THE WORCESTER LONG DISTANCE M.W. RECEIVER

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In the design of a long distance medium wave receiver, the obvious problem of obtaining sensitivity sufficient to amplify the first circuit noise to room volume is so easy to achieve that it hardly ranks as a major design characteristic. Selectivity and strong-signal-handling capability, on the other hand, are overriding design factors of great importance and merit the utmost in study and creativity. The factor of selectivity in its application to the standard superheterodyne receiving circuit is a two-part problem. Selectivity ahead of the converter must be sufficient to reject the unwanted responses that are peculiar to the superheterodyne circuit and post-conversion selectivity is required to reject unwanted signals close to the desired one in the crowded medium wave environment. This latter problem will be considered first along with a break - through solution that provides selectivity in excess of that deemed possible previously by even the "ideal" selectivity characteristics of a vertically sided rectangle!

Substantially all of the close-in selectivity is obtained in the intermediate frequency amplifier. In conventional practice, the pass-band of this amplifier must be wide enough to pass the sideband information required for intelligent speech and to pass as well, an optimum signal-to-noise compromise on music. This means a 6 db bandwidth in the range of 4.5 to 5 khz. Obviously, then, from a selectivity standpoint there will be little attenuation of an undesired carrier one, two, or perhaps even three khz from the desired signal. This usual practice, however, fails to make use of the unique properties of the standard A.M. double sideband signal. The I.F. amplifying system in this receiver employs separate amplifyers for the upper and lower sidebands; each of which passes the carrier with 6 db attenuation. These amplifiers feed a novel detection system called DIFFERENTIAL SIDEBAND DETECTION (patent pending). Figure 1 diagrams the essentials of this selectivity breakthrough. It will be noted that two transistor detectors, Q1 and Q2, are used in a series output connection. The bias batteries shown indicate that just enough opening bias is applied to each base to start collector conduction and thus provide efficient detection. actually, batteries are not used but are shown to simplify the explanation. Transistor Q1 is driven by the USB amplifier and Q2 by the LSB. The detected audio signal appears across R1. First consider what happens for an on-tune 455 khz signal. In this instance, the signal from each amplifier drives the transistors equally, and the collector currents of the two transistors are driven in unison as they must be in a series connection. Obviously, the detection process is completely normal. The DC voltage at point X then stays constant at ½ the B-plus supply.

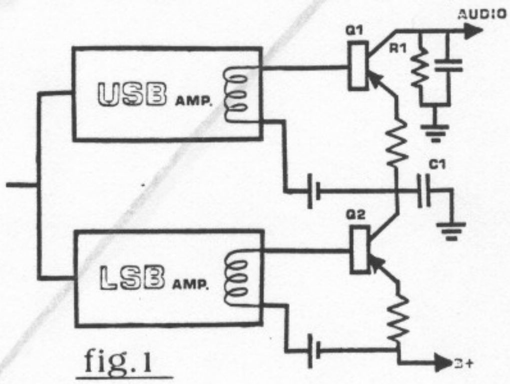


fig.1

Now consider what happens when the signal is at 456 khz. In this case the signal applied to Q1 is very strong and that applied to Q2 very weak. Q1 therefore, wants to draw a heavy current but is prevented from doing so by Q2 which is drawing only a weak collector current. The voltage across Q1 must then drop until its output current equals that of Q2. An inspection of a typical transistor collector family will show that this means the voltage across Q1 will be only two-tenths of a volt or so, indicating that this transistor has become an essential short-circuit and contributes nothing to the output. The audio output across R1 will now be determined by the weak signal from the LSB amplifier and will respond only to the amount of 456 khz signal that manages to permeate the LSB amplifier. Obviously for a 454 khz signal the process reverses. To obtain a clearer picture of what is happening, consider Figure 2.

Figure 2 is a stylized picture of two mechanical filters with shape factors of 2 bandwidths of 2.5 khz, and positioned so that the carrier is attenuated 6 db in each filter. The in-band ripple that is characteristic of these filters is not shown as it is not involved in the selectivity process. It was noted above that at 456 khz the response is that of the LSB filter showing an attenuation of approximately 50 db and at 454 khz the response will be that of the USB amplifier, also about 50 db. The 60 db bandwidth is now only 2.5 khz while the 6 db bandwidth is 5.0 khz! It will be noted that the effective selectivity curve becomes the overlap or differential of the two curves which accounts for the nomenclature "DIFFERENTIAL SIDEBAND DETECTION."

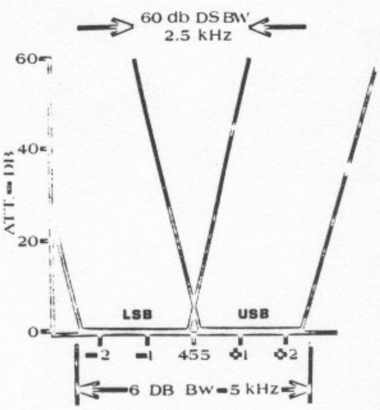


fig. 2

For a number of reasons mechanical filters are not used in the actual receiver. For one reason, very high carrier attenuations at 1 khz removed are not all that useful since the signal-to-interference ratio in the output will be determined largely by the high energy sidebands falling in the 455 khz response region. Also, the shape characteristic of a mechanical filter promotes the production of overshoots on impulse type noises such as static and the "buckshot" of sideband splatter. From a freedom from overshoot standpoint, the ideal response curve should look like a Gaussian error curve. Cascaded single tuned circuits approximate this shape closely but would require too many transistors. Double tuned circuits are a good compromise providing the coupling is maintained below the flat-flat region. Returning to the reasons for not using mechanical filters, it is felt that most DXers like to have a pleasant, realistic tone from their receivers and the substantial inband ripple that some of these filters have militates against good audio response due to the frequency distortion that results. Perhaps all of the above reasons are made academic by the realization that no suitable pairs of mechanical filters are currently available to the writer's best knowledge. For use in communication receivers, it is desirable to attenuate the carrier as much as possible, typically 25 db. In the circuit described, more than 6 db attenuation would cause amplitude distortion at high modulation percentages. To have special pairs designed for this purpose would be prohibitively expensive and time consuming at this juncture.

