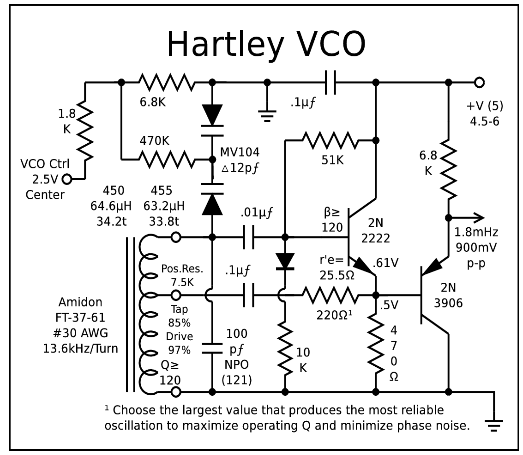
* Versatility in use in all kinds of radios and suitable for portable battery power, car, or home stereo operating from 4.5V to 6V.
* Uses low cost common off the shelf parts and an average number of coils and transformers that can be wound or bought.
* Circuit simplicity with a comprehensive and well balanced design that requires little adjustment.
* Low parts count for a discrete component circuit utilizing quad arrays to reduce cost and manufacturing complexity.

In the next drawing is an I & Q synchronous detector that utilizes quad transistor arrays for the upper half of a Gilbert cell but the two transistors in the lower half have been replaced with resistors and connected to a tuned or broadband tank that is a phase splitter and at a 5V potential. The four transistors are are biased at less than 5V so the resistors will have a voltage drop across them to supply the bias current to the transistors instead of the two transistors in the lower half of a Gilbert cell. Since this circuit operates on low voltage and would require the need for current mirrors to supply the current modulations to the upper quad transistors, eliminating this provides more headroom for the quad transistors allowing larger voltage swings in this part of the circuit. When the tank is energized the p-p voltage swing will develop an AC current imposed on the existing DC bias in the resistors which supplies the emitters, which are at an almost constant voltage, with the proper current modulations. During switching there will be a ~40mV spike in voltage across the resistors which results in a slight current change of short duration from both pairs during the switch. This is of no consequence since the tuned tank if narrow band will filter this out and for broadband the these current spikes are also introduced into the coil in common mode which rejects them especially if the coil is wound bi-filar. The bi-filar winding also neutralizes the DC magnetizing force from the voltage drop across the resistors. The ferrite core will only have AC flux through it for maximum signal handling and no DC flux to diminish this. The parallel resistance across the coil is defined by the four resistors of equal value that are in a series-parallel in relation to the p-p connection to the coil so the resistor value itself defines the parallel resistance. Defining the bandwidth is done by selection the L & C values so their reactance in conjunction with the parallel resistance will give the desired Q. Using a matched quad resistor array ensures that both detectors get the same DC bias and AC signal currents since they are both taken off from the same taps of the coil. In this drawing a third winding of various turns depending on what input impedance is desired is where the signal is injected. The other method of signal injection is a differential amplifier with the collectors connected to the coil in a push-pull center-tapped coil style creating a folded Gilbert Cell with a tuned input. Again since the tank is wound bi-filar the DC current from the drive transistors will be canceled out. The output of the quad transistor array has its collectors cross coupled and connected to 2 resistors to ground. There is a capacitor between the two resistors and collector outputs to filter out the IF portion of the signal to recover the baseband signal. Again these four resistors for the detector output should be a matched quad array providing accuracy in both DC and AC performance. This full wave rectified signal is balanced and when connect to an op-amp balun provides superior DC performance particularly in regards to the Q detector where the PLL loop amp gets its signal from so it should have tight phase tracking with little to no need for adjustment. Again the use of two quad resistor arrays for the two balun op-amp will optimize both DC and AC performance. The op-amp chosen is from the Motorola MC3407[24] series with a common mode input that includes Vee which is well suited for low voltage single supply operation especially in this application. The performance is of hi-fi quality with a 13V/µs slew rate and a 4.5mHz gain bandwidth product perfect for high performance QuAM or ISB processing working well up to a 100kHz detected signal. The precision LTC105[13] with 5µV offset, 4V/µs slew rate, and 2.5mHz GBWP offers superior DC PLL phase tracking without adjustment. The low power MC3317[24] with 2.1V/µs slew rate offers lower performance but still adequate for full frequency audio at these output levels while the LM 358/324 units from National Semiconductor has similar common mode input range but only has a 0.5V/µs slew rate limiting it to lower frequency use and also needs a pull up/down resistor on the output to eliminate crossover distortion. It is probably good for 5kHz audio but its performance is a bit short for the full frequency. The drawing also shows the PLL loop amp connected to the Q detector which develops the control voltage for the VCO and a similar one for grayed out for the AGC loop amp.



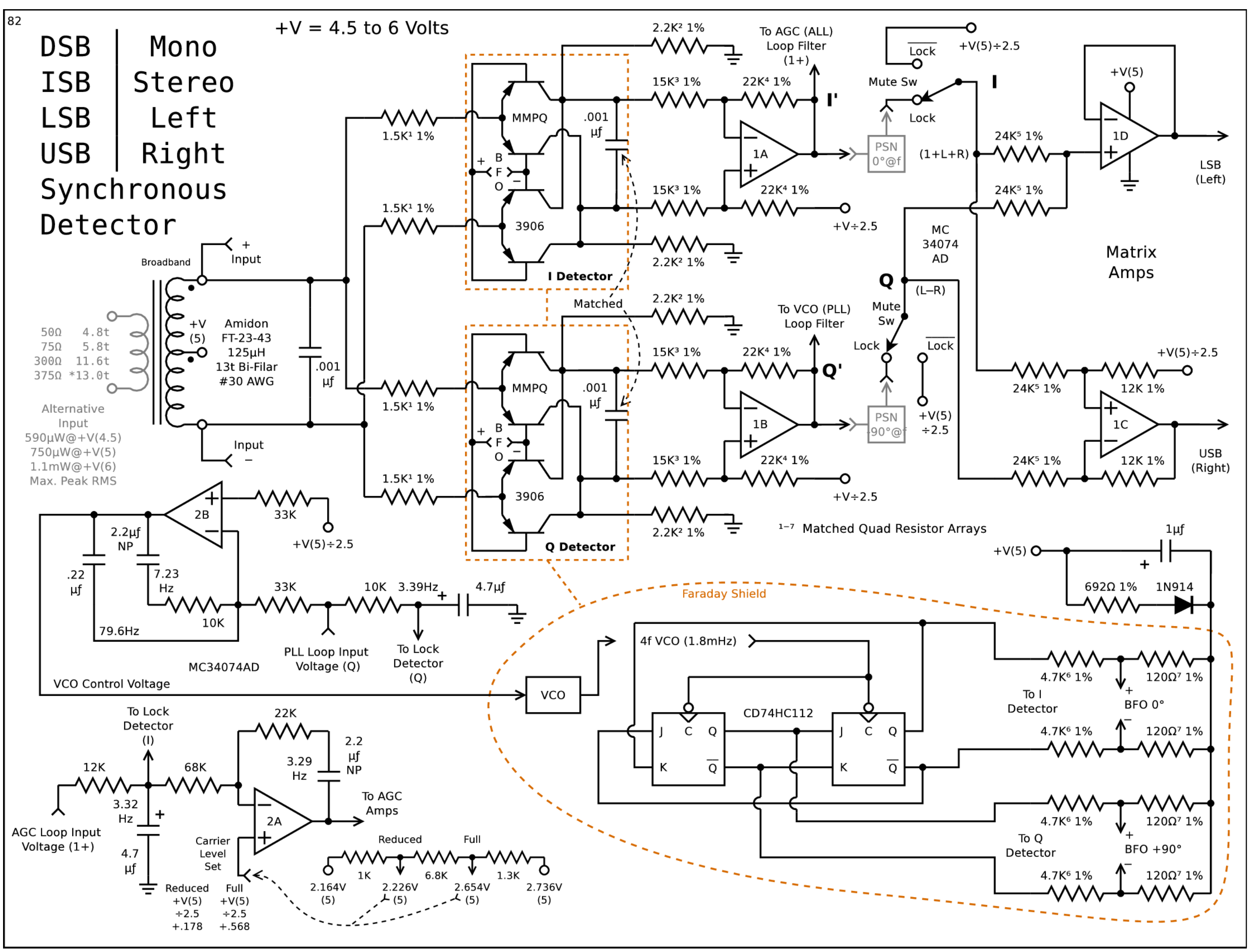
In the next picture is the VCO and the ÷4 JK Flip-Flop which generates perfect square waves in quadrature with balanced outputs. The oscillator is built using a hex inverter and connection to the coil is of the Hartley configuration. There is a varactor diode to vary the frequency that is controlled by the voltage produced from the PLL loop amp. Other typical VCO configurations built around hex inverters utilizing crystal or ceramic resonators can also be used. These balanced outputs from the Flip-Flop are used to provide a balanced BFO signal to the detectors so during both states of the ± switching the emitters of the transistors will always be at the same potential after each time they switch. This is important so after each switched state there will be no switched current offset induced into the coil. With the supply voltage at 5V the Flip-Flop outputs are 4.8Vp-p and balanced between inverted and non-inverted outputs is 9.4Vp-p. This is way too strong for the input to the bases of the transistors so a voltage divider is needed. The p-p level needed at the input of the differential pair is ~240mVp-p so a 40:1 reduction is needed. This is done with the 4.7K & 120Ω resistor sets connected to all four of the Flip-Flop outputs. Again four of each resistor is needed so using two matched quad resistor arrays will guarantee that voltage reduction for each divider will be identical in both p-p level and DC bias. To raise the DC bias level for the transistors a 300Ω pull up resistor along with a diode, common to all four dividers raises the bias voltage to its appropriate level. There is also a transistor based Hartley oscillator that can be used if the other half of the hex inverter is not needed but its output level will need to meet the clock drive requirements of the F-F.

