12.7 VHF and UHF Receivers

Most of the basic ideas presented in previous sections apply equally well in receivers that are intended for the VHF and UHF bands. This section will focus on the difference between VHF/UHF and HF receivers.

12.7.1 FM Receivers

Narrow-band frequency modulation (NBFM) is the most common mode used on VHF and UHF. **Fig 12.28A** is a block diagram of an FM receiver for the VHF/UHF amateur bands.

FRONT END

A low-noise front end is desirable because of the decreasing atmospheric noise level at these frequencies and also because portable gear often uses short rod antennas at ground level. Nonetheless, the possibilities for gain compression and harmonic IMD, multi-tone IMD and cross modulation are also substantial. Therefore dynamic range is an important design consideration, especially if large, high-gain antennas are used. FM limiting should not occur until after the crystal filter. Because of the high occupancy of the VHF/UHF spectrum by powerful broadcast transmitters and nearby two-way radio services, front-end preselection is desirable, so that a

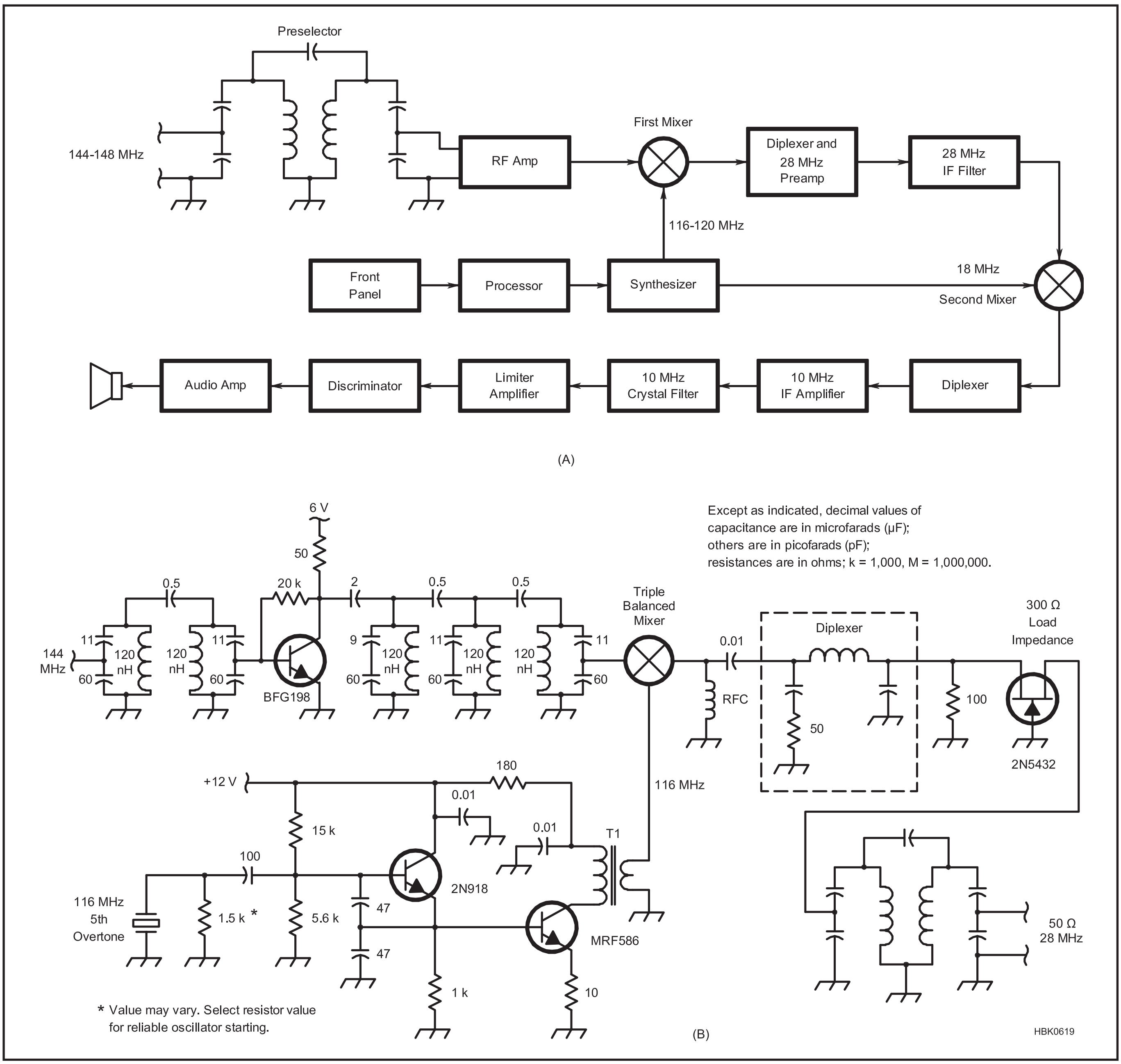


Fig 12.28 — At A, block diagram of a typical VHF FM receiver. At B, a 2 meter to 10 meter receive converter (partial schematic; some power supply connections omitted.)

low noise figure can be achieved economically within the amateur band.

DOWN-CONVERSION

Down-conversion to the final IF can occur in one or two stages. Favorite IFs are in the 5 to 10 MHz region, but at the higher frequencies rejection of the image 10 to 20 MHz away can be difficult, requiring considerable preselection. At the higher frequencies an intermediate IF in the 30 to 50 MHz region is a better choice. Fig 12.28A shows dual down-conversion.

IF FILTERS

The customary peak frequency deviation in amateur FM on frequencies above 29 MHz is about 5 kHz and the audio speech band extends to 3 kHz. This defines a maximum modulation index (defined as the deviation ratio) of 5/3 = 1.67. An inspection of the Bessel functions that describe the resulting FM signal shows that this condition confines most of the 300 to 3000 Hz speech information sidebands within a 15 kHz or so bandwidth. Using filters of this bandwidth, channel separations of 20 or 25 kHz are achievable.

Many amateur FM transceivers are channelized in steps that can vary from 1 to 25 kHz. For low distortion of the audio output (after FM detection), this filter should have good phase linearity across the bandwidth. This would seem to preclude filters with very steep descent outside the passband, which tend to have very nonlinear phase near the band edges. But since the amount of energy in the higher speech frequencies is naturally less, the actual distortion due to this effect may be acceptable for speech purposes. The normal practice is to apply pre-emphasis to the higher speech frequencies at the transmitter and de-emphasis compensates at the receiver.

LIMITING

After the filter, hard limiting of the IF is needed to remove any amplitude modulation components. In a high-quality receiver, special attention is given to any nonlinear phase shift that might result from the limiter circuit design. This is especially important in data receivers in which phase response must be controlled. In amateur receivers for speech it may be less important. Also, the *ratio detector* (see the **Mixers, Modulators and Demodulators** chapter) largely eliminates the need for a limiter stage, although the limiter approach is probably still preferred.