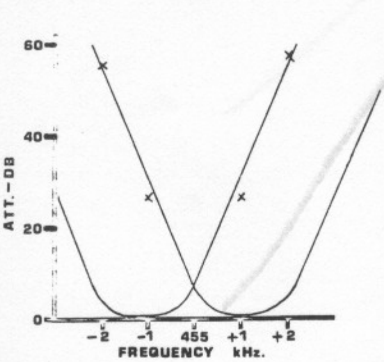


fig.3

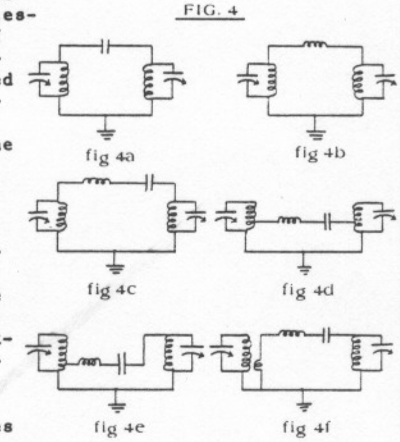
In the receiver described, each sideband amplifier comprises 16 transformers arranged in double tuned pairs plus a single special transformer to drive each detector. These filters have individual shape factors of 3 providing a 6 db bandwidth overall of 4.5 khz and selectivity as shown in Figure 3. The attenuation at 1 khz removed is approximately 30 db and at 2 khz removed about 60 db.

The preceding paragraphs have been addressed to the post-conversion selectivity problem and have described a novel breakthrough detection system to greatly facilitate the reception of foreign broadcasts on split frequencies. Selectivity prior to the converter is also very important to eliminate the various spurious responses that are unwelcome guests in the superheterodyne reception process. The effective selectivity provided is a function of the number of tuned circuits and the "Q" of same. This may sound like an easy problem to handle but the fact that the circuits must be tunable over a 3 to 1 frequency range and that the gain and bandwidth should remain constant over this range becomes a tough nut to crack when as many as three tuned R.F. circuits are employed. The use of inductive tuning would simplify these problems somewhat but such systems tend to become mechanical monstrosities and the realizable "Q's" are very low. Accordingly, this receiver uses a standard four section gang with three sections providing preselection and the other tuning the oscillator.

A tuned circuit in which the capacitor is the variable element has an impedance characteristic (gain) that rises linearly with increasing frequency and a bandwidth that also increases linearly as the frequency goes higher. The above assumes that the "Q" remains constant which is substantially the case. If three such circuits are cascaded, the gain would vary 27 to 1 over the band. Thus if the receiver were designed properly at the high frequency end of the band so that AVC control took hold with a 1 to 2 microvolt signal, the receiver would be severely sensitivity-limited at the low end of the band necessitating constant adjustment of the volume control as signals of varying strength are received. On the other hand, if the receiver were designed in the manner noted above at the low frequency end of the band, the noise between stations at the other end of the band would be intolerable and with the receiver going into severe AVC on noise alone, the signal-to-noise ratio would be affected adversely. Likewise, in the case of a 27 to 1 variation in bandwidth over the tuning range, if the "Q's" were made high enough to provide meaningful spurious signal rejection at the high frequency end of the band, such as would be provided by a 10 khz bandwidth, the bandwidth at the low end would be only .37 khz; not suitable at all for the reception of broadcast type signals. If the design situation were reversed with 10 khz bandwidth at the low end, at the high end the resulting 270 khz bandwidth would amount to virtually no rejection at all.

In view of the above, it is evident that if the exceptional spurious rejection possibilities of three tuned preselection circuits are to be realized, a design breakthrough is necessary to assure simultaneously both constant gain and constant bandwidth over the extent of the tuning range. There are other engineering reasons for doing this that are not of sufficient general interest to explain in detail. Figure 4a shows a conventional double tuned circuit with capacitive coupling. Since there is a 3 to 1 frequency coverage by the variable capacitors, their capacitive variation must be 9 to 1. Thus, with a fixed coupling capacitor the coupling coefficient is 9 times as high at the high frequency end than at the low; hardly the constant condition we are after. If the coupling capacitor could be replaced by a pure inductance as in Figure 4b, the coefficient of coupling becomes constant since the tuning inductances do not vary. This would produce constant gain if ordinary tuned circuits were involved but in order to achieve constant bandwidth it is going to be necessary in some manner to progressively reduce the "Q" of the tuned circuits as the frequency decreases. When this is accomplished, the coupling reactance must be made to vary 9 to 1 over the band rather than the 3 to 1, attained by the coupling inductance of Figure 4b, in order to maintain constant gain. This is accomplished by adding a coupling capacitor as shown in Figure 4c. If the resulting series tuned circuit is resonated at a suitable frequency below the lowest frequency covered, say 400 khz, the necessary 9 to 1 reactive coupling change can be realized to achieve constant gain. In actual practice, however, this circuit as diagrammed in Figure 4c is not realizable because the necessary inductance becomes so large that the unavoidable distributed capacitance of the coil causes it to resonate in the band covered. To eliminate this problem the coupling network can be tapped down on the tuned circuits as shown in Figure 4d. The proper tap selection will permit the use of a coupling inductance sufficiently low that its self resonance falls well above the highest frequency covered. If this proper tap should fall at N/3, to pick an example, the same result could be achieved with the same coupling values by tapping the first coil at N/9 and by returning the coupling network to the top of the second tuned circuit, as shown in Figure 4e. The reason for doing this is that it now permits us to simultaneously realize our twin objectives of constant gain and constant bandwidth as far as the second tuned circuit is concerned. Figure 4f shows such a circuit where the secondary of the first tuned circuit looks like approximately 1000 ohms which suitably loads the second tuned circuit at the low end of the band because the series coupling network is approaching resonance while at the high frequency end the reactance of the series tuned circuit is nine times higher which essentially decouples the 1000 ohm loading resistance from the second tuned circuit. It is hoped that this explanation will serve as a sufficient disclosure of the principle involved since the mathematical treatment, while uninvolved, is rather extensive.