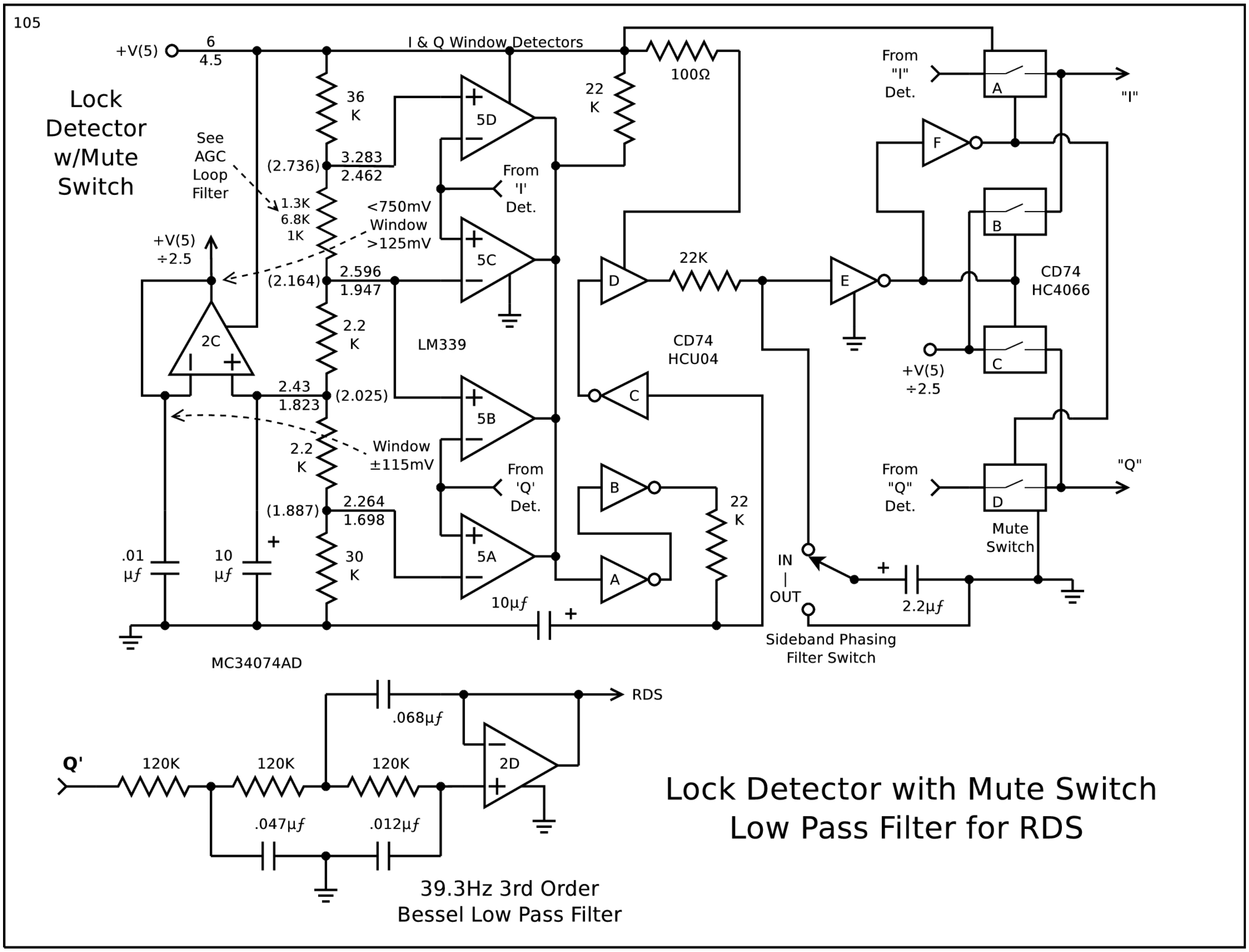


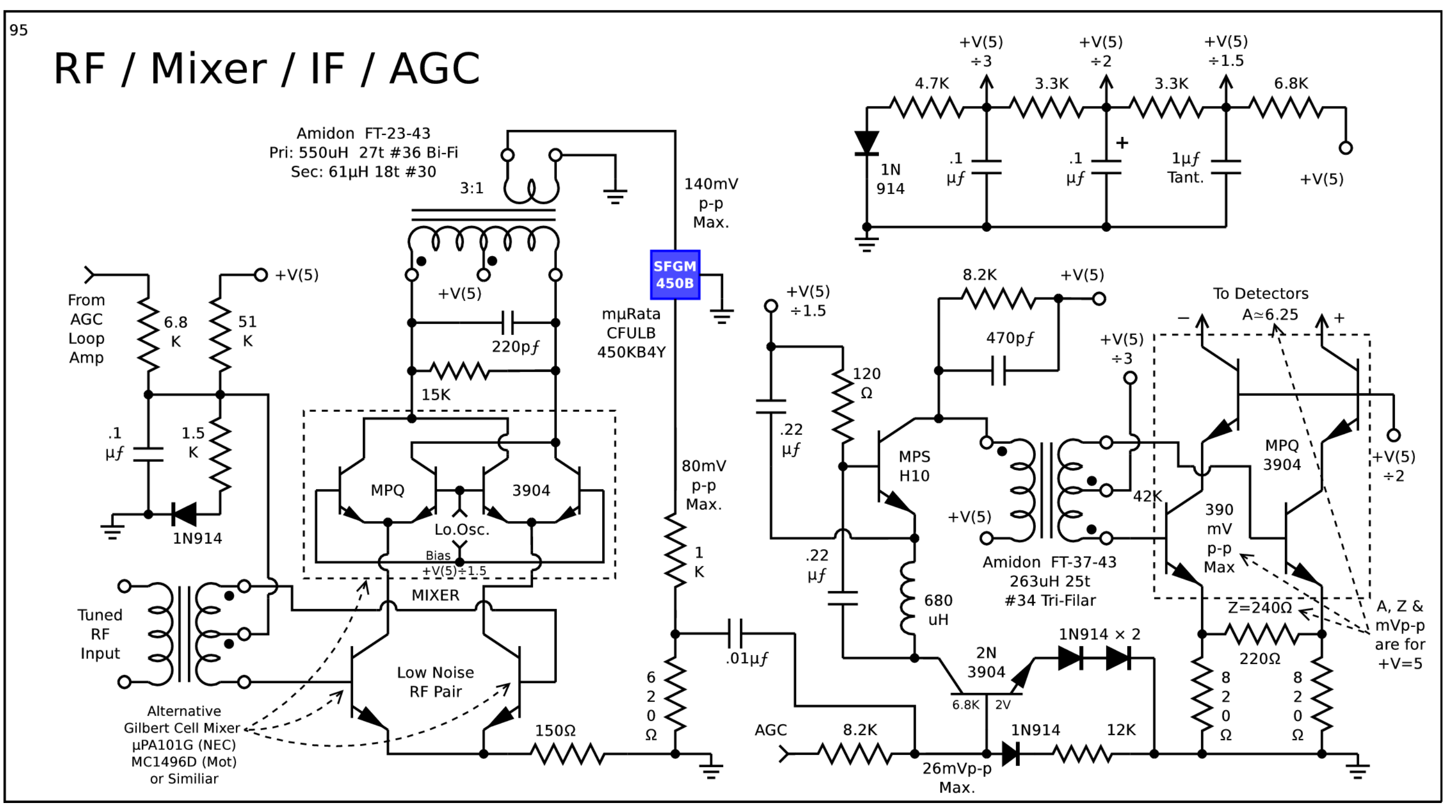


It is important to place a Faraday shield around the VCO, ÷4 F-F, and the detectors as square wave switching produces harmonic signals that may interfere with other equipment. If not then it is still good practice to do so.

In the next image are the synchronous detectors with various other support components e.g. block designation of the PSNs, de-matrix amps to develop LSB(Left) & USB(Right) signals, loop amps for both PLL & AGC, and low pass filtering for the I & Q signals for the lock detector. The F-F circuit has been repeated from the previous image but lacking the detailed VCO part designated by a block.



