XIII. Conclusions

The use of lumped-element, low-pass prototype filters in the design of the band-pass filters for parametric amplifiers and up-converters is found to simplify the problem greatly. This is because once the lumped-element, low-pass prototype is specified, only the fractional bandwidths and impedance levels of the band-pass filters remain to be specified. Reducing the problem to this number of degrees of freedom is seen to make the design problem quite tractable. In the case of up-converters, ordinary lumped-element Tchebycheff or maximally-flat prototypes are found to be quite appropriate, but in the case of parametric amplifiers with the negative-resistance type of operation, such prototypes should be modified as indicated in Sections VI and VII in order to prevent high peaks of gain at the band edges. The two-reactive-element prototype given in the last paragraph of Section VII should be suitable for many cases where two-resonator filters are desired.

As can be seen from (37), (49) to (58), and (71b), the fractional bandwidths called for in the design of the band-pass filters of parametric amplifiers and up-converters are directly proportional to $a=C_1/C_0$. Further, it will be noted that in all of the equations in this paper the degradation of performance due to diode loss is always controlled by the parameter $(aQ_d)=(C_1/C_0)Q_d$ rather than by Q_d alone. [In the case of up-converters see (74).] Thus the size of C_1/C_0 is of vital importance in determining the bandwidth capabilities of a parametric amplifier or up-converter. At present little information is available as to the largest practical values that C_1/C_0 can take without overdriving the diode so as to degrade the noise figure. The value $a=C_1/C_0=0.25$ used in the examples of this paper is probably practical, though perhaps on the optimistic side.

In the case of degenerate parametric amplifiers the fractional bandwidth of the band-pass filter is *inversely* proportional to the resonator slope parameter x_1 (or b_1 for the diode resonated in shunt). For nondegenerate amplifiers and up-converters the fractional bandwidth w is inversely proportional to $\sqrt{x_1x_1'}$ (or $\sqrt{b_1b_1'}$ for the shunt resonance case). Thus it is of great importance to design the diode resonator circuit so as to minimize the reactance (or susceptance) slope at f_0 and f_0'. Stub tuners and the like must be avoided since they introduce unnecessary resonances and, as a result, large reactance slopes. If the diode resonator is a series resonator, it is desirable that it be followed immediately by a shunt resonator (or vice versa if the diode resonator is a shunt resonator). If the diode resonator and the next resonator of the filter are separated by a connecting line, part of the selectivity of the connecting line must be charged to the diode resonator, thus increasing the effective size of the diode resonator slope parameter and reducing the bandwidth which can be used in the filter.

In summary, the bandwidths which can be achieved in engineering practice will be influenced very much by the usable C_1/C_0 ratio of available diodes, and by the degree to which the slope parameters of the diode resonators can be minimized using practical circuitry. Though the examples presented herein involve a guess as to what C_1/C_0 should be and are idealized in some respects, they should be strongly suggestive of the performance that can be expected.

Correction

G. R. Valenzuela, author of "Impedances of an Elliptic Waveguide (For the ${}_eH_1$ Mode)," which appeared on pp. 431–435 of the July, 1960, issue of these Transactions, has brought the following to the attention of the *Editor*.

On page 433, read:

into Hz, and then integrating, it can be easily shown that

$$V = i\omega\mu \int_0^{u_0} \int_{-\pi/2}^{\pi/2} H_z(u, v) ds_1 ds_2 = 2\int_0^{u_0} E_u\left(u, \frac{\pi}{2}\right) ds_1.$$

Also on page 433, pertaining to (4), read:

Now the impedances can be easily obtained. Let

$$Z_0 = 120\pi\left(\frac{\lambda g}{\lambda}\right) \quad \text{and} \quad h = q(k^2 - \beta^2)^{1/2},$$

then

$$Z_{WV} = \frac{h^4 I_1^2}{I_2} Z_0$$

and

$$Z_{WI} = \frac{I_2}{4J_{e1}^2(u_0)} Z_0. \tag{4}$$

On page 435, for (8) read:

The expression for the approximate impedances are

$$Z_{WV} = \frac{32}{3\pi} \frac{r(3 + r^2)}{(5 + 2r^2)} Z_0$$

$$Z_{WI} = \frac{3\pi}{32} \frac{r(15 + 11r^2 + 2r^4)}{(2 + r^2)^2} Z_0. \tag{8}$$