FIG. 4



It is now appropriate to examine how this principle can be applied to the complete R.F. amplifier. To meet the sensitivity requirements of a long distance receiver, it is necessary to employ an R.F. amplifier stage to mask the relatively high noise content of the converter. Also, to obtain optimum signal-to-noise ratio, only one tuned circuit should precede the amplifier in order to minimize insertion loss, with the remaining two tuned circuits located between the R.F. amplifier output and the converter input in a reactively coupled arrangement. Figure 5 shows such a circuit. In the case of the tuned antenna transformer T1, the series coupling capacitor becomes the longwire capacitance-to-earth. The correct 220 pf value is provided by a 70 foot longwire (21m), approximately. A front-panel switch permits adding additional capacitance in series with the antenna to compensate for still longer longwires. Antennas shorter than 50 feet (15m) are not recommended. Low impedance inputs such as those from a loop preamplifier have the 220 pf capacitor added by a switching arrangement as shown.

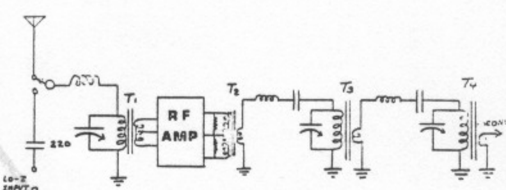
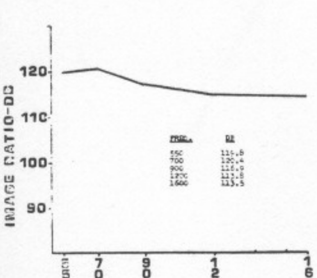


Fig. 5 RF Tuner
(Patent pending)

Transformer T2 in the output of the R.F. amplifier is broadbanded to cover the entire MW band without tuning. It couples the correct resistance through the coupling network to tuned transformer T3. T3, in turn, couples the proper resistive loading through the last series network to T4. Inspection of the measured image rejections gives a good idea of the superior rejection characteristics of this amplifier. These are shown in Figure 6, and reflect the excellent basic "Q's" of the tuned transformers T1, T3, and T4. These are enclosed in 18mm ferrite pot cores that provide realizable "Q's" in the circuit of 160 at 1600 khz, and this "Q" is reduced progressively by the circuitry described as the frequency is reduced, to approximately 55 at the low end of the band. The effectiveness of this circuit in providing constant bandwidth is demonstrated by the nearly constant image rejections over the extent of the band. Presently available receivers, even those in the $10,000 area, have image rejections that are not only considerably lower to start with but slope downward badly as the frequency is increased.

FIG. 6



Ferrite core material can be criticized from the standpoint of R.F. saturation on very strong signals. The result is that detuning results under heavy signal drive which produces some non-linearity. In order to minimize this effect, the standard 10mm construction with the "dumbell" bobbin for the winding was rejected since ferrite is present in the concentrated magnetic field. The larger pot core type, though expensive, was used instead because not only is the flux density reduced by the larger size but ferrite is not present in the highly concentrated central field except for the small tuning slug. An R.F. gain control is provided to avoid any problem of this nature that might occur with signals appreciably in excess of 110 db. Powdered iron cores were considered as a substitute, since they do not saturate, but the realizable "Q's" with this material are so low that ferrite material, all things considered, provides much the better compromise.

High image rejection, such as that indicated in Figure 6, is a good indication, also, of superior rejection of other spurious responses closer to the desired signal. With the receiver tuned to 1000 khz, no further spurious responses were observed at any frequency with an undesired signal input of 126 db (2 volts).

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Perhaps at this point a look at the receiver's sensitivity is appropiate. The antenna coil circuit, shown in Figure 5, is matched to the longwire's impedance which is in the order of 1000 ohms. Most receivers, on the other hand, match to an input impedance of 50 ohms, which just happens to be the impedance of most signal generators. This permits a direct connection between signal generator and receiver, thus avoiding the loss introduced by the intervening dummy antenna which must otherwise be used. In addition, matching to the input of the first active device by the antenna transformer results in a greater voltage step-up when you start from 50 ohms than can be realized when the starting impedance is 1000 ohms. All this means that the sensitivity measurements for publicity purposes are much better in the 50 ohm case but the actual performance with a real live longwire is very bad because of the large mismatch. In view of the above, two sets of sensitivity measurements are provided in Figure 7; one showing the sensitivity when measured through a standard IEEE dummy and the other made by connecting the signal generator through a 50 ohm matching transformer to the receiver. Curve B is applicable when comparing sensitivity measurements with other receivers having 50 ohm inputs.