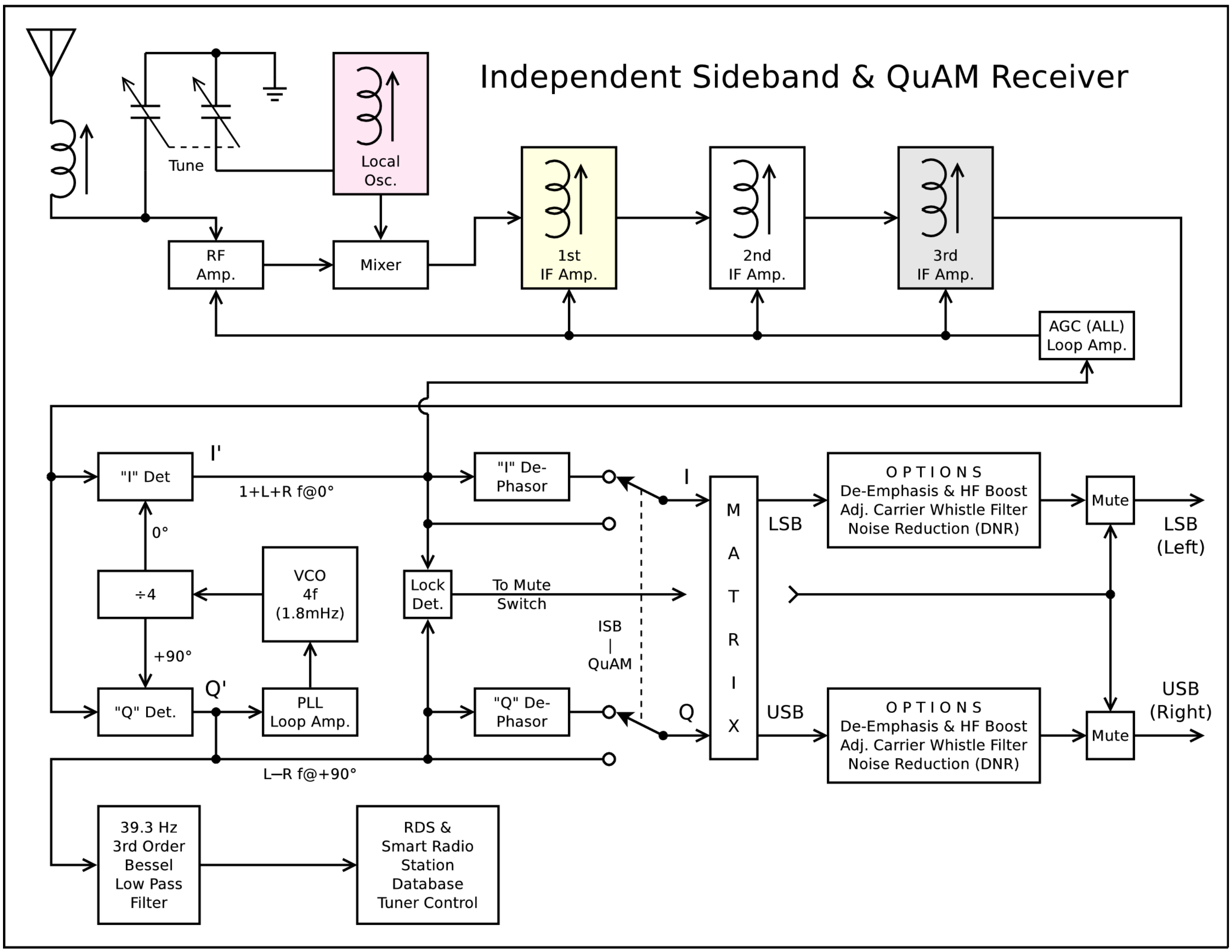

105
Lock Detector w/Mute Switch
+V(5)
6
4.5
I & Q Window Detectors
100Ω
From "I" Det.
"I"
36 K
22 K
5D
See AGC Loop Filter
(2.736)
3.283
2.462
From 'I' Det.
1.3K
6.8K
1K
<750mV Window >125mV
5C
+V(5) ÷2.5
(2.164)
2.596
1.947
22K
CD74 HC4066
2C
2.2 K
LM339
CD74 HCU04
+V(5) ÷2.5
2.43
1.823
(2.025)
5B
Window ±115mV
2.2 K
From 'Q' Det.
From "Q" Det.
"Q"
(1.887)
2.264
1.698
22 K
Mute Switch
.01 µf
10 µf
30 K
5A
IN | OUT
10µf
2.2µf
Sideband Phasing Filter Switch
MC34074AD
RDS
.068µf
Q'
120K
120K
120K
2D
.047µf
.012µf
Lock Detector with Mute Switch
Low Pass Filter for RDS
39.3Hz 3rd Order Bessel Low Pass Filter