FM DETECTION

The discussion of this subject is deferred to the **Mixers, Modulators and Demodulators** chapter. Quadrature detection is used in some popular FM multistage ICs. An example receiver IC will be presented later.

12.7.2 FM Receiver Weak-Signal Performance

The noise bandwidth of the IF filter is not much greater than twice the audio bandwidth of the speech modulation, less than it would be in wideband FM. Therefore such things as capture effect, the threshold effect and the noise quieting effect so familiar to wideband FM are still operational, but somewhat less so, in FM. For FM receivers, sensitivity is specified in terms of a SINAD (see the **Test Equipment and Measurements** chapter) ratio of 12 dB. Typical values are –110 to –125 dBm, depending on the low-noise RF pre-amplification that often can be selected or deselected (in strong signal environments).

LO PHASE NOISE

In an FM receiver, LO phase noise superimposes phase modulation, and therefore frequency modulation, onto the desired signal. This reduces the ultimate signal-to-noise ratio within the passband. This effect is called "incidental FM (IFM)." The power density of IFM (W/Hz) is proportional to the phase noise power density (W/Hz) multiplied by the square of the modulating frequency (the familiar parabolic effect in FM). If the receiver uses high-frequency de-emphasis at the audio output (–6 dB per octave from 300 to 3000 Hz, a common practice), the IFM level at higher audio frequencies can be reduced. Ordinarily, as the signal increases the noise would be "quieted" (that is, "captured") in an FM receiver, but in this case the signal and the phase noise riding "piggy back" on the signal increase in the same proportion as described in the **Oscillators and Synthesizers** chapter's discuss of reciprocal mixing. IFM is not a significant problem in modern FM radios, but phase noise can become a concern for adjacent-channel interference.

As the signal becomes large the signal-to-noise ratio therefore approaches some final value. A similar ultimate SNR effect occurs in SSB receivers. On the other hand, a perfect AM receiver tends to suppress LO phase noise. (See the reference entry for Sabin.)

12.7.3 FM Receiver ICs

A wide variety of special ICs for communications-bandwidth FM receivers are available. Many of these were designed for "cordless" or mobile telephone applications and are widely used. One is an RF amplifier chip (NE/SA5204A) for 50 Ω input to 50 Ω output with 20 dB of gain. The second chip (NE/SA602A) is a front-end device with an RF amplifier, mixer and LO. The third is an IF amplifier, limiter and quadrature FM detector (NE/SA604A) that also has a very useful RSSI (logarithmic Received Signal Strength Indicator) output and also a "mute" function. The fourth is the LM386, a widely used audio-amplifier chip. Another FM receiver chip, complete in one package, is the MC3371P.

Another recent IC is the MC13135, which features double conversion and two IF amplifier frequencies. This allows more gain on a single chip with less of the cross coupling that can degrade stability. This desirable feature of multiple down-conversion was mentioned previously in this chapter.

Design details and specific parts values can be learned from a careful study of the data sheets and application notes provided by the IC vendors. Amateur designers should learn how to use these data sheets and other information such as application notes available (usually for free) from the manufacturers or on the web.

12.7.4 VHF Receive Converters

Rather than building an entire transceiver for VHF SSB and CW, one approach is to use a receive converter. A receive converter (also called a *downconverter*) takes VHF signals and converts them to an HF band for reception using existing receiver or transceiver as a tunable IF.

Although many commercial transceivers cover the VHF bands (either multiband, multimode VHF/UHF transceivers, or HF+VHF transceivers), receive converters are sometimes preferred for demanding applications because they may be used with high-performance HF transceivers. Receive converters are often packaged with a companion transmit converter and control circuitry to make a *transverter*.

A typical 2 meter downconverter uses an IF of 28-30 MHz. Signals on 2 meters are amplified by a low-noise front-end before mixing with a 116 MHz LO. Fig 12.28B shows the schematic for a high-performance converter. The front-end design was contributed by Ulrich Rohde, N1UL, who recommends a triple-balanced mixer such as the Synergy CVP206 or SLD-K5M.

The diplexer filter at the mixer output selects the difference product: (144 to 146 MHz) – 116 = (28 to 30 MHz). A common-base buffer amplifier (the 2N5432 FET) and tuned filter form the input to the 10 meter receiver. (N1UL suggests that using an IF of 21 MHz and an LO at 165 MHz would avoid interference problems with 222 MHz band signals.) For additional oscillator designs, refer to the papers on oscillators by N1UL in the supplemental CD-ROM files for the **Mixers, Modulators and Demodulators** chapter accompanying this *Handbook*.

Based on the Philips BFG198 8 GHz transistor, the 20 dB gain front-end amplifier is optimized for noise figure (NF is approximately 2.6 dB), not for input impedance. The output circuit is optimized for best selectivity. The transistor bias is designed for dc stability at I_C = 30 mA and V_C = 6 V. Both of the transistor's emitter terminals should be grounded to prevent oscillation. NF might be improved with a higher performance transistor, such as a GaAs FET, but stability problems are often encountered with FET designs in this application.

If a mast-mounted preamplifier is used to improve the system noise figure, an attenuator should be available to prevent overload. Simulation predicts the circuit to have an IP3 figure of at least +25 dBm at 145 MHz with an I_C of 30 mA and a terminating impedance of 50 Ω.

Project: Dual-band LNA for 2 Meters and 70 Centimeters

(The following article was originally published in the Nov/Dec 2015 issue of *AMSAT Journal* and is included, along with an addendum, as a PDF file on this book's CD-ROM.)

This project describes constructing a dual-band low noise amplifier (LNA) for use on 2 meters and 70 centimeters. It includes bypass capability to permit transmitting on either band while listening to the downlink signal on the opposite band of a VHF/UHF satellite.

LNAs are needed when a long run of coax is used between the antenna and receiver. LNAs must be mounted as close to the antenna as is practical because, once the signal is lost in the coax, there is no way to get it back. LNAs mounted at the antenna amplify the weak signals before they are attenuated in coax. (See the section "Receiver Sensitivity" elsewhere in this chapter for a discussion of noise figure and the effect of feed line loss on receiver sensitivity.)

The noise figure is around 0.5 dB on 2 meters and 0.8 dB on 70 centimeters with 20 – 25 dB gain. As shown in the block diagram in **Fig 12.29**, relays are used to bypass the LNA when transmitting. The power handling capability is limited by the relays and the coax. The relay specified in this project is rated at 50 W up to 1 GHz.