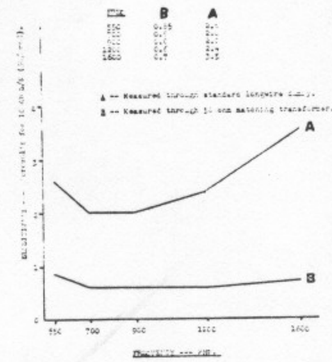

FIG. 7

Having considered the dual selectivity problems in some detail, we can now turn our attention to the equally important problem of strong signal handling capability. Referring back to Figure 5, it will be noted that the R.F. amplifier is indicated only as a "black box." What to put in this black box is the purpose of the next few paragraphs. Firstly, as regards the selection of the active device, the choice is between tubes, FET's, and bipolar transistors. The first two are simple majority-carrier devices depending on electron flow for their operation; while the latter is much more sophisticated, relying on the production of minority carriers or "holes." The effectiveness of an active device as an amplifier is indicated by its mutual conductance (Gm) which is merely the change in output current that results from a known change in input voltage. This can be labeled in micromhos or more directly as milliamperes-per-volt. A fairly accurate "rule of thumb" is that majority-carrier devices have mutual conductances of about one ma/V for every milliampere of output current. Thus an FET operating with a grain current of 5 ma would have a mutual conductance of 5 ma/v. Minority carrier devices, however, have Gm's of 40 ma/v for every milliampere of collector current and are therefore much more "vigorous" active devices. A bipolar transistor, accordingly, with a collector current of one milliampere would have a Gm of 40 ma/v. The other major distinction that must be made is in regard to the relative input impedances of the two device families. Tubes and FET's have very high input impedances, of the order of megohms at MW frequencies; but as the frequency increases further, various transit time and feedback effects gradually reduce this impedance until at 100 Mhz it is of the order of only a few thousand ohms. A bipolar transistor, on the other hand, has a low impedance even at MW frequencies. As a matter of fact it is equal to the current gain divided by the mutual conductance. Thus for a beta of 40, the input impedance becomes 40/40 ma/v or 1000 ohms. At 100 Mhz, the input impedance would still be 1000 ohms providing a transistor type was selected whose beta remained at 40 at this higher frequency.

The preceding paragraphs are preparatory to considering the widely held notion that FET's have superior strong-signal handling properties to bipolar transistors, even at MW. If a strong signal is applied to the input of an active device, basic overload occurs when the negative going excursion of the signal is sufficient to cut off the flow of output CURRENT or when the positive drive is enough to reduce the output VOLTAGE to zero because the drop across the load impedance equals the supply voltage, or a combination of both. Thus, in the case of an FET having a 5 ma drain current and a Gm of 5 ma/v, a negative going peak signal of 1 volt would cutoff the drain current. A bipolar transistor, however, with a collector current of 1 ma and a Gm of 40 ma/v would experience collector current cutoff with a peak negative signal of only 1/40 volt. Obviously, therefore, at say 100 Mhz, where the input impedances are essentially equal, and the devices driven from the same tap on the input transformer, the FET would have a strong-signal handling superiority of nearly 40. On MW, however, the FET has to be able to handle a much stronger signal to merely break even with a bipolar device because the high input impedance FET must be driven from the top of the tank while the bipolar is tapped way down at a much lower voltage level. To develop some figures, a tuning capacitance of 50 pf and a coil "Q" of 160 would provide an impedance at 1600 khz of QX, where X equals 1/6.28 fC or 1989 ohms. Z then equals 1989 x 160 or 318, 240 ohms, which becomes the impedance that drives the FET, To drive the bipolar at optimum signal-to-noise, however, requires an undermatched driving impedance of $\sqrt{\text{beta}}$ x Re. If beta is 50 and Re 26 ohms, this impedance becomes 184 ohms. The impedance ratio then becomes 318,240/184 or 1730. The turns ratio or voltage ratio is then the square root of 1730 or 41.6. The net result of all this is that an FET can handle 40 times as much signal as a bipolar but has to handle 41.6 times as much to break even due to the different voltage drive levels , so at MW it is pretty much of a standoff as far as signal handling capability is concerned.

The preceding paragraph becomes academic if the R.F. amplifier takes the form of a pair of Class AB push-pull transistors. The current cutoff condition in the output then never occurs because one transistor conducts while the other is cutoff and the only drive limitation takes place when the output voltage is reduced to zero by positive drive peaks. This drive limitation can be thwarted by increasing the voltage supply and by decreasing the load impedance. In this competition the FET is hopelessly outclassed. The much higher mutual conductance of the bipolar, 40 ma/v vs. 5 ma/v, permits the use of 1/8th the load impedance to obtain a given amplification and the much greater variety of available types permits the use of high voltage transistors suitable for MW use. The R.F. transistors used in this receiver, Siemens BF178, were basically developed for use as television horizontal sweeps. They have a maximum collector voltage rating of 160 volts which allows them to operate safely from a 75 volt supply. Their cutoff frequency is 120 Mhz which permits them to operate satisfactorily in the MW band if beta selection is employed to reject those above 75. Class AB operation, noted above, is really Class B in the audio sense as just enough quiesent current (2 or 3 ma) is allowed to flow to prevent crossover nonlinearities under strong signal conditions. The transistor's maximum collector current rating is 50 ma, but to draw anything approaching this in straight Class A operation would degrade the signal-to-noise ratio because of the high shot noise produced in the output.