95
RF / Mixer / IF / AGC
+V(5) ÷3
+V(5) ÷2
+V(5) ÷1.5
4.7K
3.3K
3.3K
6.8K
Amidon FT-23-43
Pri: 550uH 27t #36 Bi-Fi
Sec: 61µH 18t #30
3:1
1N 914
.1 µf
.1 µf
1µf Tant.
+V(5)
140mV p-p Max.
SFGM 450B
mµRata CFULB 450KB4Y
From AGC Loop Amp
+V(5)
6.8 K
51 K
+V(5)
220pf
15K
+V(5) ÷1.5
8.2K
+V(5)
470pf
To Detectors
A≈6.25
120 Ω
+V(5) ÷3
.1 µf
1.5 K
.22 µf
MPQ
3904
Lo.Osc.
Bias +V(5)÷1.5
MPS H10
MPQ 3904
+V(5) ÷2
80mV p-p Max.
1 K
42K
1N914
MIXER
+V(5)
390 mV p-p Max
.22 µf
Amidon FT-37-43
263uH 25t
#34 Tri-Filar
680 uH
Tuned RF Input
Low Noise RF Pair
Z=240Ω
220Ω
A, Z & mVp-p are for +V=5
6 2 0 Ω
.01µf
2N 3904
1N914 × 2
6.8K
2V
8 2 0 Ω
8 2 0 Ω
Alternative Gilbert Cell Mixer
µPA101G (NEC)
MC1496D (Mot)
or Similiar
150Ω
AGC
8.2K
1N914
12K
26mVp-p Max.

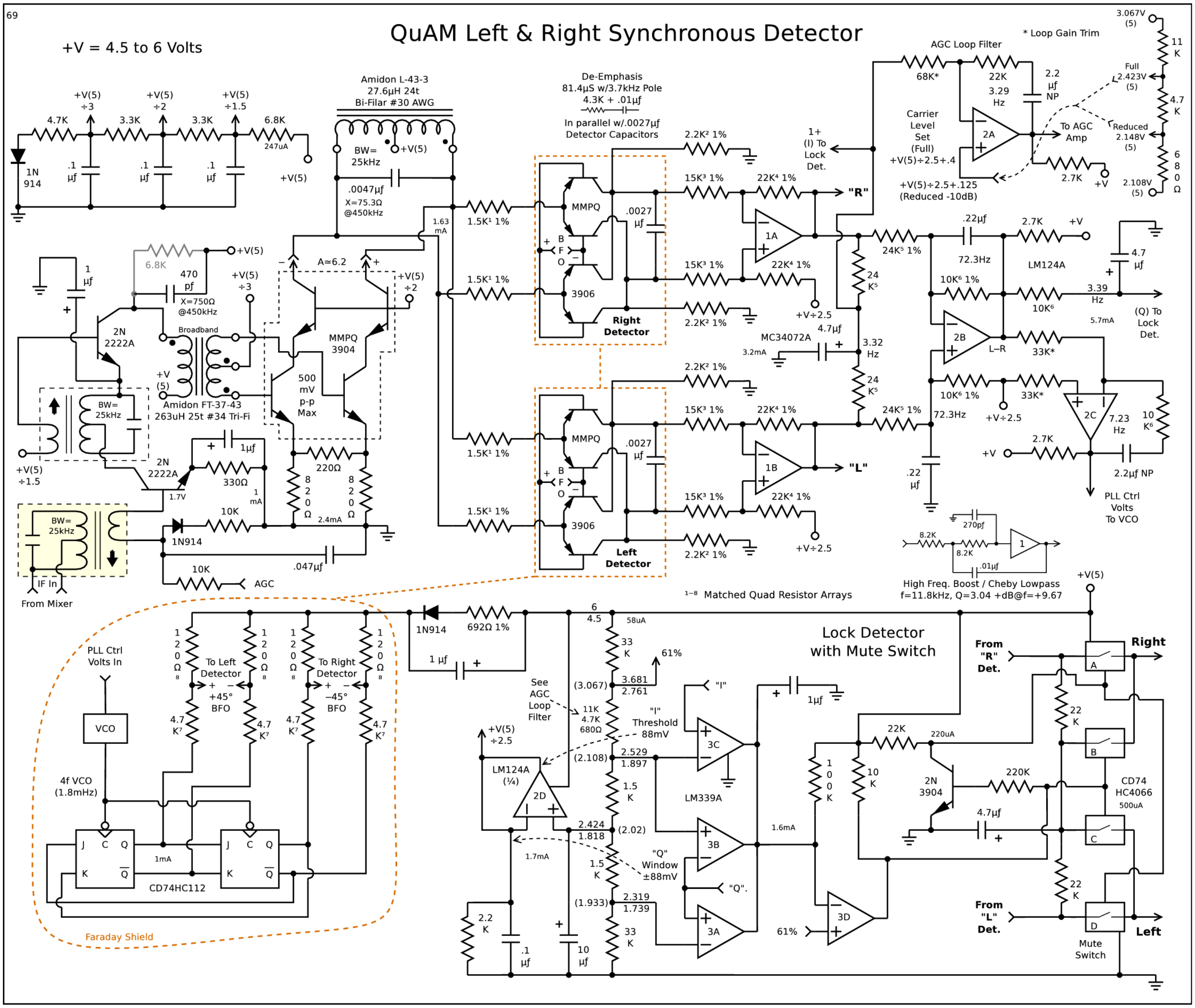
The first image on the previous page is the logic section that provides the voltage references for the window detectors on I & Q signals that determine a locked state which in turn controls the mute switch. When the PLL is not locked the signals are muted to eliminate the squeel of a unlocked synchronous detector. The voltage level for the primary op-amp bias is also buffered to ensure that it doesn't fluctuate from load variations. The mute switches shows the processing of the I & Q signals before matrixing but it may be better suited to switch the LSB(Left) & USB(Right) signals after the de-matrix and all other post filtering.

The second image on the previous page is an RF/Mixer/IF example to drive the IF input coil of the synchronous detectors. The actual driver for the detector is a cascoded differential amplifier using a quad transistor array to provide a robust signal that isolates the output from the coil driving the input.

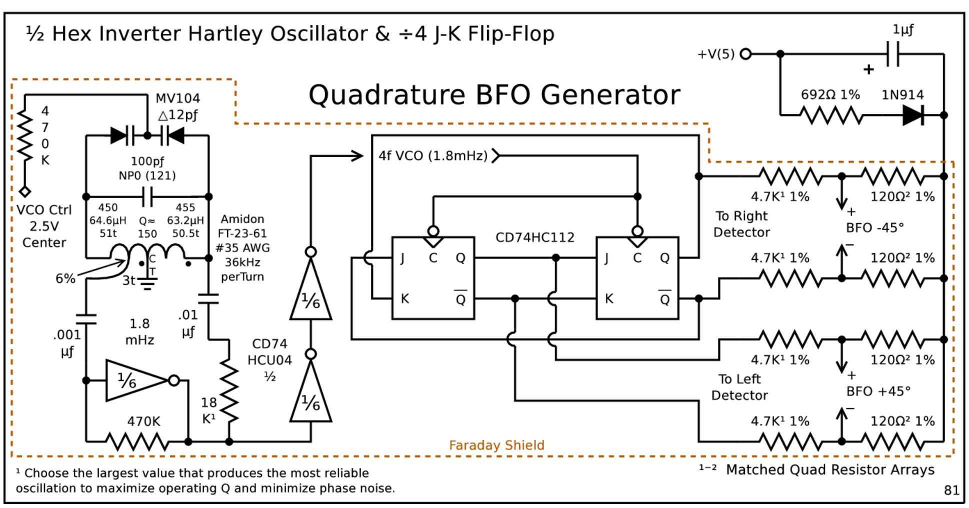
The next image is a block diagram layout of an typical ISB & QuAM receiver.