The LNAs are based on the MiniCircuits PGA-103 IC. (PCB kits referenced in the

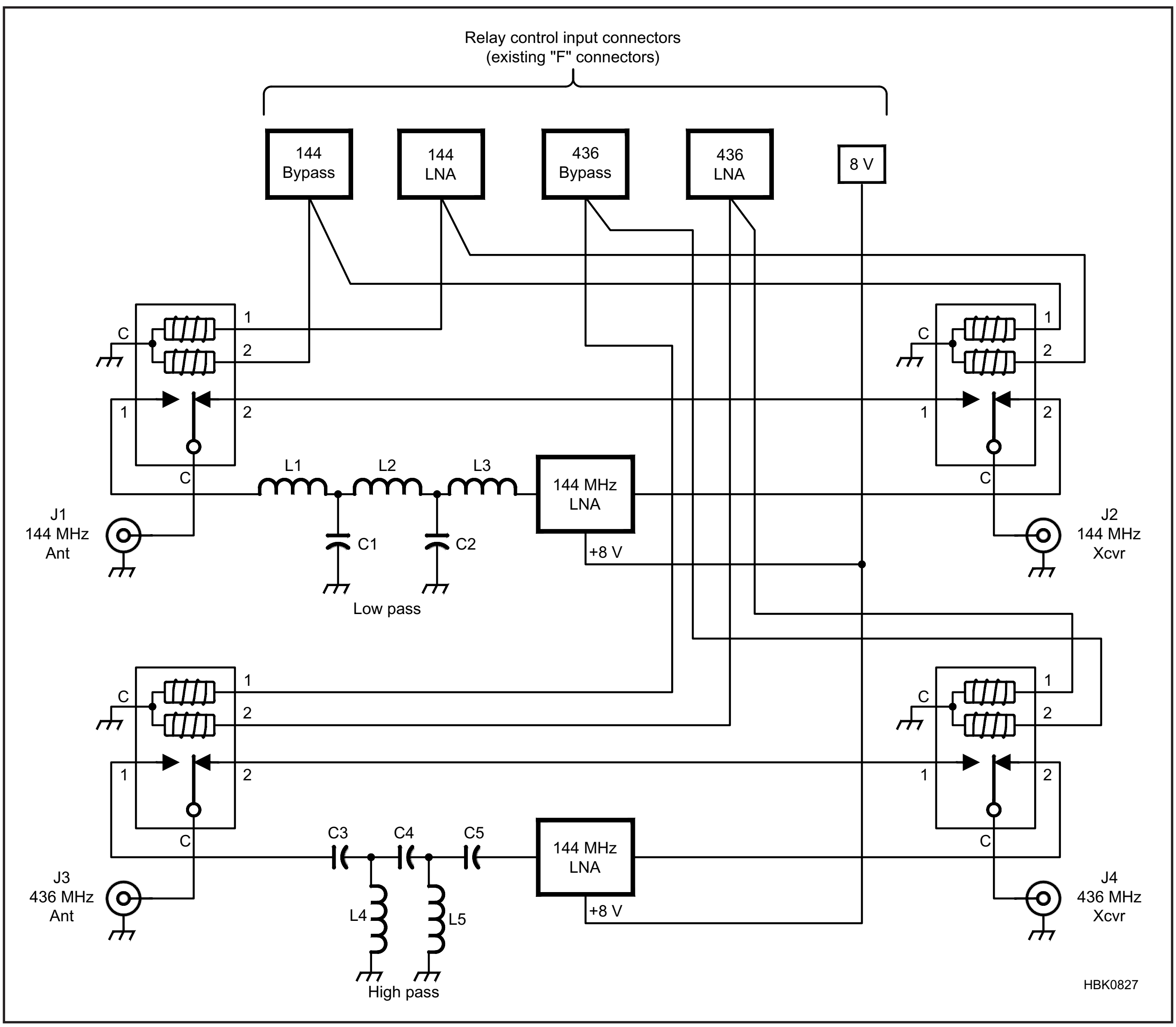


Fig 12.29 — Dual LNA block diagram. Power supply voltage (+8 V) may need to be changed if different relays or LNA modules are used from those specified in the article. J1-J4 are N connectors. J1 and J3 are located on one side of the chassis, and J2 and J4 are located on the other side of the chassis. See the PDF article for component values.

original article are no longer available but the LNA can be constructed "dead-bug style" on PCB stock or a simple PCB can be laid out for inexpensive fabrication.) The data sheet for this IC provides the necessary information for constructing the LNA. Alternatively, other LNA modules are available from several vendors ranging from bare PCBs to fully enclosed, standalone units.

A low-pass filter in the input to the 2 meter LNA rejects 70 centimeter signals from the transmitter. A high-pass filter in the input to the 70 centimeter LNA rejects 2 meter signals from the transmitter. This prevents overloading the LNAs while listening on one band and transmitting on the other (that is, when listening to your uplink signal on the satellite downlink in full duplex). The filters require silvered-mica capacitors and air-wound coils.

To use these LNAs for single-band terrestrial operation a sequencer is required to switch the state of the relays before transmitting. For satellite operation, a simple toggle switch can be used to set one LNA to bypass (uplink) and the other to LNA (downlink).

A waterproof enclosure and RF connectors are required assuming the antennas will be located outdoors. It is recommended that the enclosure be large enough to allow for convenient wiring and to make swapping out amplifier modules, relays, and filters as higher-performance units become available.

12.8 UHF and Microwave Techniques

The ultra high frequency spectrum comprises the range from 300 MHz to 3 GHz. All of the basic principles of radio system design and circuit design that have been discussed so far apply as well in this range, but the higher frequencies require some special thinking about the methods of circuit design and the devices that are used. Additional material on construction for microwave circuits can be found in the **Construction Techniques** chapter and in the series of *QST* columns, "Microwavelengths" by Paul Wade, W1GHZ.

12.8.1 UHF Construction

Modern receiver designs make use of highly miniaturized monolithic microwave ICs (MMICs). Among these are the Avago MODAMP and the Mini Circuits MAR and MAV/ERA lines. They come in a wide variety of gains, intercepts and noise figures for frequency ranges from dc to well into the GHz range. (See the **Component Data and References** chapter for information on available parts.)

Fig 12.30 shows the schematic diagram and the physical construction of a typical RF circuit at 430 MHz. It is a GaAsFET preamplifier intended for low noise SSB/CW, moonbounce or satellite reception. The construction uses ceramic chip capacitors, small helical inductors and a stripline surface-mount GaAsFET, all mounted on a G10 (two layers of copper) glass-epoxy PC board. The very short length of interconnection leads is typical. The bottom of the PC board is a ground plane. At this frequency, lumped components are still feasible, while microstrip circuitry tends to be rather large.