Using the push-pull connection as noted above and with 47 ohm emitter resistors, the R.F. amplifier can handle a 2 volt, 100% modulated antenna signal before clipping occurs. Compare this with the signal handling capabilities of a tube amplifier. With a Gm of 5 ma/v, and operating with 5 ma plate current, a 1 volt half-peak signal would clip, which means .707 volts peak rms or .354 volts carrier, assuming 100% modulation. Since the tube is a high impedance device the antenna transformer would have a gain of typically 5, making the antenna signal .354/5 or only 71 millivolts! This signal handling limitation can be increased to 0.5 volt by the application of AVC to the R.F. amplifier which makes it possible to listen to a signal this strong but is of no help to the DXer who is interested in a weak signal on an adjacent or split frequency where the AVC has essentially disappeared due to I.F. selectivity, and the R.F. attenuation in minimal.

Moving on to the converter circuit, three of the high voltage transistors mentioned above are used in a balanced push-pull arrangement. This balanced operation cancels intermodulation products to the extent that none could be found at 20 khz removed at maximum signal generator settings (2 volts from one and 0.1 volt from the other). The common-mode I.F. rejection is also outstanding as noted on Figure 8.

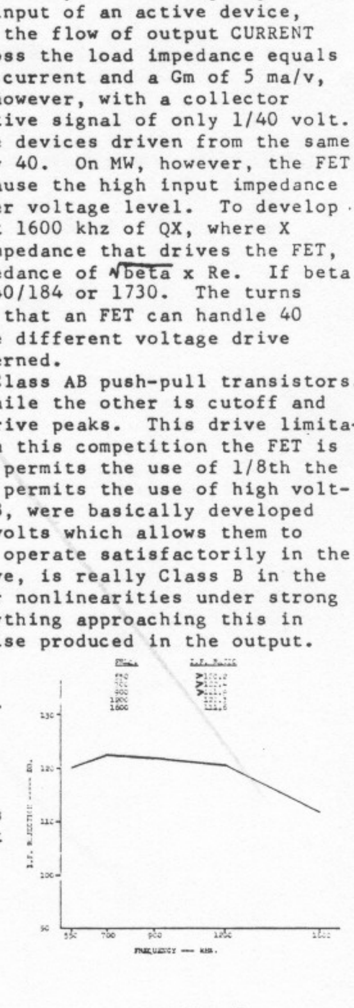

FIG. 8

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M3

A simplified schematic of the circuit is shown in Figure 9. Not shown are the various trimming, padding, and temperature compensating capacitors. In addition to the latter, oscillator drift is minimized by operating at unusually high power; 75 supply volts at 20 ma total drain current, and by zener regulation of the bias applied to the bottom transistor. Temperature drift at the high frequency end of the band is minimized by making a portion of the trimming capacitance max-negative and at the low frequency end by doing the same to a portion of the padding capacitance. Both of these compensations combine in the middle of the band to produce a drift characteristic that is somewhat over compensated as shown in Figure 10. This drift is entirely unnoticeable in practice as the receiver can be tuned to a station from a cold start and no observable improvement in tuning occurs after a period of time. This satisfactory performance of a free-running oscillator makes unnecessary the use of complicated and expensive synthesis or digital AFC schemes that introduce problems of their own.

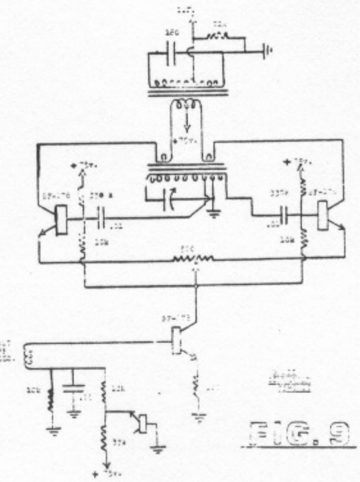


FIG. 9

Between the I.F. output shown in Figure 9 and the dual sideband I.F. amplifiers is a 455 khz amplifier module, the primary purpose of which is to provide AVC control. This is shown in Figure 11 and is probably the most sophisticated AVC control system available today. The 9 volt supply shown at the extreme left, supplies forward bias to each of the 5 silicon diodes across each of the 5 tuned circuits. The voltage applied to the diode across the last transformer is that of the two series connected silicon diodes at the extreme right, about 1.1 volts. The voltage applied to the other control diodes increases progressively due to the added voltage drop across each of the 10 ohm resistors. These forward biases applied to the anodes are bucked by the positive AVC voltage applied to the cathodes, as shown. With no-signal this voltage is approximately 2 volts and with the application of signal this voltage is made to drop by the amplified DC of the detector output until with 0.1 volt applied signal (100 db), the AVC voltage has dropped to about 1.1 volts. Thus, as the signal increases the decreasing AVC voltage permits the loading diode on the first transformer to conduct first, since it has the highest positive anode voltage, and the other diodes conduct sequentially as the signal strength continues to increase. This results in an output signal that remains essentially constant over the entire dynamic range of the receiver. This characteristic is shown in Figure 12. The attainment of AVC control by the process of "Q" reduction as just described is much superior to amplifier bias change from the dual standpoints of dynamic range and freedom from cross-modulation.