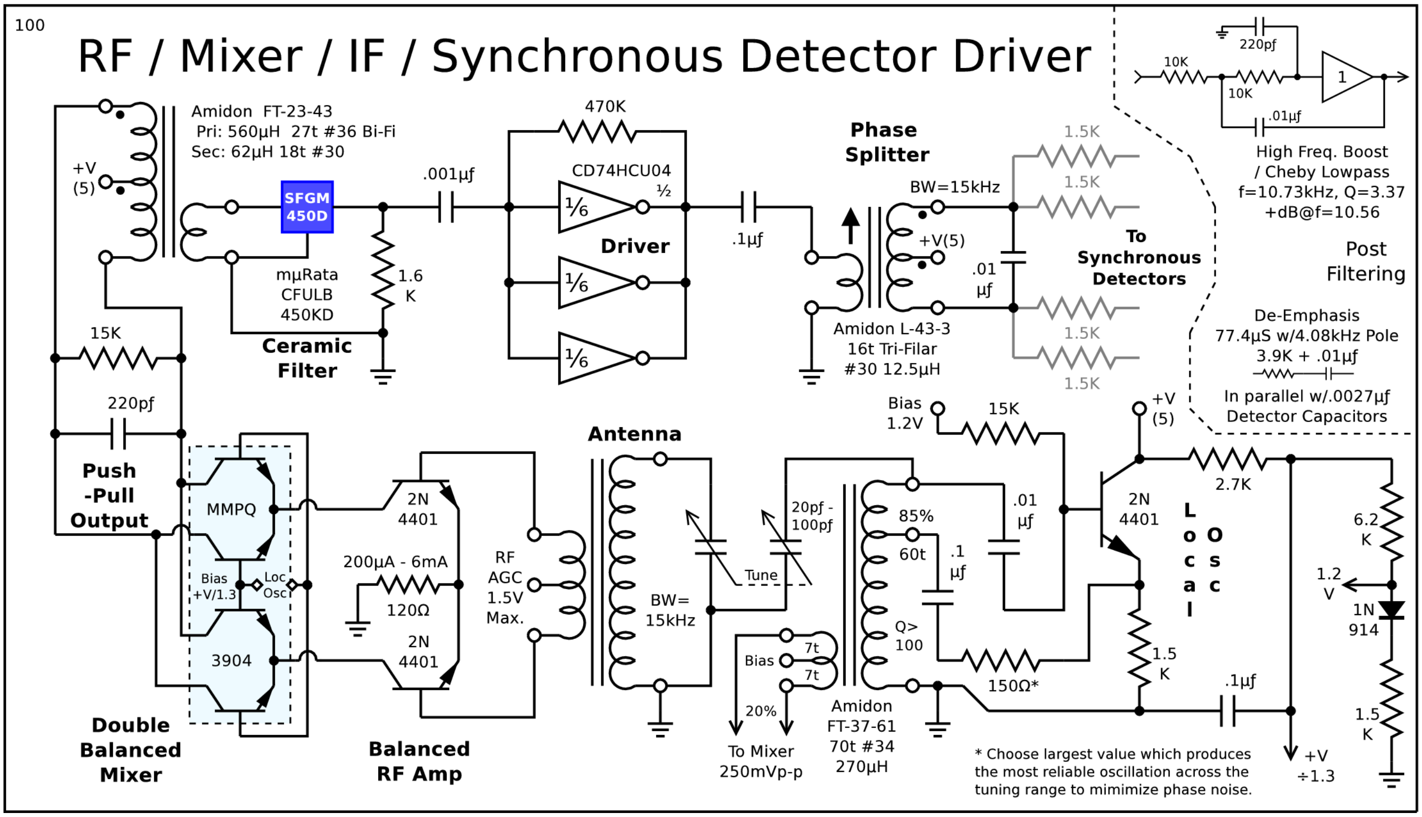
If ISB is not desired and the need for the PSNs then a simpler QuAM detector that has the detection axises aligned with ±45° for Left & Right outputs instead of 0° & +90° for I & Q axises eliminates the need for de-matrixing but a Q signal at +90° will need to be developed for the PLL loop amp. This is done with a differential op-amp to develop an L-R from L & R. The phase splitter transformer input to the detector is not broadband and is tuned for a 25kHz bandwidth. It is driven with a differential cascoded amplifier which has two other tuned IF filters with the same bandwidth. An AGC signal is applied to two of the amps providing an ~40dB adjustment range. Its input to the first IF transformer is from the mixer. The lock circuit is a bit simpler than the previous one with the elimination of the hex inverter as it doesn't need an extra control section driven by the full bypass switch for the sideband phasing filters.



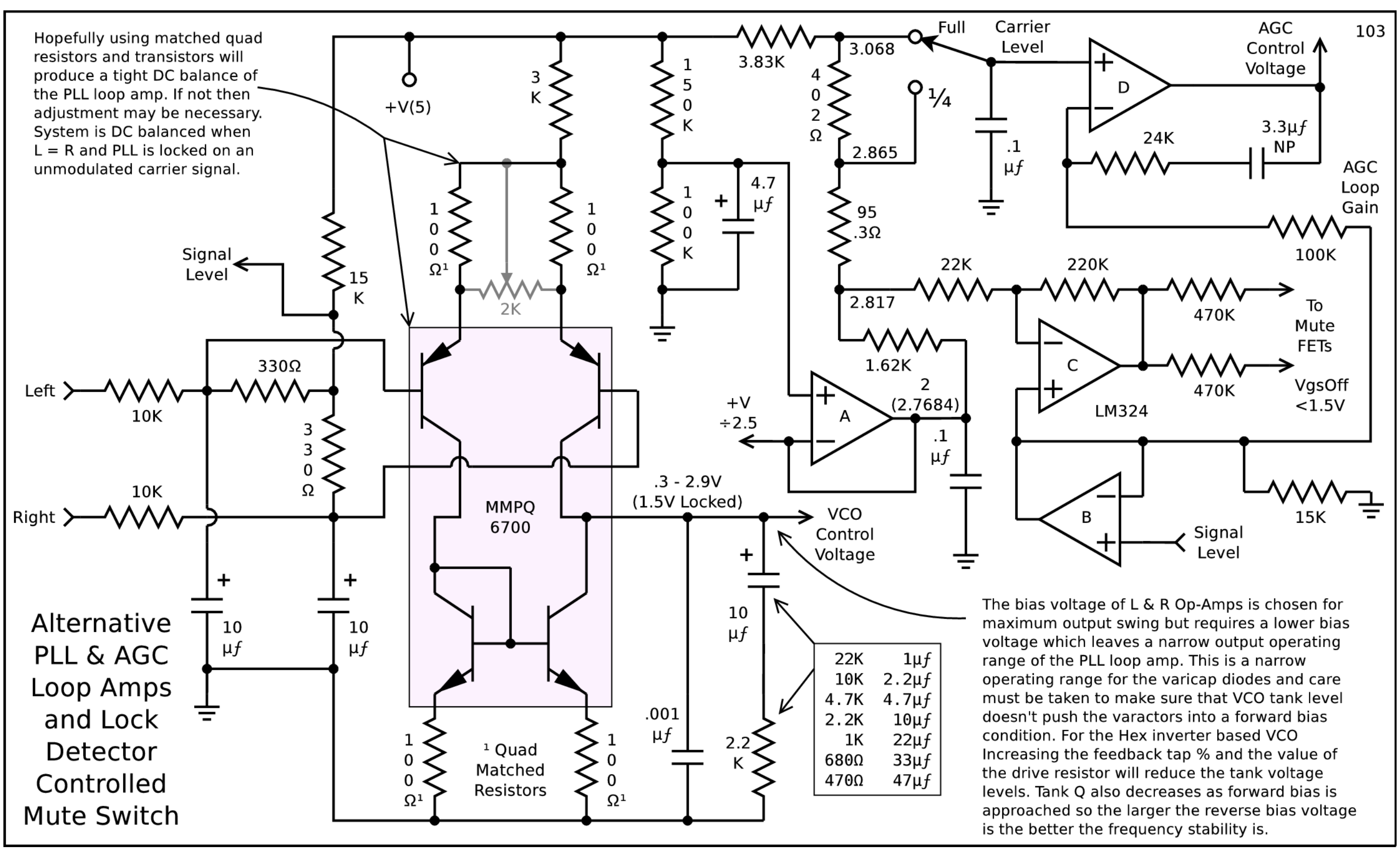
Here is essentially the same VCO circuit as previously but it is labeled for Left and Right at +45° & -45° respectively. The ÷4 F-F and resistor dividers have been repeated.