At higher frequencies, microstrip methods become more desirable in most cases because of their smaller dimensions. However, the advent of tiny chip capacitors and chip resistors has extended the frequency range of discrete components. For example, the literature shows methods of building LC filters at as high as 2 GHz or more, using chip capacitors and tiny helical inductors. Amplifier and mixer circuits operate at well into the GHz range using these types of components on controlled-dielectric PC board material such as Duroid or on ceramic substrates.

Current designs emphasize simplicity of construction and adjustment, leading to "no tune" designs. The use of printed-circuit microstrip filters that require little or no adjustment, along with IC or MMIC devices, or discrete transistors, in precise PC-board layouts that have been carefully worked out, make it much easier to "get going" on the higher frequencies.

12.8.2 UHF Design Aids

Circuit design and evaluation at the higher frequencies usually require some kind of minimal lab facilities, such as a signal generator, a calibrated noise generator and, hopefully, some kind of simple (or surplus) spectrum analyzer. This is true because circuit behavior and stability depend on a number of factors that are difficult to "guess at," and intuition is often unreliable. The ideal instrument is a vector network analyzer with all of the attachments (such as an S parameter measuring setup), an instrument that has become surprisingly affordable in recent years. (See the **Test Equipment and Measurements** chapter.)

Another very desirable thing would be a circuit design and analysis program for the personal computer. Software packages created especially for UHF and microwave circuit design are available. They tend to be somewhat expensive, but worthwhile for a serious designer. Inexpensive *SPICE* programs are a good compromise but have significant limitations at VHF and above. See the chapter on **Computer-Aided Circuit Design** for information on these tools.

12.8.3 A 902 to 928 MHz (33 cm) Receiver

This 902 MHz downconverter is a fairly typical example of receiver design methods for the 500 to 3000 MHz range, in which down-conversion to an existing HF receiver (or 2 meter multimode receiver) is the most convenient and cost-effective approach for most amateurs. At higher frequencies a double down-conversion with a first IF of 200 MHz or so, to improve image rejection, might be necessary. Usually, though, the presence of strong signals at image frequencies is less likely. Image-reducing mixers plus down-conversion to 28 MHz is also coming into use, when strong interfering signals are not likely at the image frequency.

Fig 12.31A is the block diagram of the 902 MHz down-converting receiver. A cavity resonator at the antenna input provides high selectivity with low loss. The first RF amplifier is a GaAsFET. Two additional 902 MHz band-pass microstrip filters and a second RF amplifier transistor provide more gain and image rejection (at RF – 56 MHz) for the mixer. The output is at 28.0 MHz so that an HF receiver can be used as a tunable IF/demodulator stage.

CUMULATIVE NOISE FIGURE

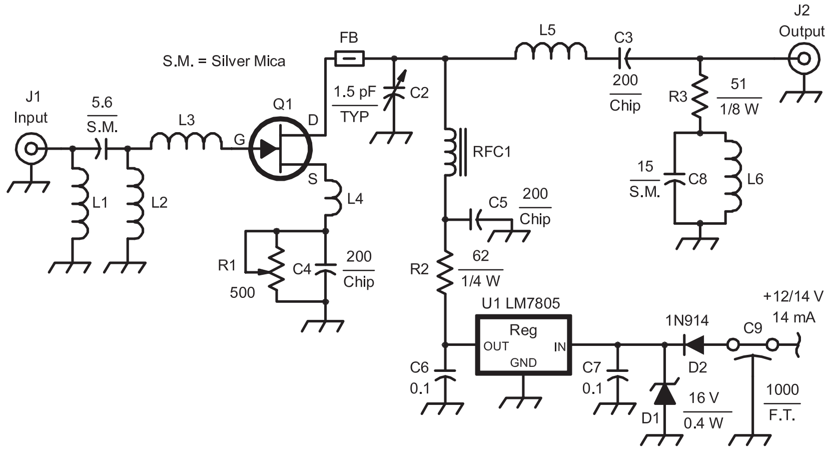
Fig 12.31B shows the cumulative noise figure (NF) of the signal path, including the 28 MHz receiver. The 1.5 dB cumulative NF of the input cavity and first RF-amplifier combination, considered by itself, is degraded to 1.9 dB by the rest of the system following the first RF amplifier. The NF values of the various components for this example are reasonable, but may vary somewhat for actual hardware. Also, losses prior to the input such as transmission line losses (very important) are not included. They would be part of the complete receive system analysis, however. It is common practice to place a low noise preamp outdoors, right at the antenna, to overcome coax loss (and to permit use of less expensive coax).

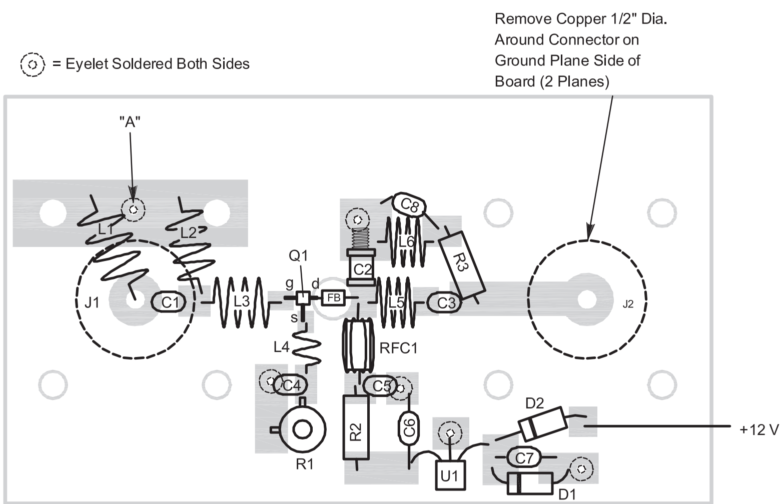
LOCAL OSCILLATOR (LO) DESIGN

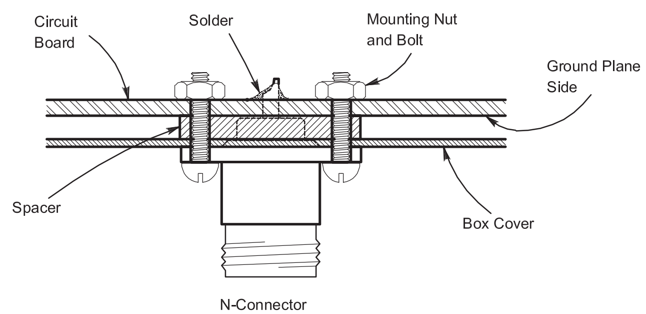
The +7-dBm LO at 874 to 900 MHz is derived from a set of crystal oscillators and frequency multipliers, separated by band-pass filters. These filters prevent a wide assortment

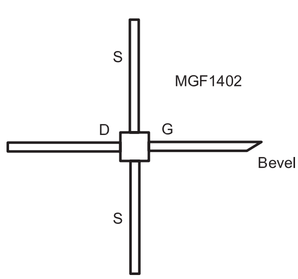
Fig 12.30 — GaAsFET preamplifier schematic and construction details for 430 MHz. Illustrates circuit, parts layout and construction techniques suitable for 430-MHz frequency range.