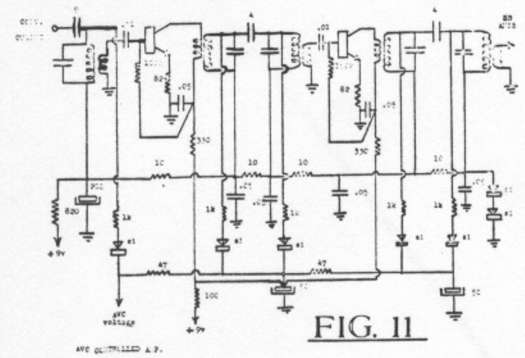


FIG. 11

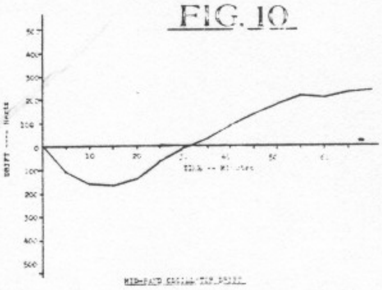


FIG. 10

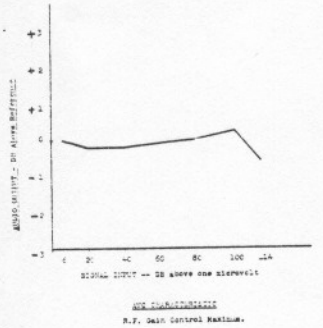


FIG. 12

A few remarks regarding the digital readout may be of interest. This operates by counting the oscillator frequency but displaying the signal frequency. This is accomplished by using presettable counters (74192) so that the count starts at 9545. After counting 455 (the I.F. frequency) the count reaches 10,000; but since only four digits are displayed the 1 does not appear. After completing the oscillator count, the displayed count is now the frequency of the signal (in khz). A crystal controlled clock circuit is used ti initiate and stop the counting procedure. Since the individual cycles are counted but the display is in kilohertz it would seem that a counting period of 1 millisecond is in order. This, however, presents a problem. There is no correlation between the time when an oscillator cycle begins and when the clock circuit initiates the count. Accordingly, the last digit fluctuates plus-or-minus one count, which of course, not only makes an accurate frequency reading impossible but is visually annoying, as well. To combat this, one more decade is counted (.1 khz) but not displayed. This requires the counting period to be 10 times as long or 10 ms. To further insure that the last displayed digit does not fluctuate, the crystal frequency is made a hertz or two high so that the count at resonance can fluctuate between 2 and 3, for instance, rather than between 0 and 9 which would produce a count reaction down the line. If the counting period is made 10 milliseconds, the off period is made this long, also, so that the display can be updated and the counter resets accomplished. The complete clock cycle thus totals 20 milliseconds or the clock frequency is 50 hertz.

In this receiver, the basic crystal-controlled oscillator frequency is 16 khz. A divide-by-16 counter, 7493, reduces this to 1 khz. This is followed by a 7490 divide-by-ten counter, connected BCD, so that the various conditioning pulses can be obtained from the remaining outputs. This is followed by ½ of a 7474 flip-flop to provide the final clock signal of a 50 hz square wave. The clock chain can be followed on the top line of the block diagram shown in Figure 13. Below this on the left is a discrete amplifier-limiter. The oscillator pick-off signal to be counted is applied to the input. The back-to-back diodes square this signal off so that there is no need for a Schmitt trigger I.C. The two amplifying transistors, being unidirectional devices, serve to isolate the steep front of the clock signal from affecting the ocsillator's output. The gain provided by this amplifier permits the use of a low amplitude oscillator signal which permits additional decoupling of the counter pulses from the oscillator to the point where no observable problem has been noted. The output of the amplifier connects to the input of a NAND gate (½ 7420) to which also is connected the clock pulse. The output of this gate will now feed oscillator pulses to the counter during the period when the clock signal is positive and turn them off during the negative portion of the clock cycle. The counter comprises one 7490 and four 74192's. The former does not drive a display as discussed previously. Each 74192 drives a 7475 latch which in turn drives a 7447 decoder-driver and this drives an LED readout. The purpose of the latch is to immunize the display during the counting period and to update the displayed digit during the quiescent period by means of the ENABLE pulse. This pulse is derived by the 7420 section shown on the extreme right. These are quad-input NAND gates, all inputs of which have to be high (plus) in order to produce an output pulse. It will be noted that one input is driven by the inverse of the gating signal clocking the oscillator pulses. This is to insure that the ENABLE pulse is produced only during the quiescent period and not while the count is proceeding. Also driving the inputs of this NAND gate are the Q1, Q2, and Q4 outputs of the 7490 in the clock chain. However, the Q4 output has its polarity reversed by an inverter (1/6 7404) and thus an output pulse is produced when Q1 and Q2 are high, Q4 low, and of course, the clock count in its quiescent state.

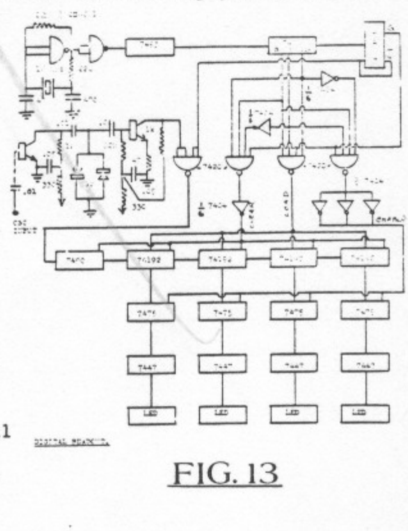


FIG. 13

This means that the ENABLE output pulse is produced during the 3 count of the 7490 (Q1 plus Q2 equals 3). Since a NAND inverts the polarity, it is necessary to further invert the output pulse in order to obtain the required HIGH ENABLE. Three parallel connected inverters (½ 7404) are necessary in order to drive four 7475's.