Since only half of the hex inverter is used the other half with three inverters can be used to drive the primary of the phase splitting tank for the detectors instead of the previous method provides a high impedance input with an appreciable amount of gain for the output of the ceramic filter.



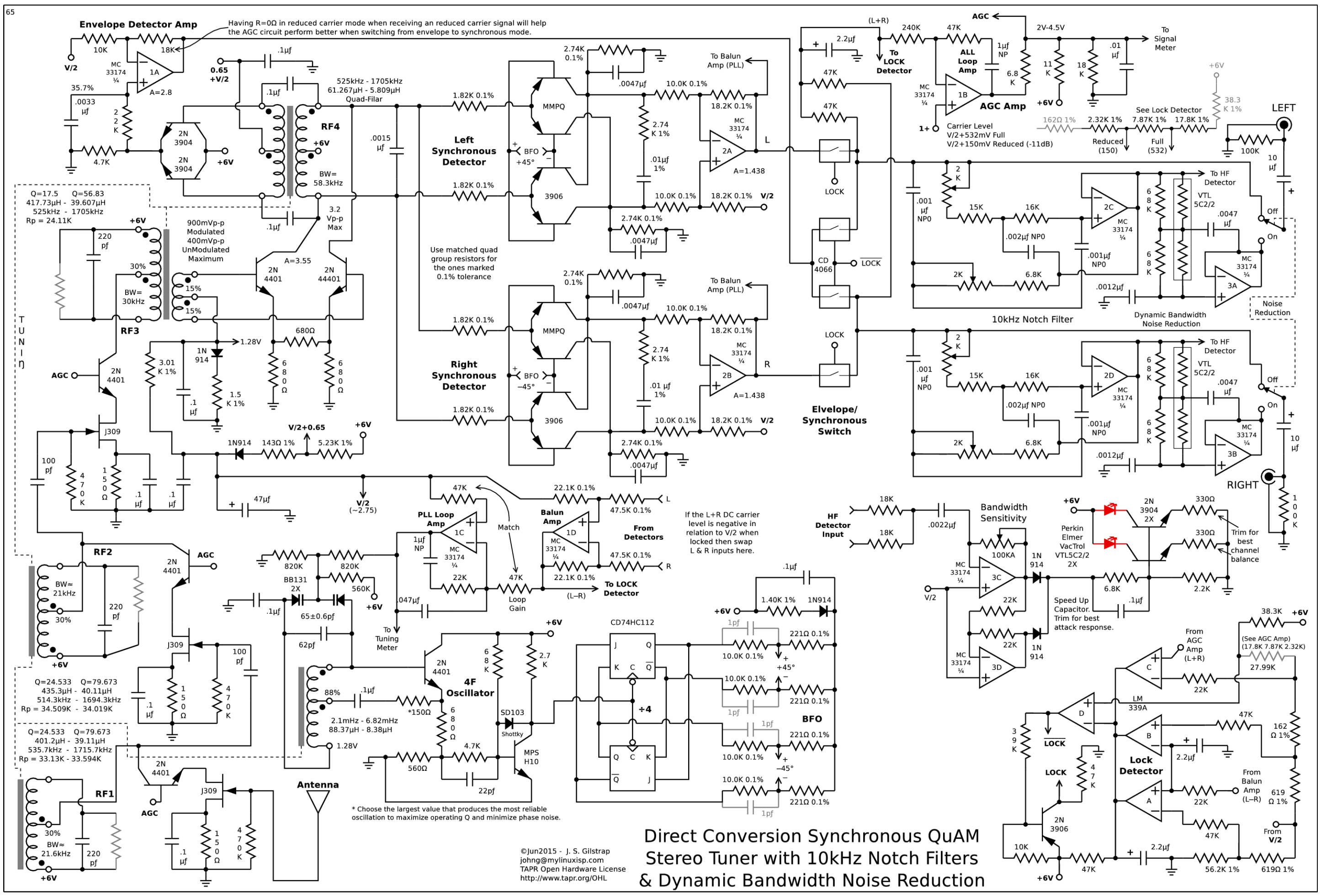
The image above contains a full RF / Loc. Osc. / Mixer / IF to go with the synchronous detectors for a complete radio. The phase splitter coil tank to the detectors is a bit narrower at 15kHz bandwidth. The input to the hex inverter is a ±10kHz ceramic filter which in turn is driven by the push-pull mixer output for a strong signal level. When the hex inverter is biased for linear use it can draw an appreciable amount of current especially during a no signal condition so a limiting resistor of 220Ω from V+ to hex inverter Vcc will help limit it the chip current. The current increases a great deal as the supply voltage increases and the gain of the inverters also decreases. It will also need to have Vcc filtered with a capacitor to ground and since the inverter drives the clock input to the F-F it would be a good idea that its Vcc filtered too with the same RC network for the same Vcc



potential. In a side block in the upper right shows the necessary de-emphasis network along with a Chebychev low pass filter which provides a >10dB boost at ~11kHz to equalize out the response produced by the RF & IF filtering to produce an ~75µs de-emphasis with an ~8.7kHz pole.

The next image eliminates the need for two of the op-amps to do the L-R matrixing and PLL loop filtering and eliminates the quad comparator also. The two extra amps in the quad op-amp are used for the AGC loop filter and the signal level controlled FET based mute switches. The two 470K resistors are connected to the gates of each FET and each FET should have a .22µ*f* connected between the gate and source. The input to the mute switch is the FET's source and the output is the drain. The FETs should have a VgsOff < 1.5V and low “on” resistance. Placing diodes with 47K resistors in series and in parallel with the 470K resistors with the anode connected to the gate of the FET will produce a fast mute response and a slower un-mute response. The PLL loop amp is preformed with a quad complimentary pair transistor array. The gain of the PLL loop amp can be adjusted by selecting the desired RC combo which has a ~7 Hz corner frequency from the table next to it. The right choice will produce a fast capture with no overshoot. With the quad transistor array and the quad 100Ω resistor array this should produce a well balanced loop amp. If not a 2K trimmer can be added as shown in gray. Whether you place the trimmer on the PNP pair or the NPN pair probably makes little difference. With the last three images all that is needed is the synchronous detectors with their op-amp baluns for a full radio. This arrangement is the least complex with the fewest parts.