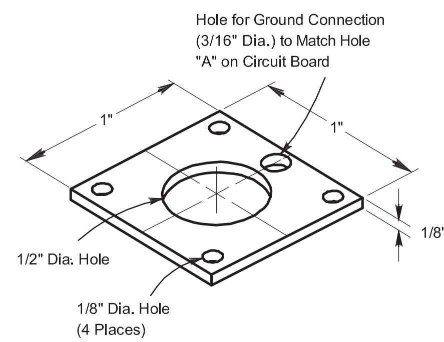
C1 — 5.6 pF silver-mica or same as C2.
C2 — 0.6 to 6 pF ceramic piston trimmer (Johanson 5700 series or equiv).
C3, C4, C5 — 200 pF ceramic chip.
C6, C7 — 0.1 µF disc ceramic, 50 V or greater.
C8 — 15 pF silver-mica.
C9 — 500 to 1000 pF feedthrough.
D1 — 16 to 30 V, 500 mW Zener (1N966B or equiv).
D2 — 1N914, 1N4148 or any diode with ratings of at least 25 PIV at 50 mA or greater.
J1, J2 — Female chassis-mount Type-N connectors, PTFE dielectric (UG-58 or equiv).
L1, L2 — 3t, #24 tinned wire, 0.110-inch ID spaced 1 wire dia.
L3 — 5t, #24 tinned wire, 3/16-inch ID, spaced 1 wire dia. or closer. Slightly larger diameter (0.010 inch) may be required with some FETs.
L4, L6 — 1t #24 tinned wire, ⅛-inch ID.
L5 — 4t #24 tinned wire, ⅛-inch ID, spaced 1 wire dia.
Q1 — Mitsubishi MGF1402.
R1 — 200 or 500-Ω Cermet potentiometer (initially set to midrange).
R2 — 62 Ω, ¼ W.
R3 — 51 Ω, ⅛ W carbon composition resistor, 5% tolerance.
RFC1 — 5t #26 enameled wire on a ferrite bead.
U1 — 5 V, 100-mA 3 terminal regulator (LM78L05 or equiv. TO-92 package).

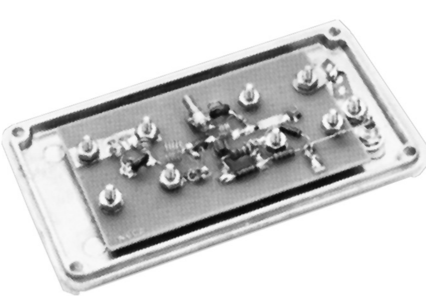












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Aluminum Spacer for N-Connector

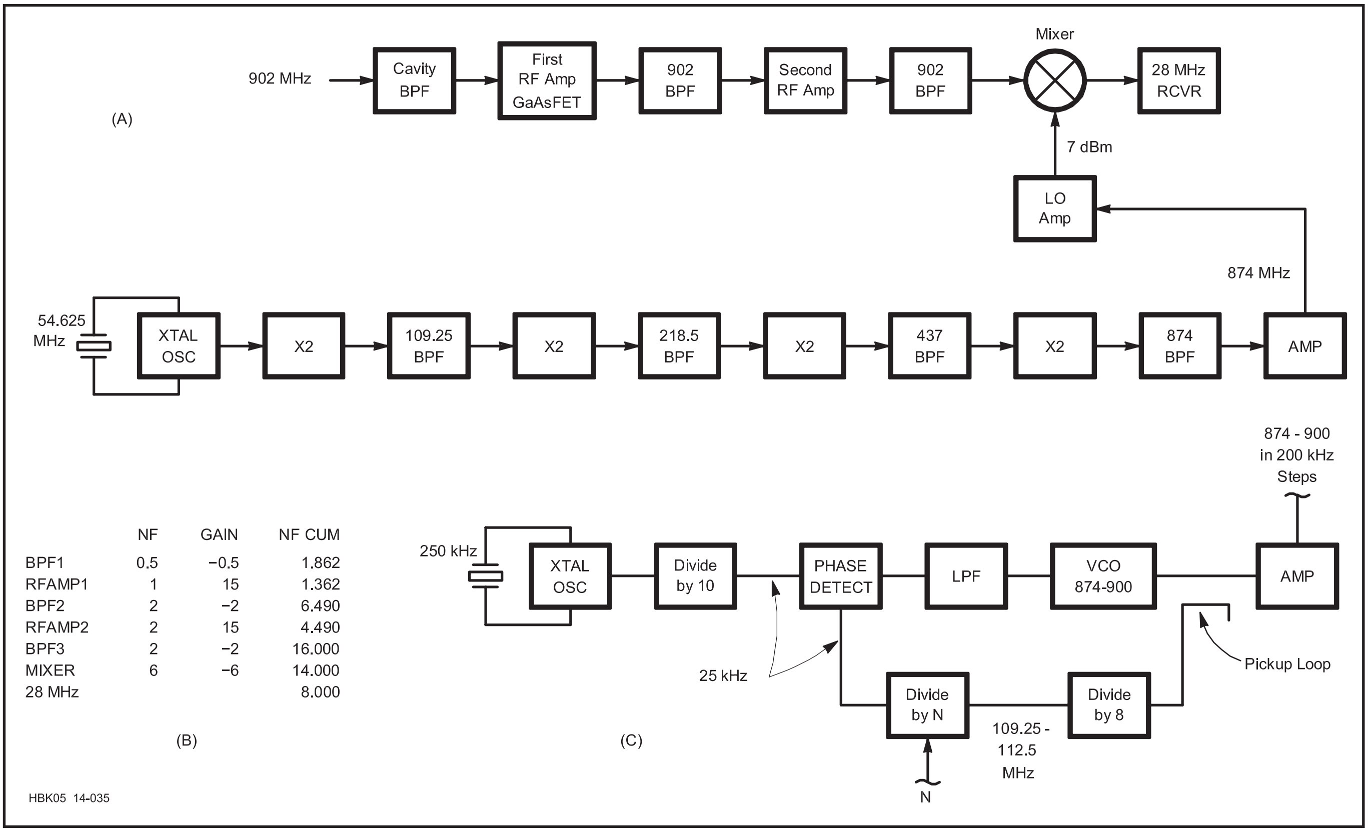

Fig 12.31 — A downconverter for the 902 to 928 MHz band. At A: block diagram; At B: cumulative noise figure of the signal path; At C: alternative LO multiplier using a phase locked loop.

of spurious frequencies from appearing at the mixer LO port. They also enhance the ability of the doubler stage to generate the second harmonic. That is, they have very low impedance at the input frequency, thereby causing a large current to flow at the fundamental frequency. This increases the nonlinearity of the circuit, which increases the second-harmonic component. The higher filter impedance at the second harmonic produces a large harmonic output.