Now that the ENABLE pulse has updated the displayed count, it is next necessary to apply a CLEAR pulse to the 74192's to reset them to zero. This is done on the 5 count by inverting Q2 (Q1 plus Q4 equals 5). The necessary CLEAR pulse is also HIGH so that an interter is required in the output of this gate, also. This must be followed by a negative LOAD pulse in order to preset the counters to 9545. This is produced on the 7 count by leaving all the Q inputs high (Q1 plus Q2 plus Q4 equals 7).

With so many pulses flying around, it is necessary to use some care in the layout and construction of the counter to eliminate the possibility of interference. As noted previously a two stage discrete amplifier is used to prevent direct conductive interference with the oscillator. Headaches are also avoided by not attempting to use gaseous-discharge nixie tubes as displays. As will be noted by Figure 14, the various I.C.'s are Elmered belly-up to a printed circuit board. Strips of printed-circuit board material are laid over the rows of I.C.'s to provide the B-plus bus lines. In order to "sink" the high current output spikes, a liberal sprinkling of .01 bypass capacitors is necessary (one for each I.C.). The board is then shielded as can be seen from the chassis photograph, Figure 15.

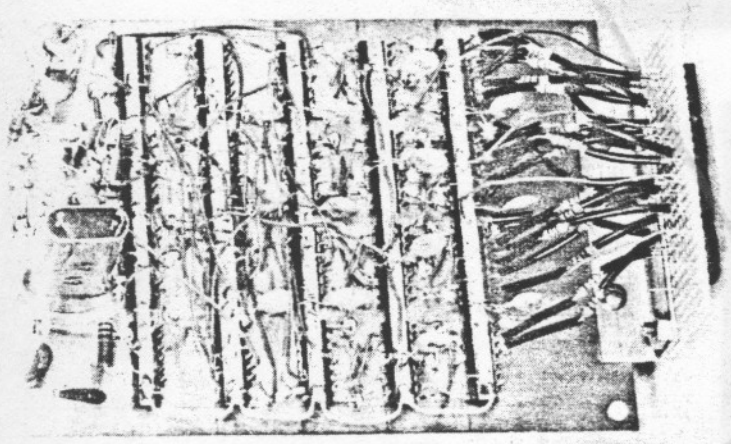


FIG. 14

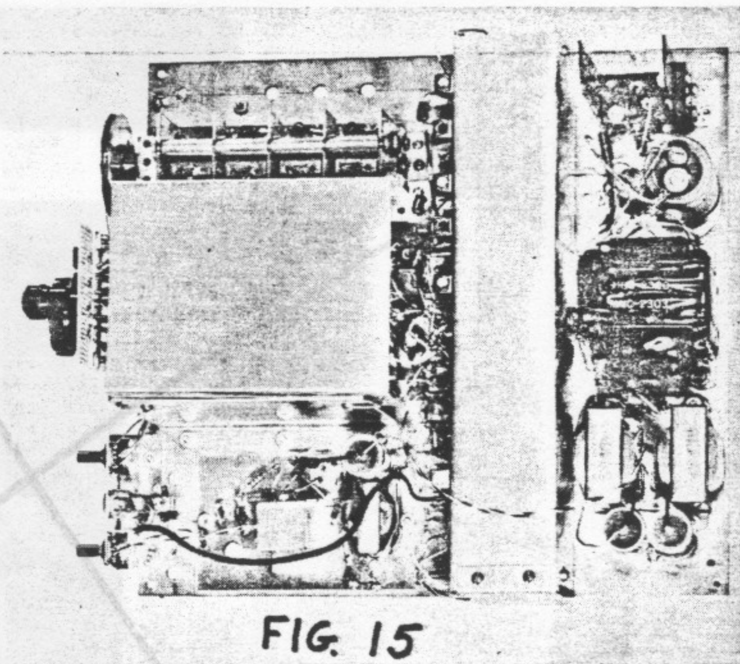
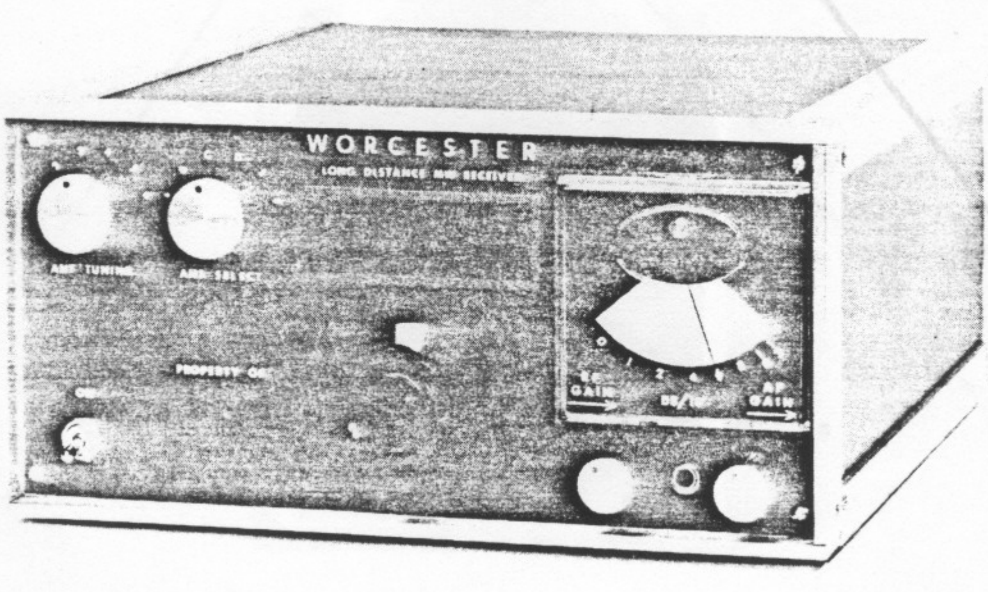


FIG. 15

This concluding paragraph will cover a few loose ends that still remain. The audio amplifier is conventional and uses a high voltage output transistor similar to RCA 40321. The power output is 1 watt at 10% distortion and 1.35 watts maximum output. The overall audio frequency response of the receiver is 150 to 2000 hertz at the 6 db points. The power transformer has dual secondaries; one to provide 100 volts DC to supply the output transistor and the high voltage R.F. and converter transistors. The other winding provides a 9 volt supply for the I.F. amplifier and auxillary circuits and a regulated 5 volt supply for the digital readout. The power consumption is 26 watts.