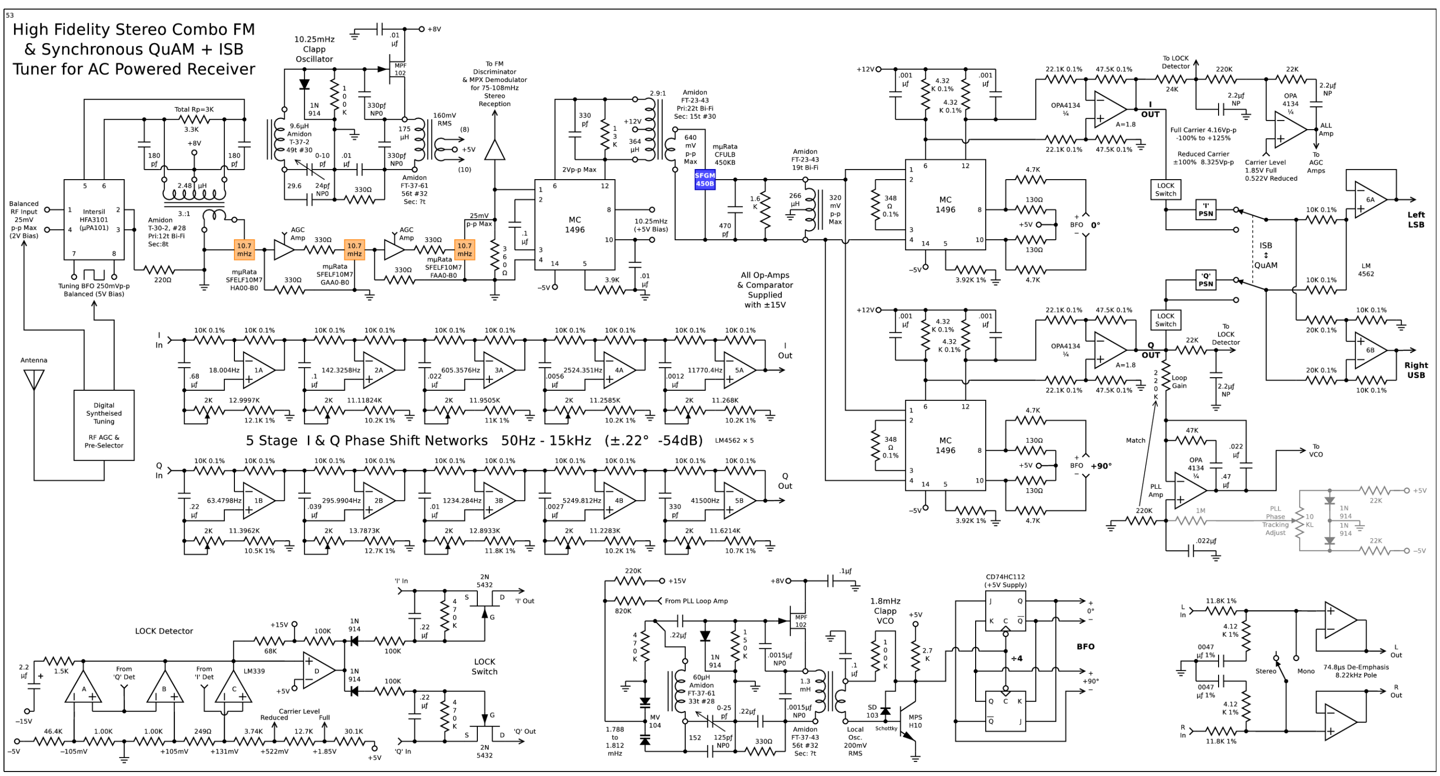
Here is a direct conversion setup using slug tuning with notch filters and dynamic bandwidth noise reduction.



A direct conversion tuner offers easier alignment accuracy for tuning across the band since all tuned circuits are changing at the same percentage including the BFO operating at 4×F. A super heterodyne tuner has a frequency offset between the RF & LO and poses greater tracking difficulty for RF center alignment in relation to the LO & IF. In the older capaitive mechanically tuned units the antenna and LO capacitance was tailored to properly track and could be fairly well aligned even for wideband use. When using a DTR with tuning diodes the same diodes are used for both antenna and LO and a series capacitance is used to approximate the frequency offset of the LO but this is not perfect. It may be satisfactory for narrow band operation but for wideband high fidelity reception the tracking is terribly inaccurate since symmetrical upper and lower sideband filtering is greatly desired for good stereo channel separation and distortion free reception for envelope detection. The downsides to mechanical capacitive tuning is microphonics and non-linear dial layout, and for DTR if not well designed center frequency modulation for RF bandpass caused by the signal which increases with signal strength can also be a problem. For both the parallel tank resistance of the tuned circuits in creases proportionally with frequency maintaining the same bandwidth. While having a constant bandwidth is greatly desired the changing resistance is not as it produces widely varying loading effects affecting the bandwidth of the circuit. When using slug tuning like used in older car radios all these issues are eliminated. While hand making a 5 gang slug tuner can be difficult it does not require the special machining needed to make a tuning capacitor. The use of staggered dual RF tanks and one center tuned tank for the front end offers a fairly flat -6dB 30kHz bandwidth response and a >60dB out of band rejection that can be fairly easily aligned. While this is a bit retro compared with today's DTR units these features are not easily obtained otherwise unless using dual conversion.

For printing this image full scale highlight and copy to clipboard and paste into an image editor. The image is at 400dpi to fit horizontally here so reset the DPI to 300 and rotate 90° for a 6.75” × 10” image to fit on 8.5” × 11” paper vertically.

This next drawing is a High Fidelity/Performance QuAM & ISB decoder using the MC1496 chip that runs on ±15V ideal for AC powered tuner in a stand alone or receiver setup. It uses a dual conversion setup to eliminate the need for the tight tracking of the RF tuned section which is an issue for a DTR front end especially for the AM/MW or LW bands. For AM/MW reception a fixed 400kHz to 2mHz RF bandpass is needed with the first IF at 10.7mHz doing the pre-selection. Regular ceramic filters for FM reception are used here offering a virtually flat response for AM RF selection. The second IF at 450kHz uses a ±15kHz ceramic filter and for ISB the phase shift networks cover the 50Hz to 15kHz audio range with a -54dB sideband suppression offering full fidelity AM reception rivaling FM. Using a very broad band mixer like HFA3101/µPA101 or a dual gate mosfet allows tuning up to 10GHz so frequency selection is limited only by the digital synthesized first local oscillator, the fixed RF broadband pre-selection and the dead zone around the 10.7mHz IF but another less conflicting IF frequency could also be used if filters are available. The first RF/LO/Mixer/IF is in semi block layout to show the general idea but could take many forms. After the 10.7mHz IF the signal can be split off for FM Stereo decoding. The second LO and PLL VCO are of the LC kind but the second LO signal could come from a ceramic, crystal or digital synthesizer although using an LC version will allow setting the second IF frequency to the exact center of the ceramic IF filter and the VCO could also be ceramic or crystal. While the output is two channel QuAM/ISB additional selection could be Left/LSB, Right/USB, or Mono from the 'I' or 'Q' signals only. The 'I' & 'Q' signals could also be provided for software decoding via an A/D converter to a computer or dedicated hardware.