For very narrow-bandwidth use, such as EME, the crystal oscillators are often oven controlled or otherwise temperature compensated. The entire LO chain must be of low-noise design and the mixer should have good isolation from LO port to RF port (to minimize noise transfer from LO to RF).

A phase-locked loop using GHz range prescalers (as shown in Fig 12.31C) is an alternative to the multiplier chain. The divide-by-N block is a simplification; in practice, an auxiliary dual-modulus divider (see the **Oscillators and Synthesizers** chapter) would be involved in this segment. The cascaded 902 MHz band-pass filters in the signal path should attenuate any image frequency noise (at RF–56 MHz) that might degrade the mixer noise figure.

12.8.4 Microwave Receivers

The world above 3 GHz is a vast territory with a special and complex technology well beyond the scope of this chapter. We will scratch the surface by describing a specific receiver for the 10 GHz frequency range and point out some of the important special features that are unique to this frequency range.

A 10 GHZ PREAMPLIFIER

Fig 12.32B is a schematic and parts list, Fig 12.32C is a PC board parts layout and Fig 12.32A is a photograph of a 10 GHz preamp, designed by Senior ARRL Lab Engineer Zack Lau, W1VT. With very careful design and packaging techniques a noise figure approaching the 1 to 1.5 dB range was achieved. This depends on an accurate 50-Ω generator impedance and noise matching the input using a microwave circuit-design program such as *Touchstone* or *Harmonica*. Note that microstrip capacitors, inductors and transmission-line segments are used almost exclusively. The circuit is built on a 15-mil Duroid PC board. In general, this kind of performance requires some elegant measurement equipment that few amateurs have. On the other hand, preamp noise figures in the 2 to 4-dB range are much easier to get (with simple test equipment) and are often satisfactory for amateur terrestrial communication.

Articles written by those with expertise and the necessary lab facilities almost always include PC board patterns, parts lists and detailed instructions that are easily duplicated by readers. Microwave ham clubs and their publications are a good way to get started in microwave amateur technology.

Because of the frequencies involved, dimensions of microstrip circuitry must be very accurate. Dimensional stability and dielectric constant reliability of the boards must be very good.

System Performance

At microwaves, an estimation of system performance can often be performed using known data about the signal path terrain, atmosphere, transmitter and receivers systems. In the present context of receiver design we wish to establish an approximate goal for the receiver system, including the antenna and transmission line. **Fig 12.33** shows a simplified example of how this works.

A more detailed analysis includes terrain variations, refraction effects, the Earth's curvature, diffraction effects and interactions with the atmosphere's chemical constituents

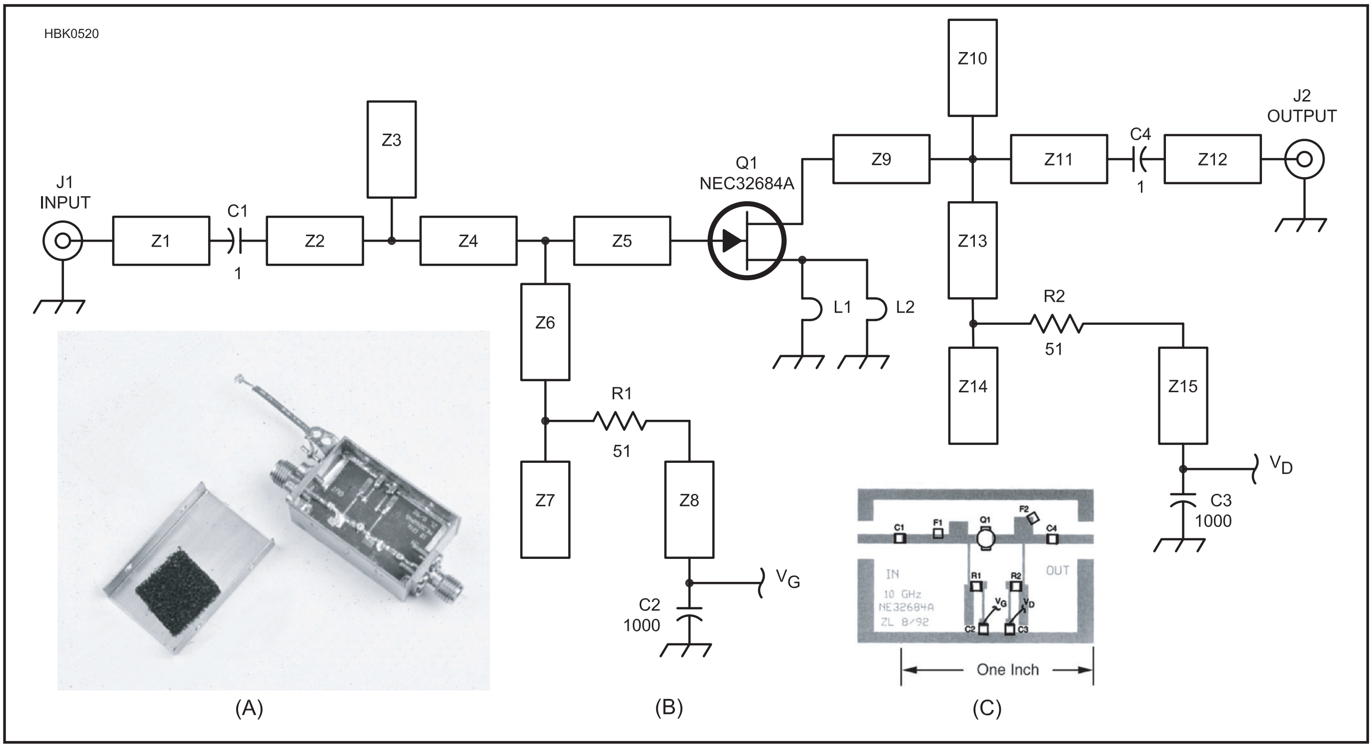

Fig 12.32 — At A, a low-noise preamplifier for 10 GHz, illustrating the methods used at microwaves. At B: schematic. At C: PC board layout. Use 15-mil 5880 Duroid, dielectric constant of 2.2 and a dissipation factor of 0.0011. A template of the PC board is available on the CD-ROM included with this book.