M3-6-6 M3

The cabinet material is aluminum and the various panels are masked at the edges during the painting process so that a reasonably good electrical shield is provided. The cabinet dimensions are approximately 11¼" length, 5¼" height, and 12" depth (28x13x30cm).

Looking at the front panel, the control at the extreme upper left is the ANT TUNING switch which compensates for variations in longwire characteristics. It will be remembered from the report, that the longwire antenna is part of the Antenna Transformer circuit and this control, which has 6 positions adds additional series capacitance to provide the required 220 pf. The switch position should be selected that provides the highest reading on the tuning meter for a weak station at the low frequency end of the band for the longwire in use.

Next to the ANT TUNING switch is the ANT SELECT switch. This is also a 6 position switch which permits the selection of any desired loop, longwire or beverage attached to the 4 antenna posts at the rear of the cabinet. These are selected as follows:

- A - Loop L1 (or beverage)
- B - Loop L2 (or beverage)
- C - Longwire A1
- D - Longwire A2
- E - Longwires A1 plus A2
- F - Disconnects antennas and grounds receiver input.

This switch should be left in position F when the receiver is not in operation to prevent possible damage from high voltage spikes from lightning and other sources.

At the center of the front panel is the digital readout display and below it the tuning control. As described in the report it reads to the nearest kilohertz. The large meter on the right indicates signal strength and is calibrated approximately in decibels above one microvolt divided by 10. Below the tuning meter on the left is the R.F. Gain control which is maintained in the extreme clockwise position unless distortion is encountered on very strong signals. For best reception of weak signals it is essential that this control be in the extreme clockwise position. Below the meter on the right is the A.F. Gain control. This is the conventional volume control and should be adjusted to the volume level desired. Between these two gain controls is the jack to accomodate headphones. The jack normally provided parallels stereo headphones, but a jack can be supplied for monaural phones, if desired. In this connection, an external loudspeaker (8 ohm) is required (and can be supplied) and this plugs into an RCA type jack on the rear panel. Another RCA jack permits the operation of a tape recorder.

It will be noted that no heterodyne elimination control is provided and the elimination of this is made possible by the DIFFERENTIAL SIDEBAND DETECTION system which automatically provides a 30 db attenuation of a carrier 1 khz removed and nearly 60 db attenuation of one 2 khz away. Regardless of whether a Het Elim notch filter takes the form of a Twin-Tee, Wien Bridge, or potentiometer-tuned network, the realizable attenuation is a theoretical 40 db if all components are precise and held to a 1% tolerance. Practically speaking, it will be less than 40 db and an unwelcome guest is an undesirable frequency distortion of the audio signal.

To the left of the main tuning control are the words PROPERTY OF. Below this, the name of the owner can be inscribed if he wishes. This personalization may serve as an effective deterent to theft. However, if it is not wanted, this area will be left blank.

The writer would like to thank Bob Mahrenholz for valuable help in adapting to the world of digital integrated circuits and also to list the following credits:

WIRELESS WORLD Dec. 1971 - C. ATTENBOROUGH - "Displaying Frequency Digitally."
WIRELESS WORLD Nov. 1974 - G. Lomas - "Signal-Frequency Meter."
LANCASTER, Donald E., - TTL Cookbook

NOTICE

Free license to use the novel features of this receiver is given to DXers who wish to construct for their personal use. If profit motives are involved, write for licensing details.
It has been the writers experience that very few DXers are interested in doing their own construction which applies particularly to something with this degree of complication. In view of this, 6 receivers have been built, aligned and tested. Each has had at least 10 hours burn-in time on stations including TA's. PRICE - $795.00 each plus transportation.

J.A. Worcester
R.D. 1
Frankfort, NY 13340

C	F	C	F	C	F	C	F
39=	102	19=	66	-1=	30	-21=	-6
38=	100	18=	64	-2=	28	-22=	-8
37=	99	17=	63	-3=	27	-23=	-9
36=	97	16=	61	-4=	25	-24=	-11
35=	95	15=	59	-5=	23	-25=	-13
34=	93	14=	57	-6=	21	-26=	-15
33=	91	13=	55	-7=	19	-27=	-17
32=	90	12=	54	-8=	18	-28=	-18
31=	88	11=	52	-9=	16	-29=	-20
30=	86	10=	50	-10=	14	-30=	-22
29=	84	9=	48	-11=	12	-31=	-24
28=	82	8=	46	-12=	10	-32=	-26
27=	81	7=	45	-13=	9	-33=	-27
26=	79	6=	43	-14=	7	-34=	-29
25=	77	5=	41	-15=	5	-35=	-31
24=	75	4=	39	-16=	3	-36=	-33
23=	73	3=	37	-17=	1	-37=	-35
22=	72	2=	36	-18=	0	-38=	-36
21=	70	1=	34	-19=	-2	-39=	-38
20=	68	0=	32	-20=	-4	-40=	-40