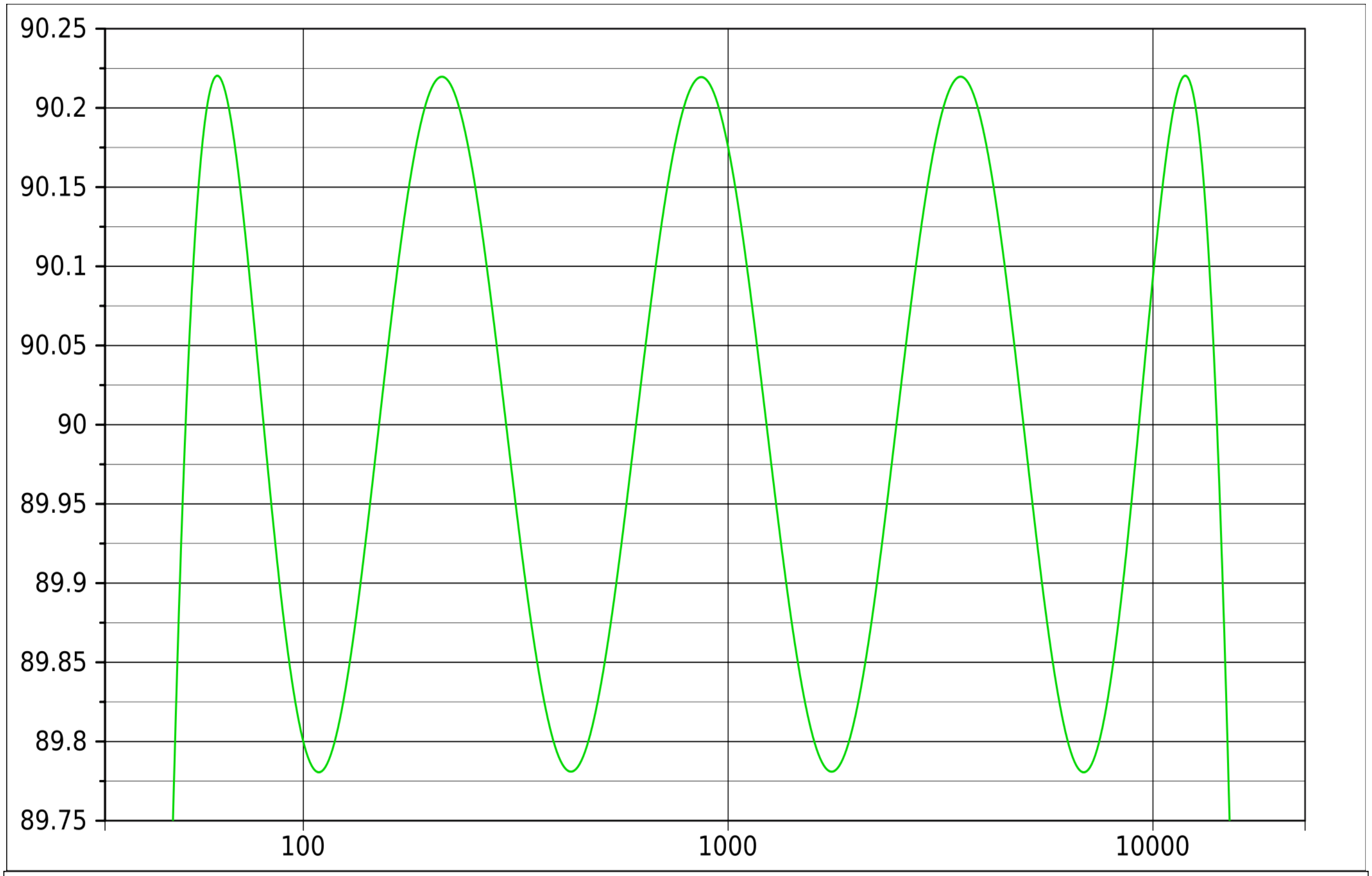


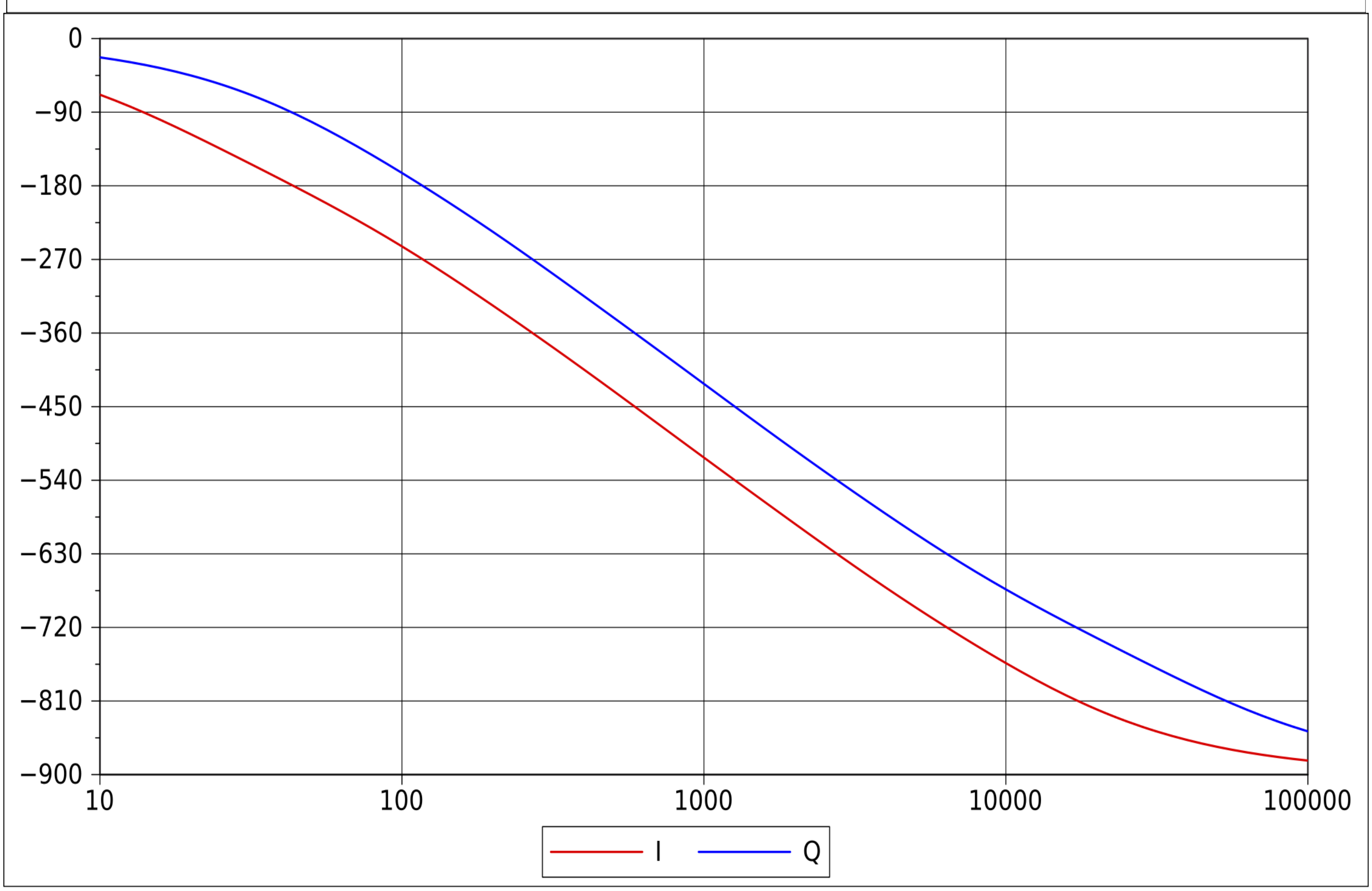
For printing follow the instructions for the previous drawing.

Super Hi-Fi 6 Stage I & Q Phase Shift Networks 40Hz – 20kHz (±0.1° -61dB)

'I'	12.633451Hz,	91.929038Hz,	341.91802Hz,	1227.7039Hz,	4439.7601Hz,	18420.899Hz
	.68µ*f* 18526Ω	.15µ*f* 11542Ω	33n*f* 14105Ω	10n*f* 12964Ω	3.3n*f* 10863Ω	680p*f* 12706Ω
'Q'	43.206761Hz,	179.26810Hz,	648.28939Hz,	2327.7726Hz,	8657.8453Hz,	63000.000Hz
	.33µ*f* 11162Ω	68n*f* 13056Ω	22n*f* 11159Ω	4.7n*f* 14547Ω	1.5n*f* 11255Ω	220p*f* 11483Ω

5 Stage Phase Shift Network Graphs





6 Stage Phase Shift Network Graphs