C1, C4 — 1 pF ATC 100 A chip capacitors. C1 must be very low loss.
C2, C3 — 1000 pF chip capacitors. (Not critical.) The ones from Mini Circuits work fine.
F1, F2 — Pieces of copper foil used to tune the preamp.
J1, J2 — SMA jacks. Ideally these should be microstrip launchers. The pin should be flush against the board.
L1, L2 — The 15 mil lead length going through the board to the ground plane.
R1, R2 — 51 Ω chip resistors.
Z1-Z15 — Microstrip lines etched on the PC board.

Analysis of a 10.368 GHz communication link with SSB modulation:

Free space path loss (FSPL) over a 50-mile line-of-sight path (S) at F = 10.368 GHz:
FSPL = 36.6 (dB) + 20 log F (MHz) + 20 log S (Mi) = 36.6 + 80.3 + 34 = 150.9 dB.

Effective isotropic radiated power (EIRP) from transmitter:
EIRP (dBm) = P_{XMIT} (dBm) + Antenna Gain (dBi)

The antenna is a 2-ft diameter (D) dish whose gain GA (dBi) is:
GA = 7.0 + 20 log D (ft) + 20 log F (GHz) = 7.0 + 6.0 + 20.32 = 33.3 dBi

Assume a transmission-line loss L_T, of 3 dB
The transmitter power P_T = 0.5 (mW PEP) = –3 (dBm PEP)
P_{XMIT} = P_T (dBm PEP) – L_T (dB) = (–3) – (3) = –6 (dBm PEP)
EIRP = P_{XMIT} + G_A = –6 + 33.3 = 27.3 (dBm PEP)
Using these numbers the received signal level is:
P_{RCVD} = EIRP (dBm) – Path loss (dB) = 27.3 (dBm PEP) – 150.9 (dB) = –123.6 (dBm PEP)
Add to this a receive antenna gain of 17 dB. The received signal is then P_{RCVD} = –123.6 +17 = –106.6 dBm

Now find the receiver's ability to receive the signal:
The antenna noise temperature T_A is 200 K. The receiver noise figure NF_R is 6 dB (FR=3.98, noise temperature T_R = 864.5 K) and its noise bandwidth (B) is 2400 Hz. The feedline loss L_L is 3 dB (F = 2.00, noise temperature T_L = 288.6 K). The system noise temperature is:
$T_S = T_A + T_L + (L_L)\,(T_R)$
T_S = 200 + 288.6 + (2.0) (864.5) = 2217.6 K
$N_S = kT_SB = 1.38 \times 10^{-23} \times 2217.6 \times 2400 = 7.34 \times 10^{-17}$ W = –131.3 dBm
This indicates that the PEP signal is –106.6 –(–131.3) = 24.7 dB above the noise level. However, because the average power of speech, using a speech processor, is about 8 dB less than PEP, the average signal power is about 16.7 dB above the noise level.

To find the system noise factor F_S we note that the system noise is proportional to the system temperature T_S and the "generator" (antenna) noise is proportional to the antenna temperature T_A. Using the idea of a "system noise factor":

$F_S = T_S / T_A$ = 2217.6 / 200 = 11.09 = 10.45 dB.

If the antenna temperature were 290 K the system noise figure would be 9.0 dB, which is precisely the sum of receiver and receiver coax noise figures (6.0 + 3.0).

Fig 12.33 — Example of a 10-GHz system performance calculation. Noise temperature and noise factor of the receiver are considered in detail.

and temperature gradients.

In microwave work, where very low noise levels and low noise figures are encountered, experimenters like to use the "effective noise temperature" concept, rather than noise factor. The relationship between the two is given by

$$T_E = 290\,(F - 1) \tag{8}$$

where the noise factor $F = 10^{NF/10}$ and NF is the noise figure in dB.

T_E is a measure, in terms of temperature, of the "excess noise" of a component (such as an amplifier). A resistor at T_E would have the same available noise power as the device (referred to the device's input) specified by T_E. For a passive device (such as a lossy transmission line or filter) that introduces no noise of its own, T_E is zero and G is a number less than one equal to the power loss of the device. The cascade of noise temperatures is similar to the formula for cascaded noise factors.

$$T_S = T_G + TE_1 + \frac{TE_1}{G_1} + \frac{TE_3}{G_1G_2} + \frac{TE_4}{G_1G_2G_3} + \ldots \tag{9}$$

where T_S is the system noise temperature (including the generator, which may be an antenna) and T_G is the noise temperature of the generator or the field of view of the antenna, usually assumed 290 °K for terrestrial communications.

The number 290 in the formulas for T_E is the standard ambient temperature (in kelvins) at which the noise factor of a two-port transducer is defined and measured, according to an IEEE recommendation. So those formulas relate a noise factor F, measured at 290 K, to the temperature T_E. In general, though, it is perfectly correct to say that the ratio $(S_I/N_I)/(S_O/N_O)$ can be thought of as the ratio of total system output noise to that system output noise attributed to the "generator" alone, regardless of the temperature of the equipment or the nature of the generator, which may be an antenna at some arbitrary temperature, for example. This ratio is, in fact, a special "system noise factor (or figure), F_S" that need not be tied to any particular temperature such as 290 K. (Note that regular noise factor (or figure) does depend on reference temperature.) The use of the F_S notation avoids any confusion. As the example of Fig 12.33 shows, the value of this system noise factor F_S is just the ratio of the total system temperature to the antenna temperature.

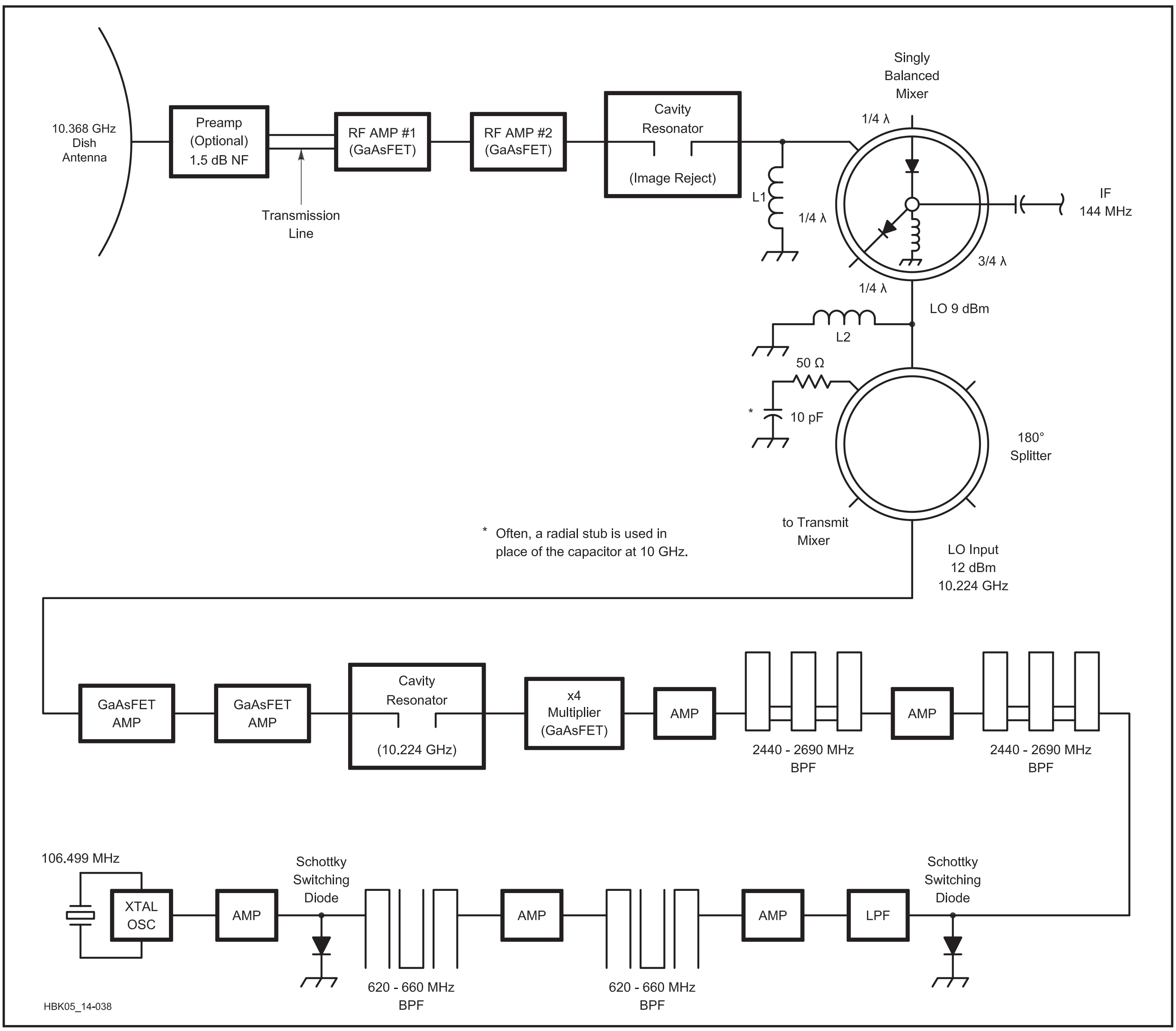

Fig 12.34 — A block diagram of the microwave receiver discussed in the text.

Having calculated a system noise temperature, the receive system noise floor (that is, the antenna input level of a signal that would exactly equal system noise, both observed at the receiver output) associated with that temperature is:

$$N = k\,T_S\,B_N \qquad (10)$$

where

$k = 1.38 \times 10^{-23}$ (Stefan-Boltzmann's constant) and

B_N= noise bandwidth

The system noise figure F_S is indicated in the example also. It is higher than the sum of the receiver and coax noise figures.

The example includes a loss of 3 dB in the receiver transmission line. The formula for T_S in the example shows that this loss has a double effect on the system noise temperature, once in the second term (288.6) and again in the third term (2.0). If the receiver (or high-gain preamp with a 6 dB NF) were mounted at the antenna, the receive-system noise temperature would be reduced to 1064.5 K and a system noise figure, F_S, of 7.26 dB, a very substantial improvement. Thus, it is the common practice to mount a preamp at the antenna, although transmission line loss must still be included in system noise figure calculations.

MICROWAVE RECEIVER FOR 10 GHZ

Here is a good example of amateur techniques for the 10 GHz band. The intended use for the radio is narrowband CW and SSB work, which requires extremely good frequency stability in the LO. Here, we will discuss the receiver circuit.

Block Diagram

Fig 12.34 is a block diagram of the receiver. Here are some important facets of the design.

1) The antenna should have sufficient gain. At 10 GHz, gains of 30 dBi are not difficult to get, as the example of Fig 12.33 demonstrates. A 4-foot dish might be difficult to aim, however.

2) For best results a very low-noise preamp at the antenna reduces loss of system sensitivity when antenna temperature is low. For example, if the antenna temperature at a quiet direction of the sky is 50 K and the receiver noise figure is 4 dB (due in part to transmission-line loss), the system temperature is 488 K for a system noise figure of 4.3 dB. If the receiver noise figure is reduced to 1.5 dB by adding a preamp at the antenna the system temperature is reduced to 170 K for a system noise figure of 2.0 dB, which is a very big improvement.

3) After two stages of RF amplification using GaAsFETs, a probe-coupled cavity resonator attenuates noise at the mixer's image frequency, which is 10.368 – 0.288 = 10.080 GHz. An image reduction of 15 to 20 dB is enough to prevent image frequency noise generated by the RF amplifiers from affecting the mixer's noise figure.

4) The single-balanced diode mixer uses a "rat-race" 180° hybrid. Each terminal of the ring is ¼ wavelength (90°) from its closest neighbors. So the anodes of the two diodes are 180° (½ wavelength) apart with respect to the LO port, but in-phase with respect to the RF port. The inductors (L1, L2) connected to ground present a low impedance at the IF frequency. The mixer microstrip circuit is carefully "tweaked" to improve system performance. Use the better mixer in the transmitter.

5) The crystal oscillator is a fifth-overtone Butler circuit that is capable of high stability. The crystal frequency error and drift are multiplied 96 times (10.224/0.1065), so for narrowband SSB or CW work it may be difficult to get on (and stay on) the "calling frequency" at 10.368 GHz. One acceptable (not perfect) solution might be to count the 106.5 MHz with a frequency counter whose internal clock is constantly compared with WWV. Adjust to 106.5 MHz as required. At times there may be a small Doppler shift on the WWV signal. It may be necessary to switch to a different WWV frequency, or WWV's signals may not be strong enough. Surplus frequency standards of high quality are sometimes available. Many operators just "tune" over the expected range of uncertainty.

6) The frequency multiplier chain has numerous band-pass filters to "purify" the harmonics by reducing various frequency components that might affect the signal path and cause spurious responses. The final filter is a tuned cavity resonator that reduces spurs from previous stages. Oscillator phase noise amplitude is multiplied by 96 also, so the oscillator must have very good short-term stability to prevent contamination of the desired signal.

7) A second hybrid splitter provides an LO output for the transmitter section of the radio. The 50-Ω resistor improves isolation between the two output ports. The original two-part *QST* article (see the references) is recommended reading for this very interesting project, which provides a fairly straightforward (but not extremely simple) way to get started on 10 GHz.

12.9 References and Bibliography

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